ELECTRONIC ENGINEERING

VOL. 37 No. 452

OCTOBER 1965

Commentary

THE Independent Television Authority has recently celebrated its tenth anniversary. The official celebration took the form of a dinner at the Guildhall, London, on 16 September, which was attended by many of the leading personalities connected with the Authority and by the Prime Minister.

The Television Act 1954, which provided for the setting up of the ITA, received Royal Assent on 30 July 1954. A few days later its first Chairman, Sir Kenneth Clark, K.C.B., was appointed and this was followed on 1 October by the appointment of Sir Robert Fraser, O.B.E., as its first Director-General.

The Postmaster General granted a licence to the ITA on 6 April 1955 and on 22 September of that year its first transmitting station at Beulah Hill, Croydon, came into operation on a vision frequency of 194·75Mc/s; the vision transmitter had a power of 10kW, the aerial mast was 175ft in height and the effective radiated power was 60kW.

The inauguration of this service broke new ground both technically and sociologically. Its allocation to Band III meant the employment of frequencies some three or four times higher than any that had previously been used for a public service in this country and this in turn meant the development and employment of new techniques both in transmitters and receivers: it has, over the years, also made necessary the development of many new studio and production methods to suit the particular requirements of commercial television. Sociologically it raised quite a storm. It was the first time commercial radio or television had been introduced to this country (with Governmental approval, that is!) and it attracted to itself many bitter and vociferous opponents who held that the impact of commercialism on a nationwide public service would prove to have a highly corruptive influence on the community. As Lord Hill of Luton, the present Chairman of the ITA said in his speech at the Guildhall, "It now seems that the nation's capacity for self-corruption is very small". There are still, no doubt, many people who would not agree fully with Lord Hill and it may be many years yet before it is possible to make a sound judgement on the impact of television, commercial or otherwise, or to decide how beneficially the technical advances in this field have been utilized. Be that as it may, however, the fact is that over the last ten years, commercial television has become an accepted part of life in this country.

From its beginning in 1955 the ITA organization grew rapidly. In 1956 a further three transmitting stations were opened and in the following year the first station in Scotland was brought into operation and this was followed, in 1958, by stations in South Wales and the Isle of Wight. In the succeeding years a number of other stations have been built to increase or consolidate the national coverage and it is of interest to note that when the Mendlesham station in East Anglia was opened in October 1959 its 1 000ft aerial mast was then the tallest in Europe.

Financially commercial television has certainly been a success and the present level of profits from this source causes considerable consternation in some quarters. It should, however, be remembered that it was not always so; to enter this new field the programme companies needed not only foresight but courage. The capital requirements were high and in the early days vast sums of money were being poured out and little was coming in; indeed, in the first year or so of operation Independent Television was tottering on the brink of bankruptcy. That it is now established on so firm a footing is a tribute to all involved.

Artistically, too, Independent Television has enjoyed a fair measure of success and over the years a great number of awards has been won at national and international events.

With a venture such as Independent Television it is obviously impossible to please all the people all the time, or indeed, some of the people for any of the time, but taken all round, Independent Television may justly claim to be a successful and smoothly working organization and many of its active participants would welcome new grounds to conquer. One way to provide this would, of course, be to allocate a second competitive channel for commercial television. This would also please another section of the industry for it is clear, from a memorandum recently sent to the Television Advisory Committee, that the British Radio Equipment Manufacturers' Association would welcome such a move.

In this memorandum BREMA restates the industry's views that the decisions taken by the Government in 1962 (Cmnd. 1770) should be implemented. This White Paper, which was published following the recommendations of the Pilkington Committee, stated, in effect that all programmes, whether on v.h.f. or u.h.f. should be on 625 line standards, that a commercial competitive programme should be introduced as early as possible and that colour should only be introduced on 625 lines. BREMA also states that, although the industry's choice of a colour system would have been NTSC, it would not, in the interests of international standardization, oppose the adoption of the PAL system; the main plea is that a decision should be made, and made now. With the publication of the Pilkington Report and the Government White Paper (Cmnd. 1770) in 1962 the way ahead seemed clear but, due to international wrangling and the internal economic position the ultimate goal is now very much obscured.

Domestic television is an important sector of industry and it should not be allowed to stagnate. To prevent this firm decisions must be taken on the questions of colour system, line standards, allocation of programmes etc, and they must be taken now. It is imperative too, that these decisions should not only provide for satisfactory internal development but that they should place the manufacturers in the best possible position to compete in export markets.

An Automatic Co-ordinate Measuring Equipment Using Moiré Fringes

By R. L. Farrell* and G. E. Moore*

Precise information on the flight paths of rockets, etc., is obtained by measurement of the positions of their images relative to those of the stars on photographic plates. With the aid of the 'Zeiss' comparator, an optical co-ordinate measuring apparatus, readings on these plates can be taken to an accuracy of better than 1 micron.

The equipment described records the movements of the plate carrier in the comparator and provides visual decimal display as well as punched card output.

The measurements of the two co-ordinates are obtained to an accuracy of 1 micron by counting Moiré fringes. The fringe patterns are read out by photo-voltaic cells and amplified. Special circuits determine the direction of movement. The decimal counters used to accumulate the fringe count are capable of operation at high speed and rapid reversal and contain some novel design principles. A special warning system based on the concept of critical deceleration detects errors due to failure of optical or electronic equipment.

(Voir page 704 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 711)

At the Weapons Research Establishment, ballistic cameras are one of the instrumentation systems used to record the flight paths of rockets, etc. Precise trajectory information is obtained from photographic plates by measuring the positions of the objects relative to the positions of stars, which also appear on the plates. In addition, camera calibration data are obtained periodically by determining the positions of a large number of known stars. The corrections so obtained are applied to the trajectory measurements by means of a digital computer. The precise measurements on the photographs are obtained with the aid of a 'Zeiss' co-ordinate measuring instrument, or comparator, which is capable of measuring the co-ordinates of a point on a plate to an accuracy of better than 1 micron (μ).

The principles employed in the comparator in order to achieve this high accuracy involve the use of X and Y scales which are arranged to be lying in a straight line with the distance to be measured, and which are divided so as to allow direct comparison with it. This minimizes errors due to tilting of the plate and due to slack in levers, screws, etc.

The plate is secured to a carrier which may be moved manually along either horizontal axis and the point whose co-ordinates are to be determined is brought under the cross-wires of a microscope. The scales slide on straight edges attached to the carrier. In former practice, they were read off through two additional microscopes. Either the object point or the scales could be viewed through the same pair of eyepieces, by switching the illumination. Archimedes' spiral were used for interpolation on the scales. The readings were spoken into a tape recorder and were later transferred manually to punched cards for use with the computer.

It is the purpose of this article to describe an automatic measuring and read-out facility for the instrument, based on electronic counting of 'Moiré' fringes[1,2]. The readings, which are displayed directly in decimal notation, may be automatically transferred on to standard punched cards.

Production of Fringes

The term 'moiré' pattern is derived from the French for 'watered' silk, in which a decorative pattern resembling ripples on water is produced by doubling the glossy corded fabric over and pressing it together. In general, moiré patterns are produced when figures of a periodic structure are made to overlap. The moiré fringes used in distance measurement are produced by the superposition of two high quality optical gratings of equal pitch. A long grating is coupled to the moving object to be measured, while a shorter, or 'index' grating is at rest with respect to the observing instrument. The rulings on the two gratings form a small angle. Interference between the two gratings produces a pattern of light and dark strips running approximately along the direction of the length of the gratings (Fig. 1). The pattern will travel across the gratings if the latter move relative to one another in a lengthwise direction. A given point on the stationary grating will go through a complete cycle of maximum and minimum brightness for each ruling-width of grating movement.

The very fine gratings used in the application under consideration are uniformly transparent, their rulings consisting of grooves of triangular cross-section. With gratings of this nature, the generation of the moiré effect is dependent on their diffraction properties, good patterns being produced only if the incident light is approximately coherent and the light output is nearly monochromatic. This is achieved by arranging the optical system to form a monochromator[3]. The straight, thin filament of the tungsten illumination lamp placed at the focus of a collimating lens provides a source of coherent light, which is dispersed by the gratings and then focused by a composite lens on the photocells, which convert the received

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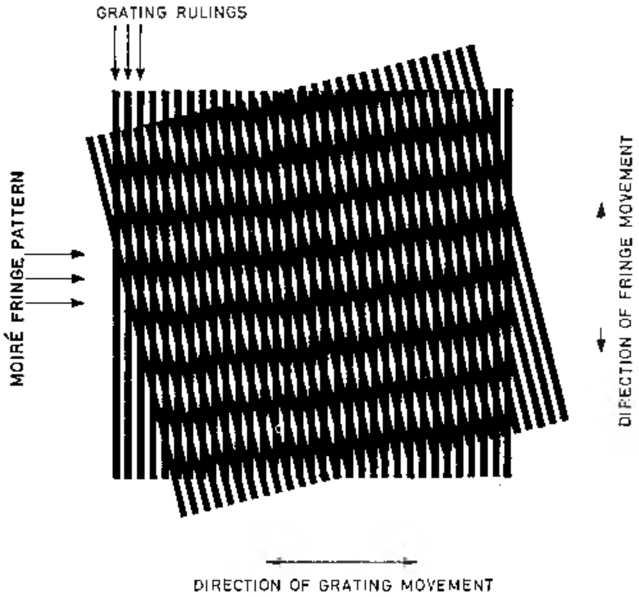


Fig. 1. Illustration of the moiré fringe principle

light intensity into electrical signals. The variation of light intensity along the fringe pattern is approximately sinusoidal. The angle between the rulings of the two gratings is adjusted to give one complete cycle of brightness variation across the width of the gratings. The width of the index grating is divided into four equal rectangles. By means of the composite focusing lens, the light from each rectangle is made to converge on a separate sensing cell. The outputs of interleaved pairs of cells are connected in anti-phase (Fig. 2). This arrangement suppresses the current due to minimum illumination, and provides information on relative movement between the gratings in terms of the number of cycles of intensity variation, and on the direction of travel in terms of the phase sequence of two sine waves in quadrature.

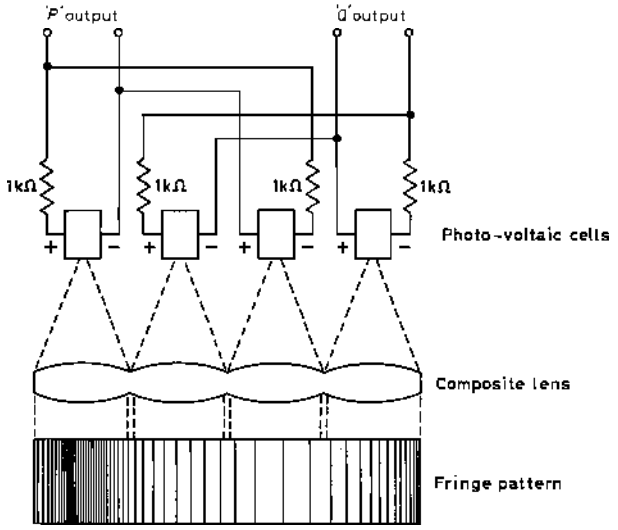
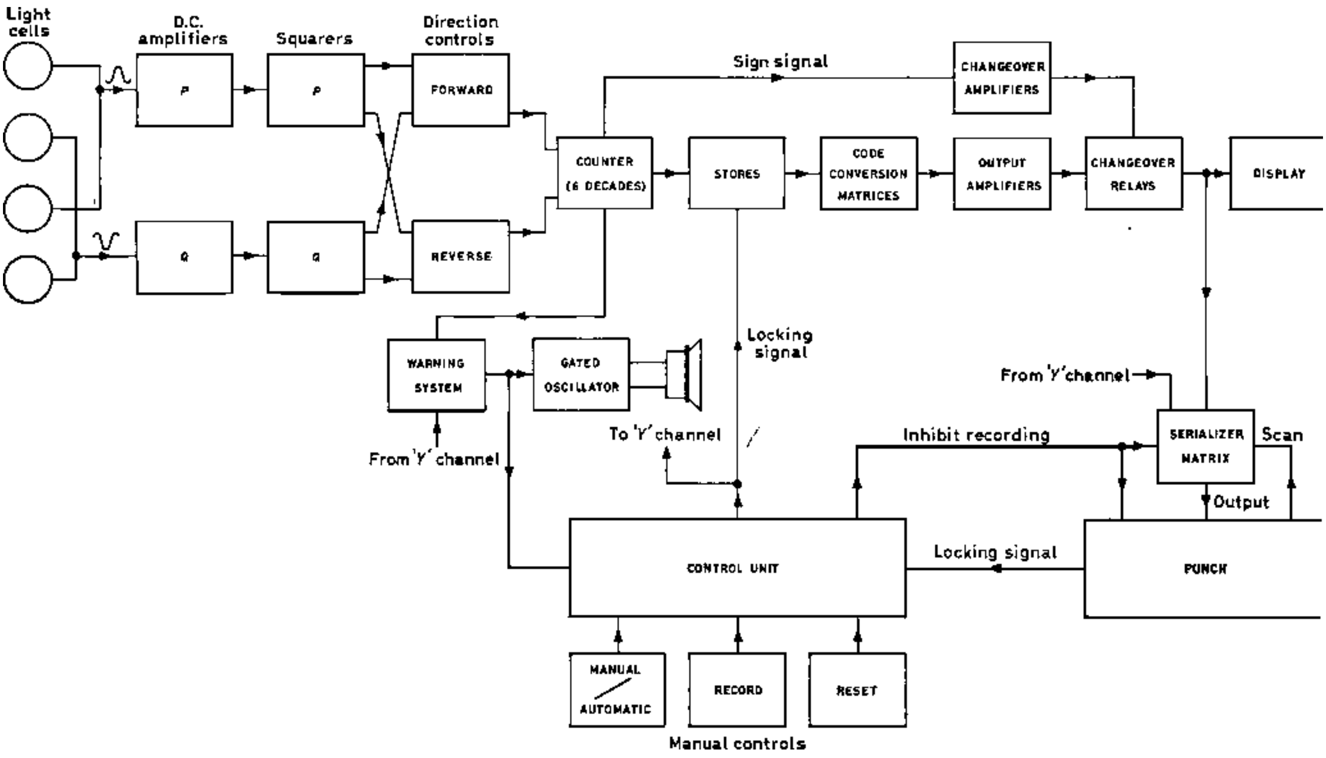


Fig. 2. Photocell arrangement

Read-out

Suitable gratings and measuring heads have been fitted alongside the X and Y scales of the plotting table, permitting the principle described above to be employed. The gratings used have a pitch of 4μ (i.e., 250 lines/mm). Circuits to be described in the following sections generate one pulse at each zero cross-over of either of the two quadrature signals, hence one pulse appears for each micron of travel. X and Y distances are measured by accumulating these pulses in reversible counters, having regard to the direction of motion. The counters have a

Fig. 3. Arrangement of equipment



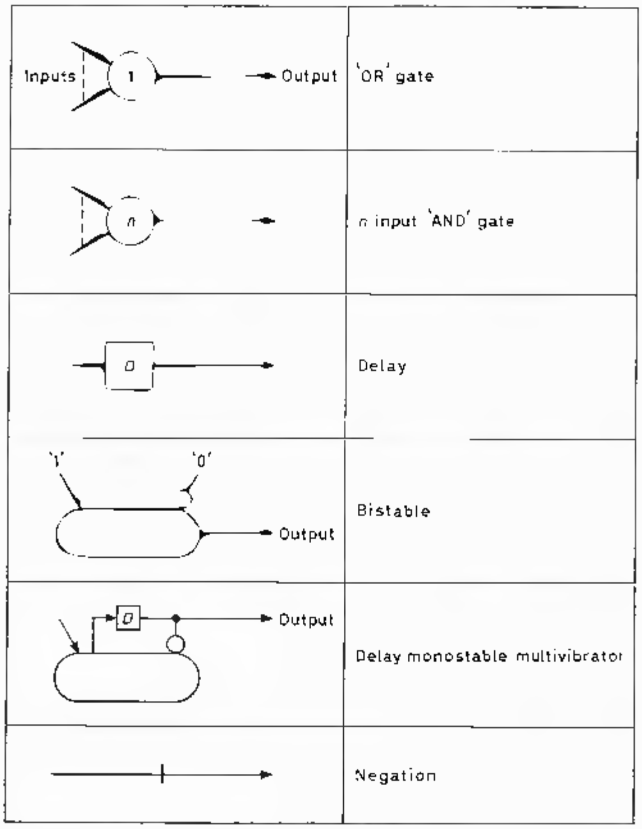


Fig. 4. Logic symbols

capacity of six decades each and are capable of resolving pulse rates of over 10^6p/s, so that movement along either scale at up to 1m/sec is permissible. The maximum acceleration is 1*g*, higher values being considered harmful to the comparator.

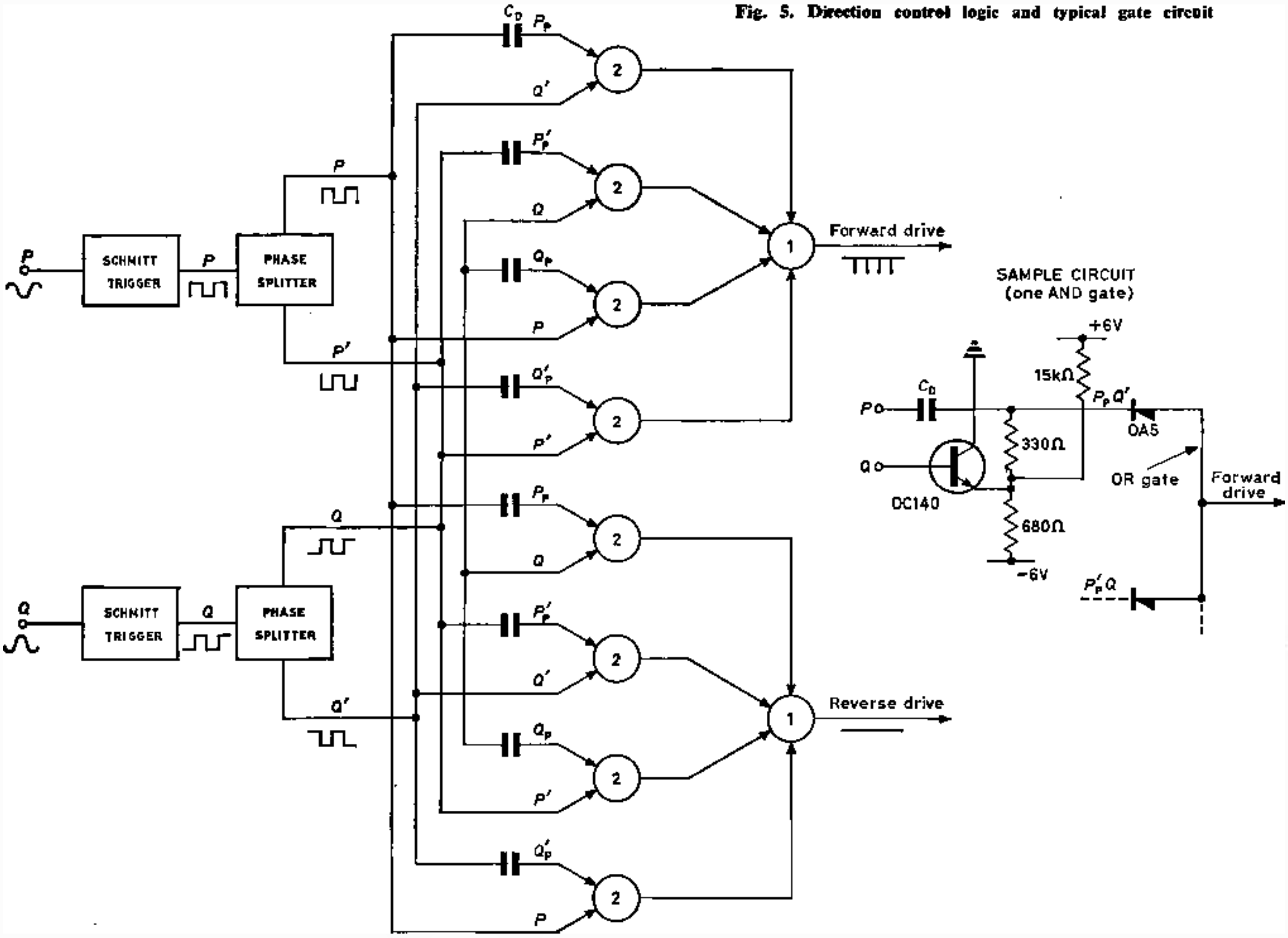
The overall arrangement of the electronic system is shown in Fig. 3, which has been condensed, since identical channels are used for the *X* and *Y* scales. Each channel consists of input signal shapers, a reversible counter and store, and a visual display. In addition, there is a card punch with its associated serializing equipment, a control unit, and a 'lost count' warning system common to both channels.

A standard 'Hollerith' programme board punch, type 25, has been modified to operate directly from the electrical signals provided, in addition to performing its normal copying and manual punching operations.

Design

The choice of light-sensitive cells was the subject of considerable experimentation. Finally, small photo-voltaic cells ('Silicon Read-out Photocells' by International Rectifier Corporation), normally used in punched tape read-out applications, were chosen. These silicon cells combine exceptionally good temperature stability with high signal-to-noise ratio. They show no ageing effects. Their non-linear characteristic is of no consequence in this application, since only a threshold level in the electrical output

Fig. 5. Direction control logic and typical gate circuit



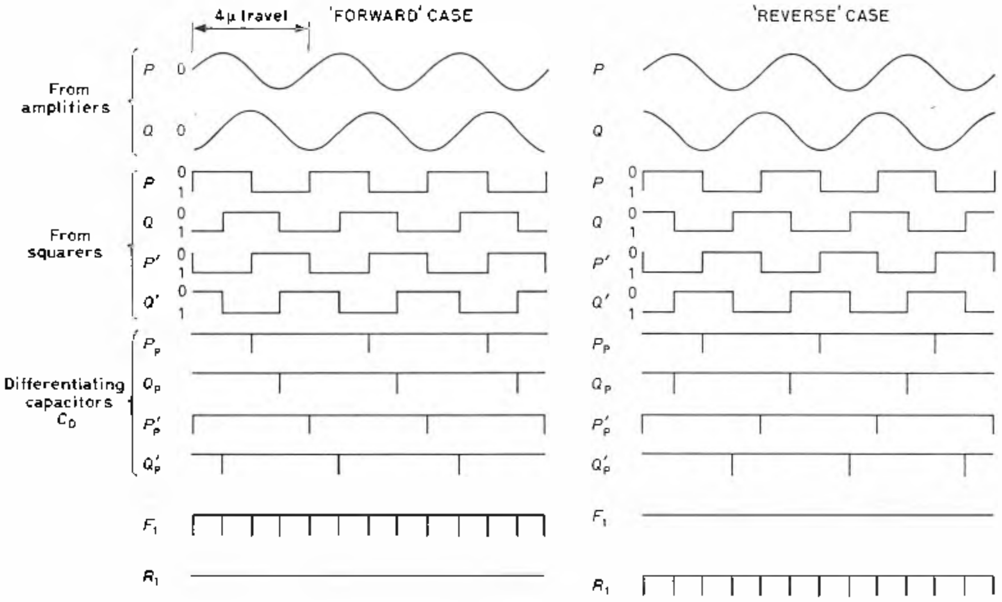


Fig. 6. Direction control, waveforms

The signals P and Q and their inverse, P' and Q' are obtained in the required form by phase-splitting circuits.

The reversible counters described below have separate input terminals for forward and reverse counting. Pulses derived from the P, Q, P' and Q' waveforms are directed to the correct point by the direction control circuits shown schematically in Fig. 5, in terms of the logical symbols defined in Fig. 4.

The pulses P_p, Q_p, P_p' and Q_p' are obtained by differentiating and selecting the negative-going transitions of the squared signals (Fig. 6). Note that within the drive gate circuit, a 0 is represented by earth potential, a 1 by a negative voltage.

The 'forward' pulse is formed by

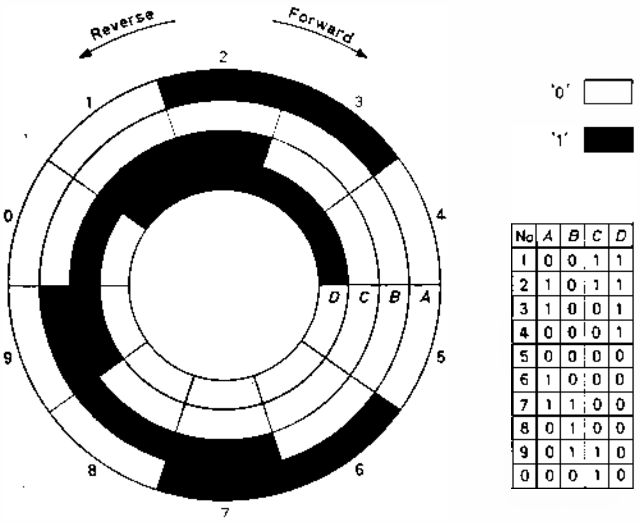
$$F_1 = P_pQ' \vee P_p'Q \vee Q_pP \vee Q_p'P'$$

the 'reverse' pulse by

$$R_1 = P_pQ \vee P_p'Q' \vee Q_pP' \vee Q_p'P$$

The counters are capable of operating at input rates of more than 1Mc/s, and of reversing instantaneously.

The operation of the individual decades is based on



No	A	B	C	D
1	0	0	1	1
2	1	0	1	1
3	1	0	0	1
4	0	0	0	1
5	0	0	0	0
6	1	0	0	0
7	1	1	0	0
8	0	1	0	0
9	0	1	1	0
0	0	0	1	0

Fig. 7. Decade code

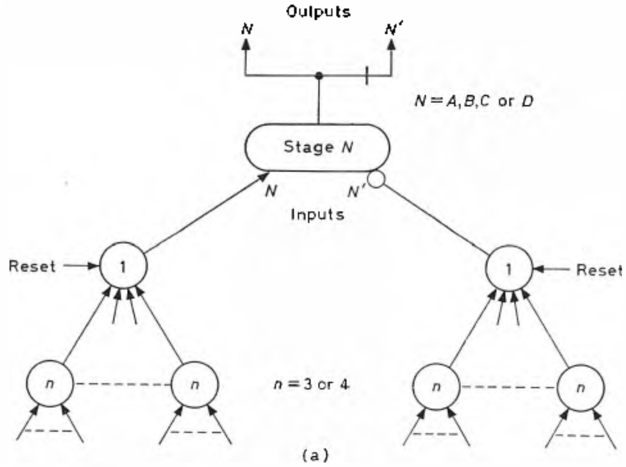
need be defined, and their high-frequency response was found to be adequate.

Since these photo-voltaic cells show best temperature characteristics at low load impedances, the two cells forming a pair are connected as shown in Fig. 2. The output current is the difference between the currents generated in the two cells. This is of the order of a few micro-amperes and considerable signal amplification is necessary. The signals derived from the two pairs are designated P and Q.

The d.c. amplifiers used to bring the photocell signals up to a usable level are based on an existing design[4]. The input was applied to the low impedance side of a balanced pair. Very good stability was obtained by mounting the transistors in a common heat sink.

The amplified signals are converted to 'up or down' functions by a 'Schmitt' trigger circuit whose aperture is adjusted to 100mV to provide optimum discrimination between the minimum signal amplitude (0·5V) and the combined noise and drift voltages which have never been observed to exceed 20mV.

Fig. 8. Decade logic
(a) Logic arrangements, single binary stage
(b) Gating functions for one decade



STAGE A
A INPUT =
BCDF V BCDF V BCDR V
BCDR V RESET
A INPUT =
BCDF V BCDF V BCDR V
CDR V RESET

STAGE B
B INPUT =
ADF V CDF RESET
B INPUT =
CDF V ADR V RESET

STAGE C
C INPUT =
ABF V ADR V RESET

C INPUT =
ADF V ABR V RESET

STAGE D
D INPUT =
ABCF V ABCR V RESET
D INPUT =
ABCF V ABCR V RESET

(b)

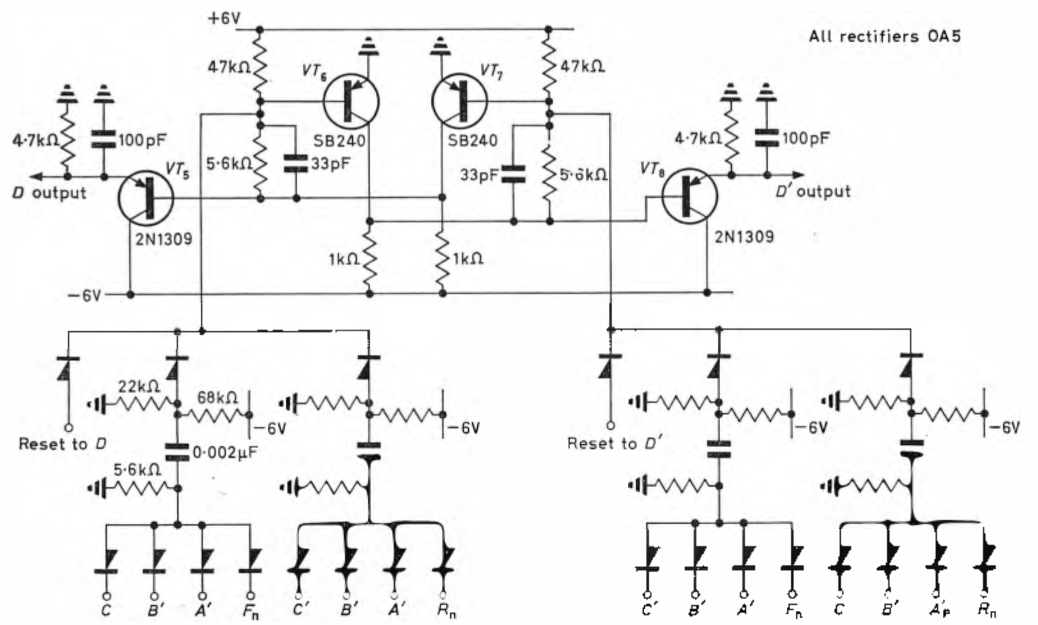


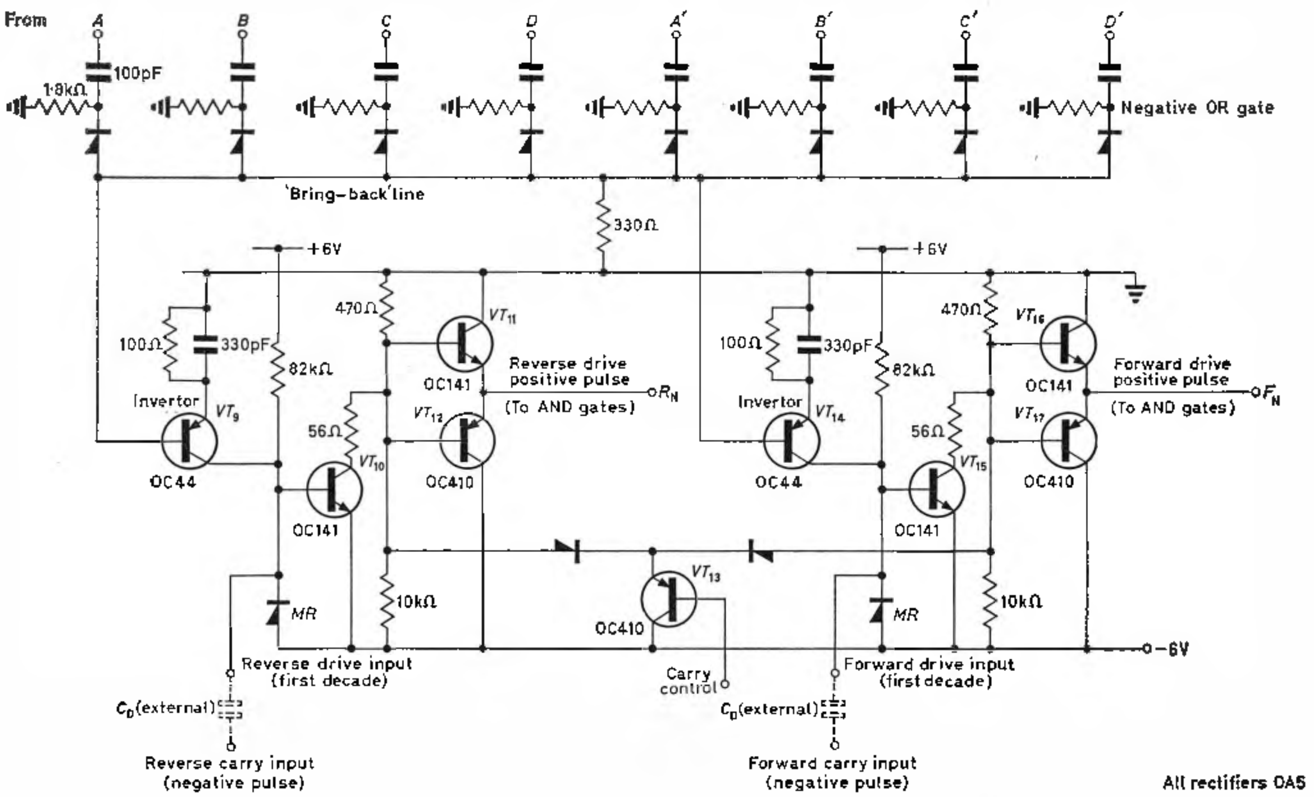
Fig. 9. Circuit of a typical bistable stage—(Stage D)

the cyclic unit distance code shown in Fig. 7. Of the four bistables A, B, C and D, only one changes state in any counting step. This avoids the time delay caused by ripple-through in conventional counters. Since no bistable changes its state more than once in two counts, the requirements of speed for the individual bistables are not severe.

The logic of the decades may be explained with the aid of Fig. 8. The diagram, Fig. 8(a), is a generalized representation of one of the four bistables. One AND gate is provided for each set of conditions in which the binary is to change its state. The inputs to the AND gates are the forward or reverse input pulse to the decade and the states of the three bistables (or only of two of these, where this is sufficient to define the transition). The gating functions are listed in Boolean notation in Fig. 8(b). The letters F and R represent the forward and reverse input pulses respectively. As an example, the A' input will be operated (i.e., stage A will be switched from '1' to '0') by the forward pulse when $B = 1$, $C = 0$, and $D = 0$, or when $B = 0$, $C = 0$, and $D = 1$, or by the reverse pulse when $B = 0$, $C = 0$, and $D = 0$, or when $D = 1$ and $C = 1$. The 'reset' inputs on all OR gates have been provided to achieve full flexibility in the choice of the resetting points. They are connected as required. The circuit of one bistable (stage D) with its associated gates is shown in Fig. 9. The convention used in the bistables is that '0' is represented by −6V, '1' by earth potential.

The correct functioning of the decades depends on accurate control of the width of the drive pulses, since the counters can 'free-run' through several steps in the

Fig. 10. Pulse forming circuits



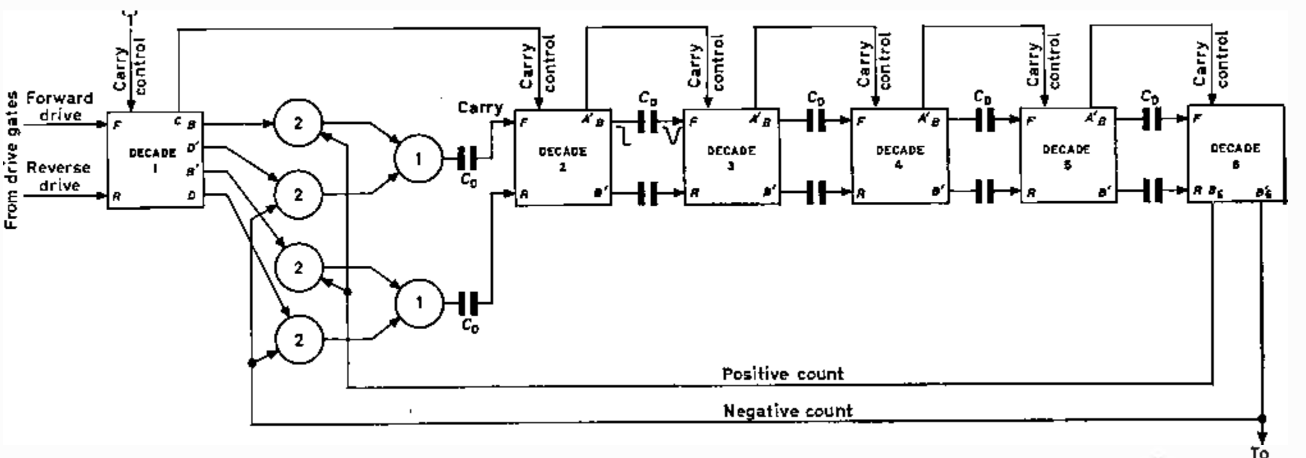


Fig. 11. Counter interconnexions

presence of an extended input pulse. To ensure reliable operation under all conditions, the pulse width is automatically adjusted to a value just sufficient to complete one counting step. This is accomplished by the pulse forming circuits shown in Fig. 10. The input signal is a negative step coupled to the forward or reverse input via one of the capacitors C_D. (In the case of the least significant decade, four such capacitors are used in each input, forming part of the direction control circuits.) The signal is amplitude-limited by the diode *MR*, inverted and amplified by VT_{10} (VT_{15} in the case of the reverse drive signal) and applied to the AND gates of the counter via push-pull emitter-followers. As explained, one of the bistables will change its state, generating a negative and positive step, respectively, on it two output terminals. The negative step is differentiated and applied via the negative OR gate seen at the top of Fig. 10, to the "bring-back' line. It is inverted by VT_9 and VT_{14} and used to restore the potential on the output side of capacitor C_D in either the forward or the reverse input, sharply terminating the drive pulse. The positive step generated by the change of state in the bistable is delayed somewhat by the 100pF capacitor connected across the ouput emitter-follower (Fig. 9). This allows for the time required for the 'bring-back' cycle to be completed, before another AND gate is made ready for the next counting step.

A schematic representation of a whole six-stage counter is shown in Fig. 11. The first (least significant) decade is driven directly by the drive gates. The third and following decades are each driven by transitions of the *B* stage of the preceding decade. Since the carry pulse is required to occur between states '9' and '0', the transition of *B* to 0 is used for carry in 'forward' the transition of *B'* to 0 in 'reverse' (Fig. 10). The unwanted change in the *B* stage, occurring between the states '6' and '7' (Fig. 7), is suppressed by the transistor VT_{13} (Fig. 10), which, in the decades mentioned, is operated by the *A'* output of the preceding stages. The carry from the first to the second stage takes place in the normal way in positive counting, but must take place between states '0' and '1' while the count is negative. This is explained with reference to Table 1. The carry transitions in this case are derived from stage *D*, the 'carry control' function is exercised by stage *C*.

As mentioned earlier, the count representing a certain grating position may be recorded by a card punch. Since this device requires several seconds to operate, it is necessary to store the count present at the start of the recording cycle and to separate the store circuits from the counters to allow free movement of the gratings during recording.

The store consists of a separate set of bistables. Through diode gates, each storing stage is d.c.-coupled to the output lines of one bistable pair in a counter. While the gates are open, i.e., while the 'lock' line is at +6V, each storing stage will be forced to take up the same state as the corresponding counter pair. During the recording cycle, the locking line is held at earth potential, the gates are closed and the store stage will remain in the state existing just prior to receipt of the recording command.

The four store stages for each decimal place are connected to a conventional diode matrix which converts the four-digit binary code to the '1 out of 10' code required for display and punch outputs.

The output of *X* and *Y* stores is presented to the punch via a resistance matrix consisting of 12 columns for the 6*X* and 6*Y* digit positions and 10 rows for the digits 0 to 9. The columns are scanned successively by a sliding contact in the punch. Each row (via transistor amplifiers) operates a relay-operated microswitch providing a signal to the punch.

TABLE 1

Display Connexions

STAGE 6 (MOST SIGNIFICANT)		STAGES 2–5		STAGE 1 (LEAST SIGNIFICANT)		
STORE CONTENT	DISPLAY	STORE CONTENT	DISPLAY	STORE CONTENT	DISPLAY	
		9	9	9	9	POSITIVE
		8	8	8	8	
		7	7	7	7	
6	6	6	6	6	6	
5	5	5	5	5	5	
4	4	4	4	4	4	
3	3	3	3	3	3	
2	2	2	2	2	2	
1	1	1	1	1	1	
0	0	0	0	0	0	
9	0	9	0	9	1	NEGATIVE
8	1	8	1	8	2	
7	2	7	2	7	3	
		6	3	6	4	
		5	4	5	5	
		4	5	4	6	
		3	6	3	7	
		2	7	2	8	
		1	8	1	9	
				— CARRY —		
		0	9	0	0	
		— CARRY —				

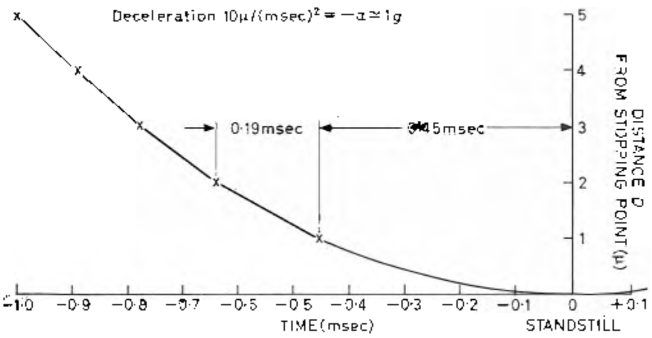


Fig. 12. Critical stopping characteristic of the comparator table

The control system provides for two modes of operation:

The 'manual' mode is intended for casual measurements, when visual readings from the Digilite display suffices. Negative numbers can be displayed in this mode. Since no provision for negative quantities exists in the card format, the recording facility is inhibited. When the reset button is pressed to establish a reference point, both counters are reset to 000000.

The 'automatic' mode is used when the data are required to be punched on cards. After selection of this mode, it is necessary to press the reset button in order to release the inhibition on recording. The counters are now reset to the values $X = 500\,000$, $Y = 200\,000$. Because the maximum traverse in either X or Y direction is about 19cm, the highest and lowest counts are approximately:

for X, 690 000 and 310 000 respectively

for Y, 390 000 and 010 000 respectively;

i.e., negative numbers do not occur in this mode.

Correct display of negative numbers in the 'manual' mode is produced as follows:

When either co-ordinate enters the negative region, the counter proceeds from 000000 to 999999, 999998, etc. As the movement in both the X and Y directions is limited to about 19cm, a negative count may be recognized by the existence of a count of 9 or 8 in the most significant decade. A signal from stage B in the latter may be used to determine the sign of the count (ref. Fig. 7). This signal is amplified and used to operate a bank of relays which alter the output connexions so that the store content is displayed as the complement to 1000000, which constitutes the required negative number. This complement is obtained by complementing the least significant decimal digit to ten, and all others to nine. The connexions for positive and negative counting are listed in Table 1. The relays also operate lamps which illuminate the vertical bars in the sign displays, changing them from '−' to '+', when appropriate.

The carry from one decade to the next takes place when the less significant digit, as displayed, changes between 9 and 0. In the least significant decade, which is complemented to ten in the negative region, this corresponds to a change between 0 and 1 in the counter (Table 1). The carry conditions must therefore be altered in this case. This is done by the gates interposed between decades 1 and 2, shown in Fig. 11, and controlled by the B waveform of the most significant decade.

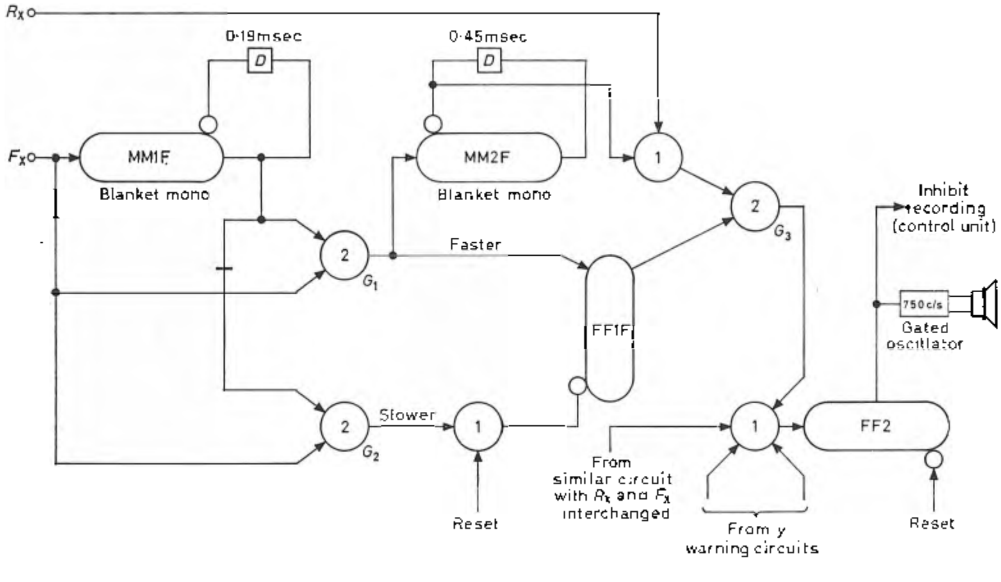
Warning System

If the train of drive pulses is interrupted while the plate carrier is in motion, an error occurs and the reference point setting will be lost. Such a disruption could be caused by maladjustment or component failure in the optical system, the light sensing cells, the d.c. amplifiers, or the squarers. Failure of either P or Q input signals to the drive gates would cause rapid alternation between reverse and forward pulses. Abrupt stopping of the plate motion may harm the equipment and is therefore also treated as a fault condition. Each of these conditions will be detected by a system of circuits called the 'warning system' which sounds a warning tone and inhibits recording until the counters have been reset manually.

The warning system will operate only at plate speeds in the X or Y direction greater than 5mm/sec. This limitation is not considered serious, since the input circuits have a falling gain-frequency characteristic, so that failure of a channel is less likely at slow plate speeds. In the following, only the case of 'forward' motion in the X direction will be considered, since similar circuits are used for reverse motion, with the roles of 'forward' and 'reverse' pulse inputs interchanged. The circuits used in the Y system are identical to those in the X system. The warning circuits are supplied with forward and reverse pulses obtained from the pulse-forming circuits of the least significant decades of the X and Y counters. The output is common to all four circuits. It consists of a bistable controlling a 750c/s gated oscillator with a telephone earpiece acting as a sound transducer.

The curve in Fig. 12 shows the distance of the moving grating from its resting position as a function of time before standstill, assuming that the grating is decelerating at the maximum permitted rate of $10\mu/\text{msec}^2$ (= $1g$). The graph shows that the last micron traversed takes 0·45msec, the second last, 0·19msec. It is uncertain where on the

Fig. 13. Warning logic



curve the μ pulses will occur, but if a given pulse follows the preceding one after less than 0·19msec, then the grating is moving at greater than critical speed (5·4mm/sec) and may not stop (or reverse) before having registered another 'μ' pulse spaced not less than 0·19msec, but not more than 0·45msec after the given pulse. The warning logic shown in Fig. 13 is based on this principle. The operation of the circuit will be described for the case of normal deceleration in the forward direction and for two typical cases of malfunction:

(a) *Normal deceleration*

While the grating is travelling at greater than critical speed, 'forward' pulses arrive at the Blanket Mono* MM1F rapidly enough to keep it in its quasistable state. G_1 is held open, MM2F is kept triggered and FF1F is at '1'.

When the grating slows down below critical speed, the pulses arrive at intervals greater than the period of MM1F, and are therefore routed through G_2. The MM2F remains in its quasistable state for 0·45msec after the last pulse which passes through G_1. If the deceleration is occurring at a permissible rate, then at least one more pulse arrives before the end of the period of MM2F. This last pulse returns FF1F to '0', closing G_3. The pulse produced by MM2F, as it returns to its stable state, is therefore prevented from reaching FF2.

(b) *Sudden stop*

If the pulse train is suddenly interrupted while the grating is moving at greater than critical speed, MM1F will return to its stable state within 0·19msec. However, no more pulses arrive during the next 0·45msec to switch FF1F to '0'; G_3 remains open and the pulse generated at the end of the MM2F period will pass through to trigger FF2, which in turn energizes the audio-oscillator and provides the signal voltage which inhibits the recording cycle. FF2 is reset together with the counters by the reset button.

(c) *Sudden reversal* (e.g. failure of P or Q waveform)

So long as the speed in the forward direction exceeds the critical value, FF1F remains at '1'. A reverse pulse arriving under these conditions will pass through G_3 and trigger FF2.

* *A 'Blanket Mono' is defined as a regenerative monostable circuit which will accept further input pulses while in its quasistable state and will return to its stable state when a specifiable time (its period) has elapsed since the last input pulse.*

Results

The equipment has been in full use for over 12 months. It has resulted in a great reduction of operator fatigue and operator error, and has considerably simplified the data reduction process for ballistic camera and other photographic records. The actual time required to take a given number of readings is reduced to less than half by use of the automatic recording feature. A further considerable saving in operator time is obtained through the elimination of the tape playback process.

Acknowledgment

The authors wish to acknowledge the work of Mr. P. Gillingham of Optical and Mechanical Instrumentation Group, Weapons Research Establishment, Salisbury, South Australia, who was responsible for the development of the optical equipment.

The permission of the Chief Scientist, Department of Supply, Australia, to publish this article is gratefully acknowledged.

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Automatic Control of Water Supply by Digital Telemetry

Westinghouse Automation, a division of Westinghouse Brake and Signal Co. Ltd, has recently completed the installation of a 'Westronic' control system and data handling equipment to automatically control the Sheffield Corporation Yorkshire Derwent water scheme. Water is extracted from the River Derwent at Elvington near York and after passing through a treatment works close to the river bank is pumped along a 40 mile pipeline to Sheffield, Leeds, Barnsley and Rotherham.

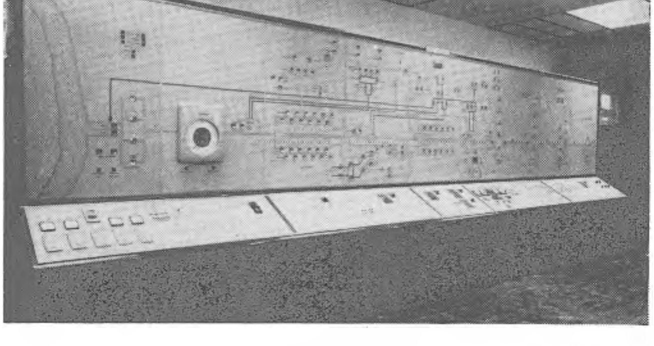
The pipeline is at present pumping 12·2 million gallons per day but this will be raised to 25 million gallons per day when the plant is operating at full capacity. The control centre at Elvington is linked to the outstations at Brayton Barff, Frickley, Hoober and Ringstone by the 'Westronic' digital telemetry system operating over a telephone line to provide automatic control and supervision from Elvington. There, a 28ft illuminated mimic panel provides an overall diagrammatic picture of the Elvington treatment plant and the entire pipeline and by automatic print-out provides a permanent record of events. The system can be left without attendance and an automatic alarm system functions to bring in staff for emergencies.

The 'Westronic' system scans 900 bits of information from five locations in 40sec while the data handling system deals with 500 functions in 12sec. Another 50 analogue functions, for reservoir depths, rate of flow and water pressures, are recorded automatically at the control centre. Chart recorders at the centre indicate various chemical plant functions.

Through the 'Westronic' system the pumping stations operate automatically, controlled by reservoir levels. Any pump in the entire pipeline can be operated automatically or manually from Elvington. Local manual control of pumps at each outstation is also provided. A safety device incorporated in the system automatically stops the required number of pumps farther back down the pipeline should failure of a pump at the delivery end occur thus preventing excess pressure causing a pipe burst.

Automatic print-out on an electric typewriter is provided for the controller showing water flow in and out of each pumping station and reservoir at preset time intervals. Another typewriter records up to 50 different functions on manual demand from the control desk. A third typewriter records the change of state of all quantities as they occur so recording automatically the indications on the mimic diagram.

The mimic display panel in the Elvington Control Room



The Design of Some Tunnel Diode Switching Circuits

By R. V. L'Archeveque*, B.Sc.A., Ph.D., and B. McA Sayers†, M.Sc., Ph.D., D.I.C.

Tunnel diode circuits for medium speed (20Mc/s) monostable and bistable action are recapitulated. Design equations for these circuits are generalized from timing calculations for a relaxation oscillator. A stable low level discriminator is also described.

(Voir page 704 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 711)

IN exploring the detailed behaviour of tunnel diode staircase circuits, the convenience has again become apparent of tunnel diode-transistor switching circuits as medium speed building blocks (e.g. switching up to 20Mc/s). Design procedures can be set up which allow calculation of such circuits, particularly for timing purposes, with fair preciion and great simplicity[1]. While circuits of this type are by no means unknown, it seems that the features of some of these circuits have been widely overlooked. By way of recapitulation, the monostable multivibrator, the bistable and the relaxation oscillator are described, and the latter is used as a basis for setting up design equations for application to the other circuits. Finally, a new conditionally-stable circuit as an adjustable level-detector with interesting properties is discussed. Elsewhere[2] a qualitative simplified account has been presented of the essential features of the switching process, arguing thence to the improved design of tunnel diode chains for fast sequential switching. The same principles apply here in these circuits, the timing action of which depends upon an inductive element; it will be seen that the use of an inductive timing element is convenient because of its high effective dynamic impedance during any fast switching operation.

Monostable and Bistable Circuits

In the monostable circuit, two timing intervals are important—the output pulse duration and the minimum period. The pulse width is determined by the time the circuit remains in the high voltage state. The minimum period, however, is determined by the time required for the circuit to return sufficiently near the stable point to allow a new switching cycle in response to the new input pulse and thus depends somewhat on the amount of input drive available.

The static operating point for the monostable circuit of Fig. 1(a) is shown as A on Fig. 1(b). The diode current is I_a, the transistor being cut off and its base current negligible. An adequate positive pulse at the trigger terminal causes current flow through R_{in}, C_{in} raising the instantaneous diode current at least beyond the peak current value I_p, point 2. The supply branch E_s, R_s, L_s of the circuit exhibits high dynamic impedance, while the transistor is effectively open-circuit. The triggering pulse current added to I_a serves to supply the diode peak current and to charge its shunt junction capacitance. Hence the diode voltage rises and rapid switching to point 3 ensues. As the input pulse falls to zero, the circuit operating point falls to point 4. At this instant the current I_a is made up of the diode current (in fact very nearly I_v), the balance $I_a - I_v$ supplying transistor base drive. This operating point 4 is unstable and as the current supplied through L_s falls to I_v, the transistor drive diminishes; when switching from point 5 to point 6 occurs the transistor is switched off under the influence of a reverse drive. Finally I_{TD} approaches the rest value of I_a at point 1. It follows that the duration of the triggered state is dominated by the time taken for the current in L_s to fall from I_a to I_v.

The bistable circuit is interesting and simple; it is identical in form with the monostable circuit of Fig. 1(a). However, the choice of load line and supply potential is made to allow two stable intersections on the diode characteristic as shown in Fig. 2. It will be recognized that the precise selection of stable points demands some consideration of driving pulse tolerances and duration, and of the switching operation itself. The choice of inductance is dictated partly by the need to complete the switching action within a minimum time to allow a given switching rate to be accommodated. The presence of the inductance is required to ensure that any input pulse current is caused to flow only in the diode (and transistor, if this is in the conducting state). Consider the circuit to be in the low voltage state, at operating point A; a brief input positive pulse of adequate magnitude causes the diode to switch to the high voltage state, turning on the transistor. If the inductance causes the circuit to remain near point P until the input pulse is terminated no reverse switching will occur due to the trailing edge of the pulse. The circuit will then settle to the second stable point B with the transistor in saturation.

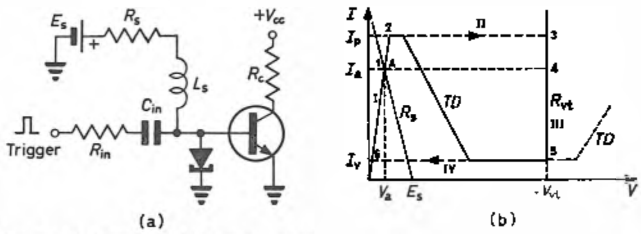


Fig. 1. (a) Tunnel diode monopulser or bistable; the choice of E_S and R_S determines the mode of behaviour

(b) Monopulser circuit linearized characteristics; the effect of the shunt transistor is approximated by R_{vt}

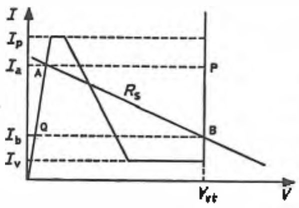


Fig. 2. Bistable circuit linearized characteristic; the stable points are A and B

When a second pulse of the same polarity is applied to the input, C_{in} charges rapidly, increasing temporarily the base current of the transistor (already saturated). At the end of the input pulse, C_{in} discharges causing, in particular, the diode current to fall below the valley current value. Switching to the low voltage state occurs and the circuit settles towards point A as the current through the inductance approaches its final value.

Thus by using a short input time-constant, switching of the bistable takes place on pulses of either polarity

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(employing successively the positive-going and negative-going edges of successive impulses). This action is made possible only by the action of the inductance in preventing double-switching at each AB transition. It is to be noted that as with all high speed switching devices adequate supply line decoupling is essential in using these circuits.

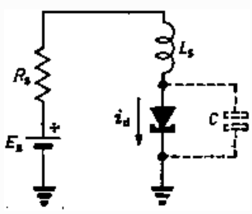

Fig. 3. Single tunnel diode relaxation oscillator

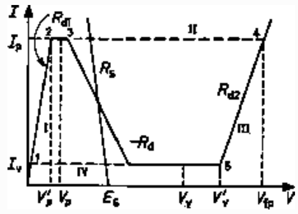

Fig. 4. Oscillator circuit linearized characteristic

Development of Design Equations

Relaxation Oscillator

Fig. 3 shows the circuit diagram of a tunnel diode relaxation oscillator; Fig. 4 is a piecewise-linear approximation of its characteristic.

Expressions for the various times determining the period of oscillation can be derived using the approximating characteristic of Fig. 4. The analysis relies on the assumption that L_s is large enough to ensure constant current switching from the low to the high voltage state and vice versa. Charging of the parallel capacitor C largely dictates the behaviour of the circuit in the regions 2 to 4 and 5 to 1. It is assumed for the analysis that the charging current is constant and equal to $I_p - I_v$. It is also postulated that L_s controls the changes in the current and that, in the two regions 1 to 2 and 4 to 5 (Fig. 4), the presence of the parallel capacitance may be neglected. In this way, the derivations are greatly simplified; the expressions for constant current charging of a capacitor and for the build up (or decay) of a current in an inductance-resistance network are respectively sufficient to yield the time spent in corresponding regions described by Fig. 4.

Loading of the Tunnel Diode by a Transistor

In most practical circuits to be discussed here, the tunnel diode is directly coupled to a common-emitter transistor, Fig. 5. Fortunately, the analysis previously given[3] needs only minor alterations to take into account the influence of the transistor.

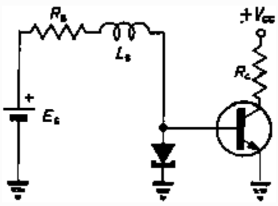

Fig. 5. Hybrid tunnel diode-transistor oscillator circuit

The forward voltage, V_{fp} for a germanium tunnel diode is of the order of 500mV. On the other hand, the base-to-emitter voltage for a medium power germanium transistors does not normally exceed 450mV, even at maximum rating. The collector current must therefore be limited to a safe value by a resistor R_c (Fig. 5) and the transistor saturates when the tunnel diode switches to the high voltage state. Fig. 6 reproduces the V_{BE}/I_B curve for a typical medium power high speed switching germanium transistor, superimposed on the characteristic of a tunnel diode. A piecewise-linear approximation of the combined characteristic is also illustrated. In Fig. 6 V_{vt} represents the voltage appearing between the base and the emitter when the transistor is in saturation. For the two typical transistors used, 2N797 (npn) and ASZ21 (pnp), and for the current range of interest (I_c about 10 to 15mA), that voltage lies between 350mV to 450mV. For design purposes, V_{vt} may be selected as 400mV. The maximum base current level at which the transistor enters the saturation is adjusted below I_v by limiting the collector curren with R_c according to the minimum value of current gain β

The input dynamic resistance (R_{vt} in Fig. 6) of a saturated transistor in the common-emitter configuratior is nearly equal to the forward resistance of the emitter-base diode at a point corresponding to the base curren level, and is therefore very small (using a typical 1N294] $R_{vt} \sim 6\Omega$). For values of $E_s \geqslant \frac{1}{2}V_v$, the method of analysis tends to underestimate the transition time from 3 to 4 (Fig. 6). The error can be compensated in practice to a large extent by taking $R_{vt} = 0$ in the calculations, thus increasing the time-constant $L_s/(R_s + R_{vt})$ in the equations which will now be given.

Timing Calculations for the Relaxation Oscillator

Fig. 7 shows the two linearized networks representing circuit behaviour in the two dominant phases of operation The following expressions result:

Region I—from 1 to 2, (Fig. 6)

$$t_{1\text{-}2} = (L_s/R_{t1}) \ln \frac{(E_s - R_{t1} I_1)}{(E_s - R_{t1} I_p)} \quad \ldots\ldots\ldots \text{(1)}$$

where $R_{t1} = R_s + R_{d1}$

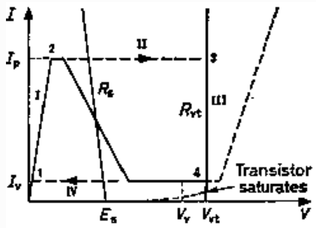

Fig. 6. Hybrid relaxation oscillator linearized characteristic

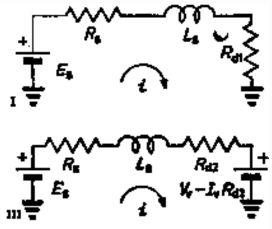

Fig. 7. Linearized equivalent networks representing circuit action in the dominant timing phases I and III of the oscillation cycle

Region II—from 2 to to 3, (Fig. 6))

$$t_{2\text{-}3} = \frac{(V_{fp} - V_p)}{I_p - I_v} C \quad \ldots\ldots\ldots\ldots\ldots \text{(2)}$$

C represents the total capacitance in parallel with the tunnel diode, including the junction capacitance.

During the switching process, some current is drawn by the transistor before it saturates, thus reducing the capacitor charging current in the last phase of the transition to point 3. It is possible readily to compensate for neglect of this effect by using V_{fp} instead of V_{vt} in equation (2); however the interval $t_{2\text{-}3}$ is often too small to contribute significantly to the oscillator period.

Region III—from 3 to 4 (Fig. 6)

$$t_{3\text{-}4} = (L_s/R_s) \ln \frac{(E_s - V_{vt} - R_s I_p)}{(E_s - V_{vt} - R_s I_v)} \quad \ldots\ldots \text{(3)}$$

Region IV—from V_v to 1 (Fig. 6)

The interval involved here is extremely short, but to calculate the order of magnitude concerned in all such circuits it is easiest to compensate for factors such as non-linear shunt capacitance, by assuming a capacitance charging current of $I_p - I_v$.

$$t_{4\text{-}1} = \frac{(V_v - V_1)}{I_p - I_v} C \quad \ldots\ldots\ldots\ldots \text{(4)}$$

Equations (1), (2), (3) and (4) can be used to compute the period of a tunnel diode relaxation oscillator loaded by a common-emitter transistor, to an accuracy better than 5 per cent. It is convenient to re-arrange these equations to give a useful design expression for the magnitude of inductance, L_s (in terms of circuit parameters), in order to produce a given period of oscillation, T.

$$L_s = \frac{T - CK_3}{K_1 + K_2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

$$\left.\begin{aligned} \text{where } K_1 &= (1/R_{t1}) \ln \frac{(E_s - R_{t1} I_v)}{(E_s - R_{t1} I_p)} \\ K_2 &= (1/R_s) \ln \frac{(E_s - V_{vt} - R_s I_p)}{(E_s - V_{vt} - R_s I_v)} \\ K_3 &= \frac{V_{fp} + V_v - V_p - V_1}{I_p - I_v} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots \quad (6)$$

With corresponding adjustment of the parameters contained in K_1, K_2 and K_3, the same equation can be applied to the design of tunnel diode monostable, bistable or conditionally stable circuits.

The Tunnel Diode Monostable Circuit

Equation (5) derived for the relaxation oscillation can still be used to compute the maximum acceptable value of L_s which would allow the present circuit (Fig. 1(a)) to be operated by pulses separated by a minimum time interval, T. In this case, the expression for K_1 must contain one additional factor to take into account the presence of a stable point in the low voltage region (point A, current I_a, in Fig. 1(b)). The factors K_2 and K_3 remain the same and are obtained from the expressions given for the relaxation oscillator (except that I_p is replaced by I_a in the expression for K_2).

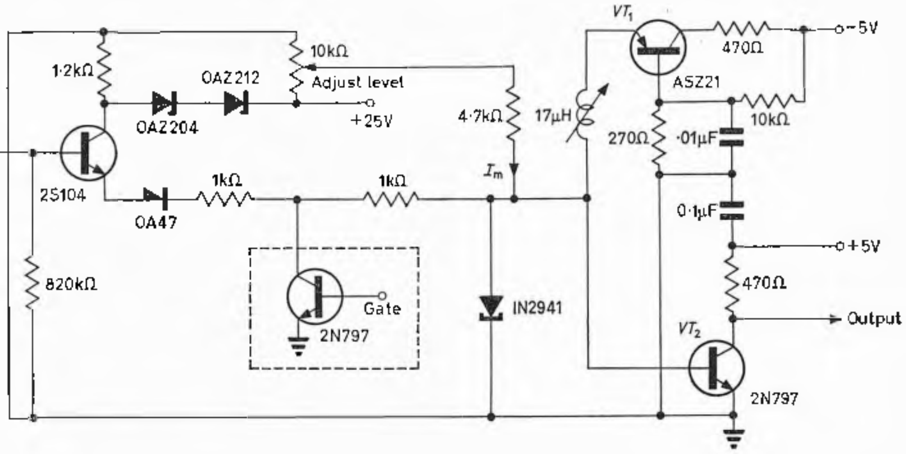

Fig. 8. Level detector circuit for positive going signals; for negative going signals: 2N2412 replaces 2S104, 2N797 (VT_2) interchanges with ASZ21 (VT_1), supply polarities, diode, Zener diodes and tunnel diode (1N2941) all reversed

$$\left.\begin{aligned} K_1 &= (1/R_{t1}) \ln \frac{(E_s - R_{t1} I_v)}{(E_s - R_{t1} N I_a)} \\ N &= \frac{(I_p)_{max} - (I_{in})_{min}}{I_a} \end{aligned}\right\} \quad \ldots\ldots\ldots \quad (7)$$

The factor N is a function of circuit tolerances. It is dependent on the maximum distance that the instantaneous operating current in the tunnel diode can depart from I_a at the time of arrival of a second input pulse to allow successful triggering under worst conditions.

The Tunnel Diode Bistable Circuit

The design equation (5) can be applied to the bistable circuit (also Fig. 1(a)) to compute the maximum value of inductance as a function of the minimum specified time interval between switching pulses. Two different cases must be considered.

(1)—*Switching from B to A* (Fig. 2).

In this case, the term T in equation (5) represents the minimum spacing between a pulse triggering the circuit from B to A and the next one switching it from A to B.

$$\left.\begin{aligned} K_1 &= (1/R_{t1}) \ln \frac{E_s - R_{t1} I_b}{E_s - R_{t1} N I_a} \\ K_2 &= 0 \\ K_3 &= \frac{V_v - R_{d1} I_b}{I_p + I_{in(b)} - I_b} \end{aligned}\right\} \quad \ldots\ldots\ldots \quad (8)$$

where N is as described for the case of the monostable; I_a and I_b are defined as in Fig. 2; $I_{in(b)}$ is the current available from the pulse triggering the circuit from A to B as determined by the input gating network.

(2)—*Switching from A to B* (Fig. 2).

In this case, the interval T in equation (5) represents the minimum spacing between a pulse triggering the circuit from A to B and the next one, switching it from B to A.

$$\left.\begin{aligned} K_1 &= 0 \\ K_2 &= 1/R_s \ln \frac{(I_a - I_b)}{M I_b} \\ K_3 &= \frac{V_{fp} - V_p}{I_{in(a)} + I_a - I_v} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots \quad (9)$$

where I_a and I_b are defined in Fig. 2, $I_{in(a)}$ is the current available from the pulse triggering the circuit from A to B; M = maximum relative difference from I_b of the current in the inductance on arrival of a second input pulse. This factor is a function of the tolerance in the circuit.

$$M = \frac{(I_{in(b)})_{min} + (I_v)_{min} - (I_b)_{max}}{I_b} \quad \ldots\ldots \quad (10)$$

A detailed design procedure, taking into account all tolerances in the circuit is given elsewhere[1].

The Level Detector

Operating Principle

Another circuit of general interest, used in a quantizer[2], acts as a level detector, or voltage discriminator, (Fig. 8, and in basic form, Fig. 9).

The tunnel diode peak current point is used to define a detected level of input signal; a single pulse, say, is required to mark the instant of detection in the quantizer. An increase of applied signal level, V_{in}, from zero, causes a corresponding increase of the diode current, all input current being taken by the diode until the peak current level is reached. The diode then switches to the high voltage state and the resulting voltage change initiates the diversion of the input current into the emitter of the transistor VT_1 and away from the diode; the diode is thus caused to reset to the low voltage state. The system can be well described, in terms employed with the relaxation oscillator, by the characteristic of Fig. 10. This figure shows a linear approximation of the electrical static characteristic corresponding to the circuit of Fig. 9; the transistor VT_1, (detecting the change to the high voltage state) is considered as a load on the tunnel diode; the presence of VT_1 can be taken into account in a manner previously explained. Frequently the base of VT_1 is biased with a small negative voltage, V_B, such that the input characteristic of this transistor in a common base con-

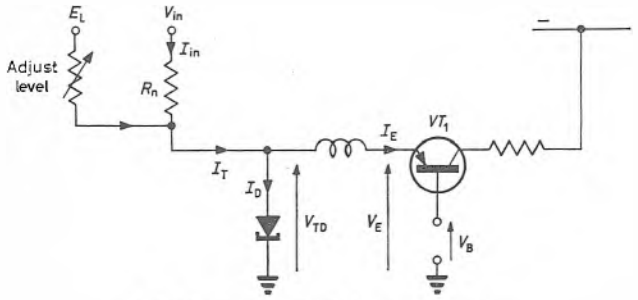

Fig. 9. Essential level detector elements

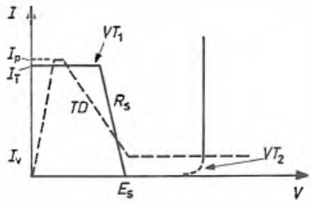

Fig. 10. Transistor VT_1 acts as non-linear load for the tunnel diode in the level-detector circuit

figuration, drawn as a load line (see Fig. 10), does not intersect the tunnel diode curve above the valley voltage, V_v.

A relaxation oscillation is sustained as long as the input current, due to V_{in}, is greater or equal to I_p. However, if V_{in} is decreased such that the input current is smaller than I_p, the circuit becomes stable at a point below the peak of the diode characteristic.

The combination of the inverted transistor characteristic and the diode characteristic, Fig. 10, is useful in describing certain dominant features of this circuit. Referring to Fig. 9 and neglecting VT_2, it can be seen that as V_{in} is increased slowly, the total current I_T, made up of the input current plus the steady bias current, is taken entirely by the diode until $I_T = I_p$.

Switching to the high voltage state occurs and transistor VT_1 begins to conduct. Assuming that the total current remains at I_T, the question now arises as to whether the high voltage state ultimately reached is a stable state. In any stable or quasi-stable state which is reached the diode voltage V_{TD} and the emitter voltage of VT_1, V_B, must be equal; of the total current I_T the diode takes an amount I_{TD}, the remainder $(I_T - I_{TD})$ being available for the transistor. The characteristic of Fig. 10 describes this situation and the higher voltage settling point is thus represented by the intersection of the diode characteristic with the inverted transistor characteristic. The effective base-line to which the inverted emitter characteristic is referred, is the value of total current I_T. If V_B is increased or decreased positively, the VT_1 characteristic shifts laterally to the right or left respectively. The behaviour of the circuit after switching is determined by this intercept point; if it occurs above the valley of the diode characteristic, the circuit will act as a level-sensitive bistable. If the intercept occurs below the diode valley point a stable high-voltage point will not be found and reverse switching to the low-voltage state will occur; the circuit will act as a pulse generating discriminator.

In this case the circuit hysteresis depends upon the change of I_T required to take this intercept stably above and below I_p thus initiating or inhibiting forward switching. The hysteresis is associated with the diode characteristic near the peak current, and this has been found to be well-defined and stable. It will be seen that it is the use of a non-linear load for the diode which allows this feature to be exploited so that hysteresis is determined stably at the peak current point. That the circuit output is a pulse train need be no inconconvenience for other, low speed uses since the pulses may be low-pass filtered to produce a d.c. level shift to mark level detection.

The method of analysis developed for the tunnel diode relaxation oscillator can be applied here without any modification. Measurement have confirmed that equation (5) predicts the period of oscillation of the circuit under study with an accuracy better than 5 per cent for $I_{in} = I_p$, $E_s = 200$ to 300mV and for pulse repetition rates up to 10Mc/s. (Note that E_s is measured across the tunnel diode terminal points when VT_2 and the diode are removed, and $I_T = I_p$.)

Temperature Coefficient

The magnitude of the input voltage necessary to trigger the tunnel diode can be written as in equation (11).

$$E_L = V_{BE} + V_1 + R_n(I_p - I_m) + V_p \quad \ldots . \quad (11)$$

where E_L = input voltage necessary to bring the current in the tunnel diode to its peak point.

V_{BE} = input transistor base-emitter voltage when the emitter current is equal to $(I_p - I_m)$.

V_1 = series diode voltage when $I = I_p - I_m$.

I_m = current supplied to the tunnel diode by the level adjustment network (Fig. 8).

The differentiation of equation (11) with respect to temperature yields equation (12) for the temperature of the level detector. It is assumed that the current, I_m, and the resistor, R_n, have a negligible temperature sensitivity.

$$(dE_L/dT) = (dV_{BE}/dT) + (dV_1/dT) + (R_n dI_p/dT) + (dV_p/dT) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (12)$$

The first two terms of equation (12) represent the temperature coefficient of the voltage drop across semiconductor diodes and approximate $-2{\cdot}5$mV/°C for both germanium and silicon. The last term of equation (12) is[3] of the order of -60μV/°C. If $R_n dI_p/dT$ is adjusted to a value of about +5mV/°C, a very small temperature coefficient is obtained for the level detector. In fact, this term can be adjusted easily to produce a positive or a negative temperature coefficient if desired. A good estimate of the coefficient dI_p/dT of a tunnel diode can be obtained[3] from a knowledge of the peak voltage, V_p, and selection is relatively simple. Alternatively, the resistor R_n can be adjusted to alter the overall temperature coefficient of the detector, the level of discrimination being re-established with I_m.

Conclusion

Hybrid tunnel diode-transistor circuits for medium speed switching operations may well be convenient for many applications. Attention is drawn particularly to the simplicity of the various circuits and the design procedure outlined here, to the flexibility with which tunnel diode chains can be used as fast-stepping staircase level generators of some precision[2], and to the special features of the amplitude discriminator discussed above. The circuit simplicity (and the simple polarity inversion) makes such circuits of special interest for laboratory purposes; it is important to note the advisability of adequate decoupling (by capacitor by-passing each supply line to ground at the circuit input point) as usual with high speed switching devices.

Acknowledgment

This article includes some material taken from a University of London thesis by R. V. L'Archeveque.

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High Voltage Generation from Piezoelectric Ceramics

By C. Pearcy*

Recent developments in piezoelectric materials have led to the production of high voltage impulse sources based on the direct conversion of mechanical energy to electrical energy. This principle has been exploited in the production of spark generators incorporating PZT† (lead zirconate titanate) ceramic cylinders. Devices of this type have been successfully adapted for the ignition of various gas appliances, missile impact fuses and for low current high voltage power supplies such as those needed for portable X-ray equipment.

The purpose of this article is to outline the various design techniques needed to fabricate such a system. Design equations are listed and explained together with the relevant physical characteristics of PZT ceramic material.

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

THE piezoelectric effect is a phenomena exhibited by certain single crystals, of which quartz and rochelle salt are the most well known. In recent years certain polycrystalline ceramic materials have been shown to exhibit a strong piezoelectric effect, and of these barium titanate and lead zirconate have been the most widely used. Ceramic elements of this type can be fabricated into a variety of shapes and sizes, the most useful shape being the simple disk or cylinder.

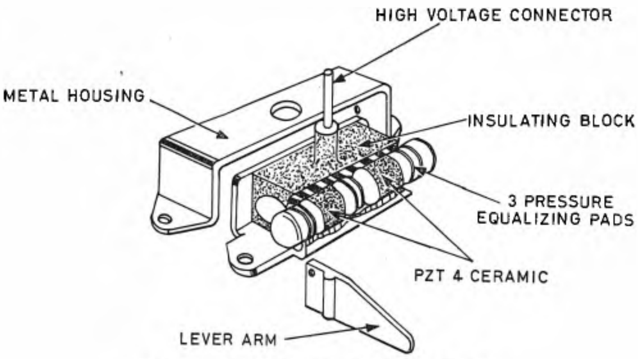


Fig. 1. Arrangement of 'Spark Pump'

If a piezoelectric crystal or ceramic element is coated with conducting electrodes on certain faces and it is subjected to mechanical strain, an electrical charge will appear across these electrodes. Strain can be introduced by pressure or tension applied between two faces of the element, the resulting charge produced being directly proportional to the induced strain. Since a ceramic material constitutes a dielectric it can be considered as a low loss capacitor, while the charge induced in it appears as a voltage across the electrodes. This charge can be released in the form of a high voltage spark when the voltage rises to a certain level, an equal charge of opposite polarity is produced when the initial strain is released. This second charge produces another spark which will travel in the opposite direction to the first. The successful exploitation of this principle for high voltage generation makes certain electrical and mechanical parameters of the material particularly important.

The piezoelectric ceramic needs to have high mechanical strength, a fairly high dielectric constant with low leakage, a high voltage output coefficient (known as the *g* coefficient), and it must retain these properties over a wide range of operating pressures and temperatures. The only known materials possessing these properties are the range of high efficiency lead zirconate titanate ceramics with the trade name of PZT, which are manufactured in the United Kingdom by Brush Clevite Co. Ltd. The most suitable PZT material for this application has the type number PZT-4. PZT materials and a piezoelectric ignition system known as the 'Spark Pump'‡ were both developed initially by Clevite Corporation of U.S.A., and are covered by several U.K. and foreign patents[1].

The 'Spark Pump'

The 'Spark Pump' is a typical example of a practical system for producing high voltages from PZT. This system, although not available in the U.K., is described here in detail to illustrate one way in which PZT can be used to make a successful high voltage generator. The important design considerations and performance specifications are included for those who wish to fabricate their own systems for a specific application. It was originally developed for the ignition of internal combustion engines such as outboard motors for small boats, lawn mowers, etc. The overall volume of the complete assembly is about $3\frac{1}{2}$in^3 and it weighs approximately 3oz.

When the 'Spark Pump' is incorporated in this type of motor, the starting rope or pedal is eliminated, being replaced by a small hand operated lever which generates sufficient voltage from the PZT for starting purposes. The 'Spark Pump' utilizes two ceramic PZT-4 cylinders $\frac{1}{2}$in long and $\frac{3}{8}$in diameter which are enclosed in a housing made of a low leakage plastic material. This housing is itself contained in a metal framework as shown in the exploded view of Fig. 1.

The two PZT cylinders are arranged mechanically in series and connected electrically in parallel, with the two electrodes of like polarity in contact with a conducting metal disk in the centre. From this conducting disk an insulated lead is taken out through the metal housing. Electrical contact is made between the metal housing and the two outer electrodes by means of washers so that this effectively 'earths' the system. The elements, washers and metal disk are prestressed by an adjustable bolt sufficient to take up any slack movement in the system.

The lever arm shown can be driven by an eccentric

* *Brush Clevite Co. Ltd.*

† *PZT is a registered trade name of Brush Clevite Co. Ltd.*

‡ *'Spark Pump' is a registered trade name of Clevite Corporation.*

on the engine crankshaft, camshaft or by a secondary lever which is hand operated. The lever arm thus exerts a smoothly varying axial pressure on the PZT elements, giving an overall contraction in length of about 0·00009in for each 1 000 lb/in^2 of stress applied, under open circuit conditions. If the charge generated is totally spark discharged as in short-circuit conditions a further overall contraction of ·00007in occurs. This contraction is very small due to the high Youngs modulus of PZT, and it is necessary to use pressure equilization pads as shown, consisting of conducting rubber or soft metal foil to give an even stress distribution across the ceramic/metal interfaces.

With a lever multiplication of 15:1 a static force of 80 lb will stress the PZT to about 7 000 lb/in^2 generating a voltage of approximately 22kV. With stresses of this order it has been established that the operating life of the system would considerably exceed 10^9 cycles of operation. The PZT elements used have an electrical capacitance C_o of 40pF each, giving an overall capacitance of 80pF. Hence, with 7 000 lb/in^2 stress applied, the energy output is 0·0195 joules for each half cycle of operation.

The 'Spark Pump' produces its full rated voltage at all engine operating speeds. The actual voltage rise time for such a system to give 10kV at the spark plug is as short as 10nsec. This is about 3 000 times faster than a magneto sparking system. A comparison of the voltage response of these two types of ignition system is shown graphically in Fig. 2.

Due to this very fast response time, a further advantage of the spark pump over a magneto system is that it can tolerate over 100 times lower shunt resistance such as that encountered due to fouled plugs. This comparison is shown graphically in Fig. 3.

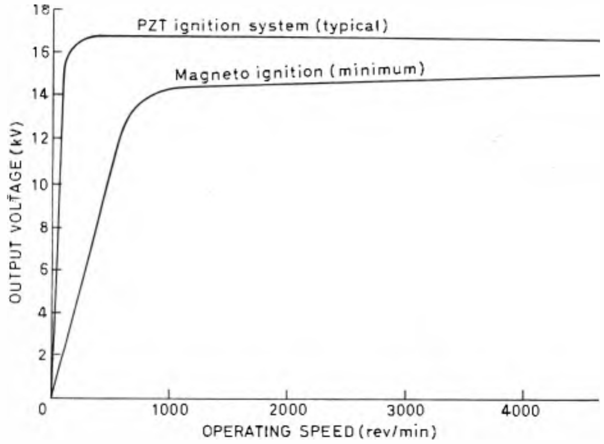
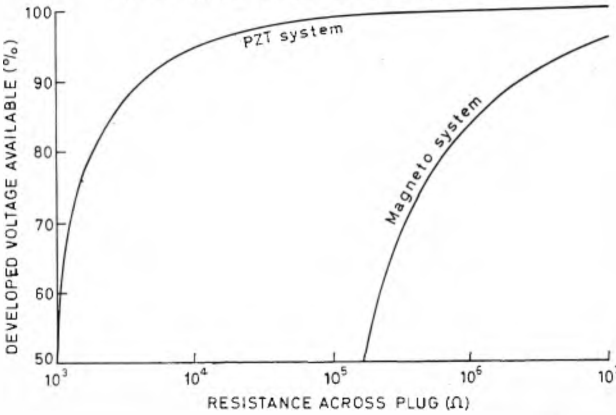


Fig. 2. Comparison of output voltage/speed characteristics

Fig. 3. Output voltage/shunt resistance curves



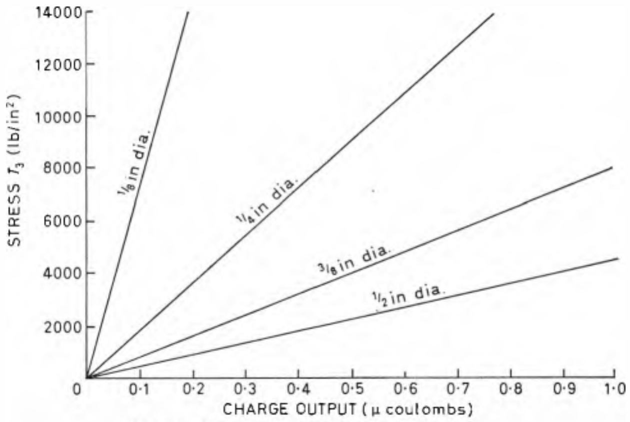


Fig. 4. Output charge variation with stress

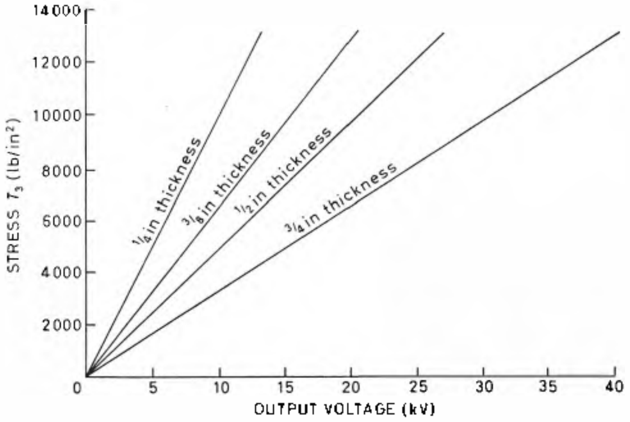


Fig. 5. Output voltage variation with stress

Useful Stress/Voltage Characteristics of PZT

When a PZT element is stressed to a value T_3, the charge density produced Q_3/A is given by the equation:

$$Q_3/A = d_{33}T_3 \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

Where Q_3/A is in coulombs/metre^2, d_{33} is the piezoelectric charge output coefficient in coulombs/newton and the stress T_3 is in newtons/metre^2. The first suffix denotes the plane of polarization and the second denotes the plane of mechanical deformation within the ceramic element. In the case of a PZT cylinder of diameter D metres, the actual charge developed would be:

$$Q_3 = 2D^2T_3 \times 10^{-10}\,\text{coulombs} \quad \ldots\ldots\ldots \quad (2)$$

This relationship is shown graphically in Fig. 4 for several PZT standard diameter cylinders.

For a constant applied stress the voltage output of a PZT element is proportional to the thickness. The equation connecting the open-circuit voltage V_3, the piezoelectric voltage output coefficient parallel to the direction of polarization g_{33} and the applied stress is:

$$V_3 = g_{33}T_3t \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Where V_3 is in volts, g_{33} is in volt metre/newton and t is the thickness of the ceramic in metres. This relationship between the output voltage and applied stress for PZT-4 cylinders of several thicknesses is shown in Fig. 5.

The energy output W at very low frequency operating conditions, such as in the case of gas lighting appliances, is given by the capacitor equation:

$$W = \tfrac{1}{2}C_oV_3^2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Where C_o is the low frequency capacitance of the ceramic element and when W is expressed in joules. From this equation it can be shown that for a PZT-4 cylinder diameter D metres used in axial compression:

$$W = 2{\cdot}4\, D^2 t T_3^2 \times 10^{-12} \text{ joules} \quad \ldots\ldots\ldots \quad (5)$$

Thus the variation of energy output against stress for PZT cylinders of two standard sizes can be expressed as in Fig. 6.

The maximum safe stress level under which PZT-4 will retain its linear piezoelectric properties is approximately 10 000 lb/in². A further voltage output may be produced at stresses exceeding 75 000 lb/in² due to the ferroelectric effect. This ferroelectric output is the remanent charge which was originally applied to the ceramic in order to give it piezoelectric properties. Once this remanent charge is extracted, however, the effect is not reversible under these conditions and can only be restored by the polarization process.

The contraction x of a ceramic element may be calculated for open-circuit and short-circuit conditions from the stress T_3, Youngs modulus Y_{33} and the thickness t.

$$\text{Open-circuit } x_{33}{}^{D} = T_3 t / Y_{33}{}^{D} \quad \ldots\ldots\ldots \quad (6)$$

$$\text{Short-circuit } x_{33}{}^{E} = T_3 t / Y_{33}{}^{E} \quad \ldots\ldots\ldots \quad (7)$$

Where $Y_{33}{}^{D}$ and $Y_{33}{}^{E}$ are the open-circuit and short-circuit values of Youngs modulus in newtons/metre², and dimensions are in metres.

At very low frequency operation the equivalent circuit of any PZT element may be shown as in Fig. 7.

The values of the capacitance C_o and the shunt internal loss resistance R_i can be calculated from the following equations:

$$C_o = KA\epsilon_0 / t \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

Where K is the dielectric constant, A is the cross-sectional area in square centimetres, t is the thickness between electrodes in centimetres, and ϵ_0 is the permittivity of free space.

In the case of a PZT-4 disk of diameter D cm at 25°C

$$C_o = \frac{83{\cdot}4\, D^2}{t} \quad \ldots\ldots\ldots\ldots\ldots \quad (9)$$

The value of shunt internal leakage R_i may be calculated from the resistivity σ in ohm centimetres, the thickness t, and the cross-sectional area A.

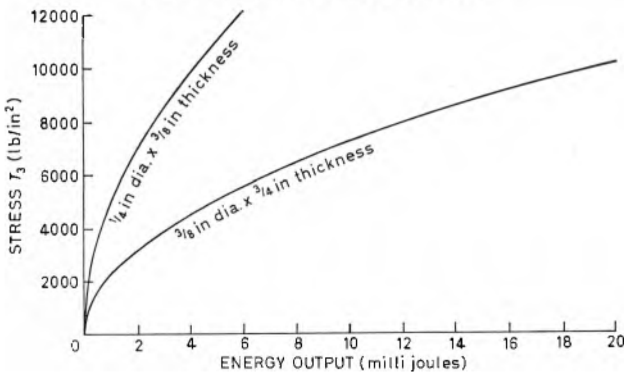
$$R_i = \sigma t / A \text{ ohms} \quad \ldots\ldots\ldots\ldots \quad (10)$$

For a PZT-4 disk at 25°C:

$$R_i = \frac{1{\cdot}27 \times 10^{12} \times t}{D^2} \text{ ohms} \quad \ldots\ldots\ldots\ldots \quad (11)$$

If a PZT element is subjected to continuous sinusoidal operation at a certain frequency f cycles/second which is

Fig. 6. Variation of energy with stress



well below the natural resonant frequency of the element, the value of shunt loss resistance is dependent on frequency. R_p may then be calculated from:

$$R_p = \frac{1}{2\pi f C_o \tan\delta} \text{ ohms} \quad \ldots\ldots\ldots\ldots \quad (12)$$

where $\tan\delta$ is the dielectric loss factor.

When PZT systems are used for the ignition of gases where only one cycle of operation is normally required, an unwanted side effect occurs in the creation of a shock wave around the spark. This shock wave sounds rather like a miniature explosion, and it can be considerably reduced by increasing the time-constant of the circuit. Although the 'Spark Pump' principle easily ignites the majority of fuels satisfactorily, when gases with no free hydrogen present and with a rather high flash point are being used it may not be such a successful ignition method. The shock wave tends to blow the gas away from the spark, thus reducing the chance of ignition. By increasing the circuit time-constant, the spark becomes more 'spread out' and is maintained for a longer period with correspondingly less shock.

The most convenient method of suppressing the shock in this way is by introducing a series resistor into the circuit. The optimum value of such a resistor is dependent on the particular ignition properties of the gas to be lit, but practical experience has shown that it should be of the order of 500kΩ.

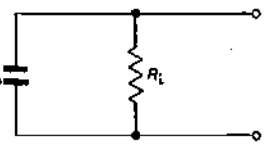


Fig. 7. Equivalent circuit of PZT element

Apart from the normal squeezing technique for producing high voltage from PZT, it has been shown that a voltage can also be produced by subjecting it to a transient stress by means of a hammer blow[2]. Such a blow creates a compressional stress wave within the ceramic causing a short series of alternating electric fields to be produced across the electrodes as the ceramic vibrates. This is due to the stress wave reflecting backwards and forwards several times within the PZT causing it to resonate. When a spark gap is connected to an element stressed in this way a short burst of sparks is produced. Due to uneven stress distribution within the PZT, the voltage produced in this way can be at least 10 per cent below that produced by the squeezing system. However, due to the resonance effect within the element, the output impedance is lower and more resistive. Thus, losses due to stray capacitance are much less important than in the case where the PZT elements are 'squeezed'. A short series of sparks produced by one hammer blow, will, for example, ignite natural gas more reliably than the two sparks produced in the other method if no extra resistor is used.

If the hammer mechanism is constructed to have a low mechanical loss, a further number of spark producing pulses may be obtained by the hammer bouncing on the PZT. This series of sparks will continue until the damping of the system reduces the voltage below that required to break down the gap.

When designing a transient system it is advisable to place a disk of aluminium or a similar material between the hammer and the PZT in order to spread the stress wave evenly across the ceramic surface. This has the effect of increasing the efficiency, and also radically reduces the possibility of the ceramic shattering due to uneven contact with the hammer. Care should also be taken to limit the peak value of induced compressive stress in the PZT during impact to about 12 000 lb./in². The peak stress T_3

in newtons/metre^2 can be calculated approximately when impact occurs between a hammer of cross-sectional area A_H metre^2, and a PZT element of cross-sectional area A_C, from the equation:

$$T_3 = -\frac{A_H U Y Y_{33}^D}{A_C v Y_{33}^D + A_H v_3^D Y} \quad \ldots\ldots\ldots (13)$$

Where Y and Y_{33}^D are the values of Youngs modulus in newtons/metre^2, and v and v_3^D are the compressional wave velocities in metres/second in the hammer material and PZT respectively. The velocity of impact being U metres/second.

This formula shows that if the hammer is made of steel, and has the same cross-sectional area as the ceramic, the maximum safe impact velocity is about 7m/sec. This velocity can be exceeded when an intermediate 'buffer' layer is used as suggested above, the value then also depends on the elastic properties of the material used to distribute the stress.

A PZT spark ignition system does not generate any appreciable internal heat, but prolonged exposure to temperature over 150°C is not recommended. The voltage output/unit stress is substantially independent of temperature up to about 120°C.

In designing high voltage generator systems based on the above principles, the following characteristics of PZT-4 ceramic can be used. These values given are valid for cases where an average operating stress of about 3 000 lb/in^2 is applied parallel to the direction of polarization of the PZT, i.e. the '33' direction.

Dielectric constant $K = \epsilon_{33}/\epsilon_0 = 1200$

Dielectric loss factor at 1kc/s, $\tan\delta = 0{\cdot}005$

Piezoelectric charge output
coefficient $d_{33} = 255 \times 10^{-12}$C/N

Piezoelectric voltage output
coefficient $g_{33} = 24{\cdot}0 \times 10^{-3}$V.m/N

Youngs Modulus, open-circuit
conditions $Y_{33}^D = 11{\cdot}4 \times 10^{10}$N/m^2

Youngs Modulus, short-circuit
conditions $Y_{33}^E = 6{\cdot}71 \times 10^{10}$N/m^2

Velocity of a compressional wave
open-circuit $V_3^D = 4\,600$m/sec

Maximum safe compression 10^8N/m^2

Resistivity at 25°C $\sigma = >10^{10}\Omega$.m

Resistivity at 100°C $\sigma = >10^{10{\cdot}5}\Omega$.m

Maximum operating temperature = 150°C.

Conclusions

The development of practical and economic PZT ignition systems for a variety of purposes has opened up an entirely new field. The output energy available from PZT elements is considerably more than sufficient to ignite most air/gas mixtures. For example, an optimum mixture of methane and air can be ignited with a spark of only 0·3 millijoule. The energy transfer from a spark discharge can be very efficient, thus making it an attractive proposition for ignition purposes. Since the piezoelectric effect is everlasting the PZT elements do not need replacing, thus reducing the need for visits by service engineers, and eliminating replacement costs. This is a very important advantage compared with rival systems such as battery and flint operated devices. For the ignition of natural gases PZT has been shown to be vastly superior to battery operated red hot wire igniters. The most successful applications of PZT so far exploited include fully integrated ignition systems for gas fires and space heaters, portable igniters for domestic and industrial purposes and gas filled cigarette lighters. Using the same principle impact fuses for missiles can also be made using PZT elements to operate a detonator. Recently voltages exceeding 1MV have successfully been produced by a PZT portable d.c. power supply for X-ray apparatus.

Acknowledgment

Acknowledgment is due to the Directors of Brush Clevite Co. Ltd for their encouragement to write and permission to publish this article.

REFERENCES

1. British Patents 780 673, 818 710, 885 994, 918 925 and 918 926.
2. British Patent No. 712 803 and U.S. Patents 2 649 488, 2 717 589 and 2 717 916.

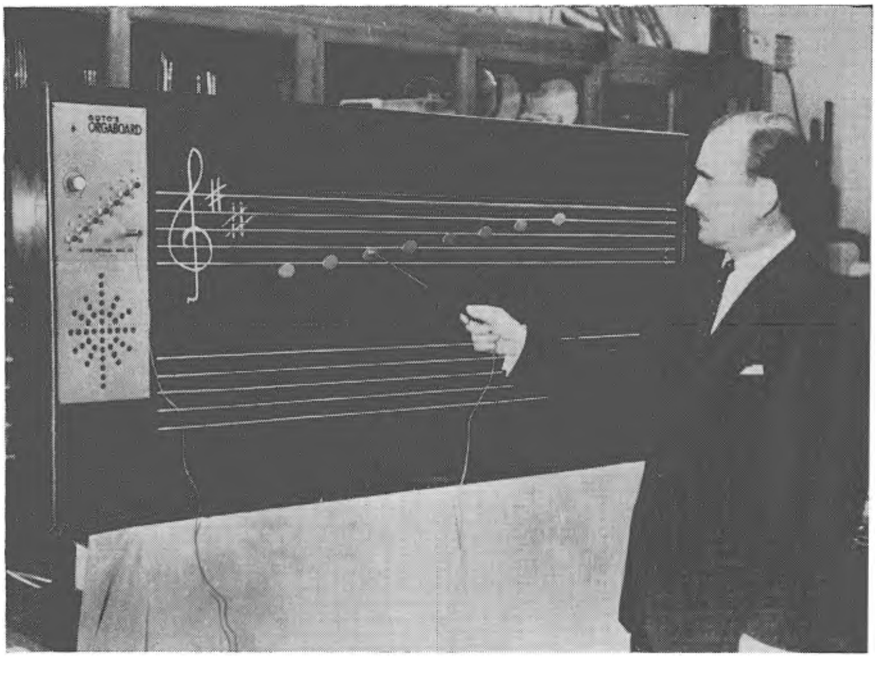
A 'Singing Blackboard'

The photograph shows an electronic device for use by music teachers in primary education schools for auditory instruction in score reading and composition of music.

Known as a 'Singing Blackboard' it consists of a transistorized oscillator, amplifier and loudspeaker mounted behind the blackboard on the face of which are two sets of lines forming the musical staves.

Jacks with metal heads shaped like notes can be plugged into the blackboard and when touched with a special pointer the corresponding musical note is produced. Each note can be altered in pitch by switches on the left-hand side of the board to sound natural, sharp or flat as required and twelve notes are thus available.

The apparatus, a prototype manufactured by Goto of Tokyo, has been imported by Boosey and Hawkes and is being studied for its possible use for education in music.



AN ANALOGUE COMPUTER . . .

. . . for Forecasting and Stock Control

By R. Calvert* and J. Mildwater*

Recent advances in operational research have resulted in improved methods of short term forecasting. The computer described employs ratio transformer techniques to solve a series of equations giving statistical information for forecasting and stock control.

A general description of circuit arrangements for addition, subtraction, multiplication and division is followed by the more detailed application of the system to a particular set of equations.

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

NEW operational research methods for routine forecasting have been applied most successfully in devising nodern stock control procedures. Typical results have been o achieve reductions of more than 25 per cent while at he same time reducing the frequency of stockouts. By hese means capital invested in stocks can be released for rofitable investment elsewhere. One of the practical probems in applying these methods on a routine basis is proiding the means of making the necessary calculations ndividually for each stock item. Even with access to a igital computer there are practical problems in applying hese methods due to the difficulty in transmitting data o and from a central computer sufficiently quickly. A entral computer can also, in some cases, have the disadantage that local knowledge of product demand is not asily included in the assessment of stock requirements. 'he need for a means of carrying out the calculations ocally and with the necessary speed has led to the deelopment of an analogue computer for this specific urpose.

* *The Wayne Kerr Laboratories Ltd.*

The Importance of Forecasting

No forecast can be expected to be perfect; errors in forecasts must arise. The approach of operational research is to accept this and to measure and constantly monitor the size of these errors so that the maximum information is available for the best management decisions to be reached under the prevailing circumstances. This simple concept is very important in modern stock control procedures.

Having accepted that forecast errors will arise, it follows that allowance should be made for such errors in planning stock levels. By the continuous measurement of errors in forecasting the future demand of each product, it is possible to ensure that stocks held are as small as possible but nearly always sufficient to meet the demands which arise.

The system has been designed to

(a) Compute short-term forecasts
(b) Assess the possible errors in the forecast
(c) Decide the stock required for future periods
(d) Control the risk of running out of stock.

The Forecast System Used

The simplest form of statistical forecasting employs the arithmetic mean of the values obtained over a certain number of past periods. This mean gives equal weight to both past figures and the most recent. Alternatively, by giving less weight to earlier figures, the mean will be influenced more by recent events. In a system employing exponential smoothing the weight given to past periods compared with the most recent can be varied to suit the conditions. For example, sometimes more accurate forecasts can be obtained by paying little attention to the demand which existed several periods ago; whereas under certain conditions the reverse is true and little notice should be taken say, of a sudden upsurge in demand, since based on previous experience it may be unlikely to continue. The freedom to select the relative emphasis given to current and past figures of demand by varying the time-constant enables requirements of the particular sales characteristic of each product to be met. If upward or downward trends in demand exist suitable allowance in the calculated forecast figures can be made. Also, if forecasts of more than one period ahead are of interest these can be made either specifically for a certain period in the future, or cumulatively over the periods between the present and some given date in the future.

In addition to these simple forecasts, other analogous forecasts can be made of the maximum expected demand or the minimum expected demand. These forecasts can be computed for varying levels of confidence as required. For example, it is often a requirement to calculate the maximum and minimum expected demand on a basis which ensures a 95 per cent confidence that the actual demand will in fact lie between the two figures.

The system enables any desired confidence level to be set.

The Application to Stock Control

With systems of stock control based on operational research methods a forecast of the maximum expected demand for a product during a period ahead is often required to control the level of the safety stocks. The forecasts of expected demand are therefore used as inputs for further calculation to determine order quantities, or when to place orders. The machine is flexible and can make the necessary calculations for a variety of stock control systems. It does not however, remove the need to design the control system carefully so that the best stock decision rules are formulated appropriate to each situation and the need to take account of the many theoretical and practical problems usually involved.

Definitions

EQUATIONS

The data which must be collected and used as input information consist of seven parameters. These are given shortened names, each of five letters, to provide a convenient method of meaningful reference, and also a symbol for use in equations when these are written in algebraic form.

The term 'NEW' in the shortened names infers a parameter associated with the 'present' review period, whereas the term 'OLD' refers to the values at the preceding review.

The seven inputs are:

X_1, NEWDE, the new actual value of demand in the most recent review period

X_2, OLDAV, the old average value of demand

X_3, OLDTR, the old trend in demand

X_4, OLDEV, the old expected value of demand, that is the previous forecast

X_5, OLMAD, the old mean absolute deviation of forecast errors

X_6, ORDUE, the orders outstanding and due into stock during the lead time

X_7, STIND, the stock in hand.

The solution of the equations gives twelve outputs. Not all of these outputs will necessarily be used in any given circumstances, but their availability gives considerable flexibility to the system.

The outputs are as follows:

0_1, NEWAV, the new average value of demand

0_2, NEWTR, the new trend in demand

0_3, LEPEV, the expected value of demand L periods ahead

0_4, NEMAD, the new mean absolute deviation of forecast errors

0_5, MAXEV, the maximum expected value of demand L periods ahead

0_6, CUMEV, the cumulative expected value of demand through L periods.

0_7, CUMAX, the cumulative maximum expected value of demand through L periods

0_8, STORE, the stock required, equivalent to expected demand plus safety stock

0_9, ORDER, the order to be placed

0_{10}, NEPEV, the expected value of demand one period ahead

0_{11}, MINEV, the minimum expected value of demand L periods ahead

0_{12}, CUMIN, the cumulative minimum expected value of demand through L periods.

While it would at first appear that the inputs X_1-X_7 are all new data, with a system in operation

X_2, (OLDAV) was 0_1, (NEWAV) in the previous review
X_3, (OLDTR) was 0_2, (NEWTR) ,, ,, ,, ,,
X_4, (OLDEV) was 0_{10}, (NEPEV) ,, ,, ,, ,,
X_5, (OLMAD) was 0_4, (NEMAD) ,, ,, ,, ,,

and these outputs 0_1, 0_2, 0_{10}, and 0_4, while they are available to give immediate information, are carried through to become inputs at the time of the next review period. It will be seen that only one new item of information, namely X_1, the new demand, is required at each review. Since the outputs would normally be recorded on some form of stock control card, the labour involved in manually recording the information is kept to a minimum.

Five other factors appear in the equations as pre-set values, namely

α, the smoothing constant

L, the lead time

k_1, k_2, k_3, multiples of 0_4, NEMAD

These factors have been called the 'management controls.' The value of α, the smoothing constant, is chosen to suit the particular sales characteristic of the product under consideration, and affects the relative weight of the past and present figures. The lead time, L, is the number of review periods required between placing an order for stock and the items being available, and this obviously varies with the type of product. The three k's, being multiples of the mean absolute deviation of the forecast errors, can be chosen to give any selected degree of confidence for the maximum and minimum values.

The confidence levels are set as follows:

MAXEV and MINEV with k_1

CUMAX and CUMIN with k_2
STORE with k_3

The statistical argument leading to the derivation of the calculations to be made will not be discussed here since it is outside the scope of this article and is simply a development of what has been described before[1].

The equations are as follows:

NEWAV $= \alpha$ NEWDE $+ (1-\alpha)$ OLDAV	$O_1 = \alpha X_1 + (1-\alpha)X_2$ ...... (1)
NEWTR $= \alpha$ (NEWAV-OLDAV) $+ (1-\alpha)$ OLDTR	$O_2 = \alpha(O_1 - X_2) + (1-\alpha)X_3$ .. (2)
LEPEV $=$ NEWAV $+ \left(\frac{1-\alpha}{\alpha} + L\right)$ NEWTR	$O_3 = O_1 + \left(\frac{1-\alpha}{\alpha} + L\right)O_2$ .... (3)
NEMAD $= \alpha$ |NEWDE-OLDEV| $+ (1-\alpha)$ OLMAD	$O_4 = \alpha\lvert X_1 - X_4\rvert + (1-\alpha)X_5$ .. (4)
MAXEV $=$ LEPEV $+ k_1$ NEMAD	$O_5 = O_3 + k_1 O_4$ .... (5)
CUMEV $=$ L. LEPEV $- \frac{L(L-1)}{2}$ NEWTR	$O_6 = LO_3 - \frac{L(L-1)}{2} O_2$ .... (6)
CUMAX $=$ CUMEV $+ k_2$ NEMAD	$O_7 = O_6 + k_2 O_4$ .... (7)
STORE $=$ CUMEV $+ k_3$ NEMAD	$O_8 = O_6 + k_3 O_4$ .... (8)
ORDER $=$ STORE $-$ ORDUE $-$ STIND	$O_9 = O_8 - X_6 - X_7$ .... (9)
NEPEV $=$ NEWAV $+ \frac{1}{\alpha}$ NEWTR	$O_{10} = O_1 + 1/\alpha\, O_2$ .... (10)
MINEV $=$ LEPEV $- k_1$ NEMAD	$O_{11} = O_3 - k_1 O_1$ .... (11)
CUMIN $=$ CUMEV $- k_2$ NEMAD	$O_{12} = O_6 - k_2 O_4$ .... (12)

TRANSLATION OF THE EQUATIONS INTO ELECTRICAL TERMS

To carry out the calculations electrically the inputs (X's) are produced as voltages, and the multiplying factors such as α, L and the k's are given by transformer ratios. These last being numerical, the outputs (O's) are given as voltages.

VOLTAGE SOURCE

The source for all the voltages to give the inputs, X, consists of a 1000c/s oscillator with transformer output. This transformer has several secondary windings and is designed in such a way that the ratio error between these secondaries is much smaller than the acceptable error in the calculations. The technique of the design of such transformers is well known in the field of transformer ratio arm bridges, and the ratio errors are easily kept below 0·01 per cent, and for the purposes of all the following calculations can be ignored.

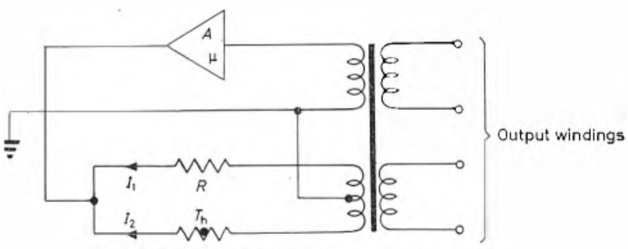
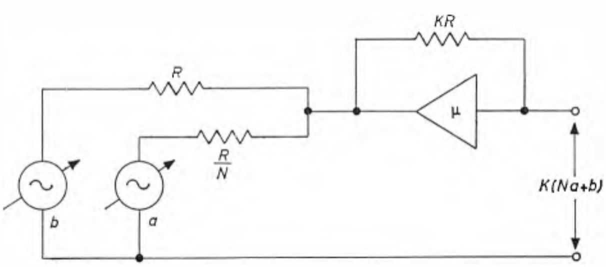


Fig. 1. Schematic diagram of source

Fig. 2. Computer block using resistive standards



One secondary winding is used to give opposing voltages for a thermistor bridge, used in the feedback path of the oscillator. This provides a pure, stable output. The arrangement of the source is shown in Fig. 1.

The transformer windings are arranged such that the current I_1 through R reinforces the input to the amplifier A. This positive feedback produces oscillation in the circuit at a frequency determined by a tuned circuit in the forward path of the amplifier. At low levels the voltage applied to the thermistor Th is small, and there is little power dissipated. In this way the resistance of the thermistor is high, and the current I_2 which provides negative feedback is small. As the level increases more power is applied to the thermistor and its resistance falls, increasing I_2. This increase of negative feedback tends to cancel I_1 and the system stabilizes at a level where I_1 and I_2 are very nearly equal, if the loop gain, μ is sufficiently high. If the thermistor bridge winding is symmetrical the stabilization occurs with the thermistor resistance very nearly equal to R.

The two main requirements are that μ should be high enough to ensure stability, and the natural limiting level of the amplifier should be sufficiently above the running level to provide adequate power in hand to stabilize the thermistor initially.

Formation of Equations

GENERAL

Examination of the equations shows that certain basic operations need to be carried out.

These operations consist generally of the addition or subtraction of a number of voltage sources, each of which may be operated upon by some preset multiplier.

The resultant voltage may be taken directly as an output, or used in conjunction with other input voltages, and through further preset multiplication processes to provide later outputs. One method for the accurate addition and subtraction of voltages which is widely used employs operational amplifiers to sum the currents through a number of resistive paths. Multiplication can be effected by a suitable choice of values of the input and feedback resistors.

For example, if the requirement is $K(Na + b)$, with a and b as input voltages and K and N as preset coefficients this can be achieved by the circuit of Fig. 2.

Provided the gain μ is high enough for the input voltage to be ignored in the presence of a, b, and the output, the accuracy is determined by the resistors.

If a variety of preset values for N and K are required the number of resistors required can become large, and if high accuracy is required at the same time, the cost can be considerable.

In a typical case N may have 10 preset values and K

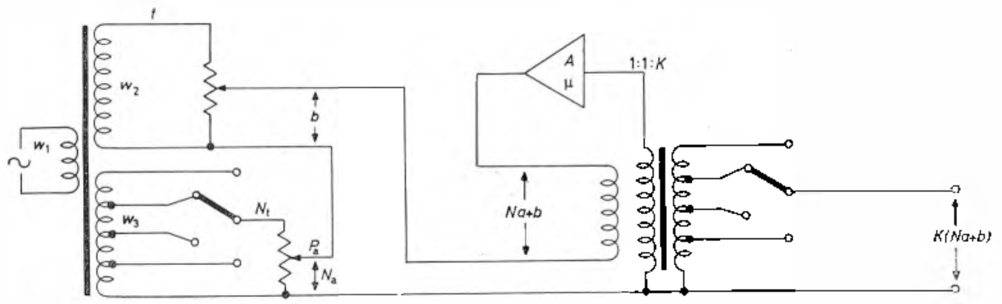


Fig. 3. Computer block using transformers

may have as many as 20. If an accuracy of 0·1 per cent is required, which is also quite typical, the circuit requires 31 resistors of high accuracy. The cost of this is considerable.

If on the other hand transformers are used to produce the preset coefficients, the number of values required makes very little difference to the cost, needing simply a further tap on a transformer.

The arrangement of Fig. 3 shows how transformers can be used to give the same solution in a simpler, cheaper, and more accurate way.

The input transformer is arranged to carry unit volt-

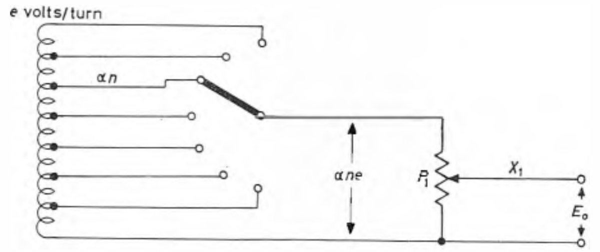


Fig. 4. Arrangement giving αX_1

age in a t turn secondary, then the input voltage to the amplifier is $(Na + b)$.

The amplifier has a transformer output, one secondary of which is used as a feedback winding effectively in series with the input. If the amplifier has sufficiently high gain for its effective input voltage to be negligible compared with its output then the feedback winding carries a voltage $(Na + b)$.

A further secondary has a turns ratio K to the feedback winding, and from this the required output $K(Na + b)$ is obtained. The accuracy of the output is determined by the loop gain of the amplifier as before, and the transformer ratio accuracies. The accuracy obtainable from transformer ratios transcends that available from resistors giving the system superiority in accuracy, and having provided a transformer, it is a fairly simple matter to arrange any number of taps to provide a wide variety of values for N and K.

Given the same gain, of say 2 000, in the feedback amplifiers, and an accuracy of 0·1 per cent for the resistors of Fig. 2, the error in the arrangement of Fig. 3 will be no greater than 0·05 per cent whereas the error in the arrangement of Fig. 2 could be as high as 0·25 per cent.

This improvement of a factor of five is brought about by the ratio accuracy of the transformers being so high that the errors introduced can be ignored, compared with the error due to the finite gain of the amplifier.

Having described the system in general the formation of particular equations will be discussed in greater detail.

Particular Equations

Equation (1) will be the first described in detail. To avoid repetition, since many of the equations use only the same principles as that of equation (1), the next equation considered will be (3) illustrating how the term $(1 - \alpha)/\alpha$ is produced. Equation (4) contains a modulus and this, being a new operation, is described in full. Finally equation (6) is described, as it contains the term $L(L - 1)/2$, the formation of which is of some interest.

Equation (1)

$$\theta_1 = \alpha X_1 + (1-\alpha)X_2$$

The output of the secondary windings of the source transformer is e volts/turn, say. The maximum, value, X_1' of X_1 corresponds with ne volts. If a secondary is arranged to have taps of αn turns selected by a switch, the output of the switch will be αne volts. This switch feeds a potentiometer P_1 as shown in Fig. 4.

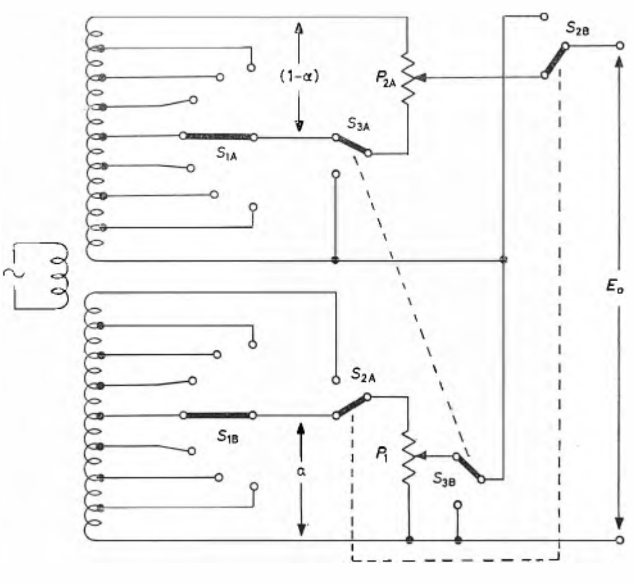
It will be seen that the output E_o is variable with the setting of P_1 from zero to $\alpha.X_1'$.

The potentiometer could carry a calibrated dial with the values of X_1, and the output $E_o = \alpha X_1$ could thus be obtained.

The second term of the equation is obtained in a similar way. A further secondary with taps $(1 - \alpha)n$ is used with a second potentiometer P_2. Again, P_2 could carry a dial calibrated with the values of X_2. This however, requires expensive calibration of dials or accurately linear potentiometers, and is unnecessary. Since the output θ_1 is to be read on some form of indicator, this indicator can be used also for setting up the values of X_1 and X_2. Fig. 5 shows the system employed.

The value of α is selected by S_1. Operation of S_2 applies a voltage to the potentiometer P_1 equivalent to an α of

Fig. 5. Method for setting inputs



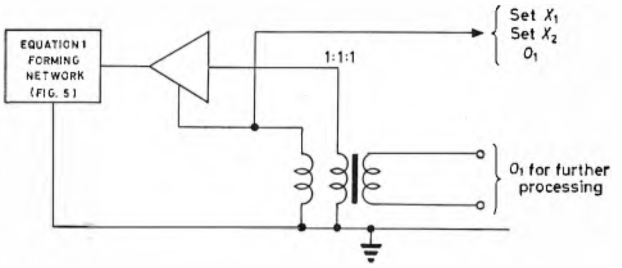


Fig. 6. Output arrangement for equation (1)

unity by S_{2A}, and connects the output terminal to the wiper of P_1 by way of S_{2B}. P_1 can now be set to give an output equivalent to X_1. Resetting S_2 and operating S_3 applies the standard voltage to the potentiometer P_{2A} by way of S_{3A}, and S_{3B} earths the bottom end of P_2. P_2 can now be set to give an output equivalent to X_2. Resetting S_3, the output becomes

$$\alpha X_1 + (1 - \alpha)X_2 = 0_1$$

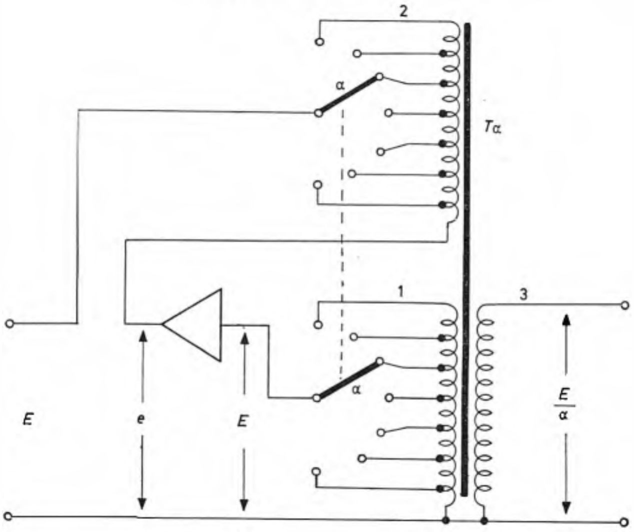
By using this system of setting-in the values of X_1 and X_2 using the same indicator as is used for the outputs from the calculations, the need for either calibrated or linear potentiometers is eliminated.

In practice the values for X_1 and X_2 lie in the range 0 to 1 000 units and α has values 0, 0·05, 0·1 and in steps of 0·1 to unity. If the transformer has an output of 25mV/turn, then a 40 turn secondary would give an output of 1V. Taps of 2t, 4t, 8t, 12t. . . 36t, 40t give the required values of α. The values of X_1 and X_2 are set in directly in millivolts, and the output 0_1 is also given directly in millivolts.

Since the output 0_1 is required in subsequent equations for further processing, to remove any load from the potentiometers P_1 and P_{2A}, and also to provide 0_1 free from earth, the output E_o is taken to a unity gain amplifier with transformer output. A high gain amplifier feeds an output transformer, one winding of which provides feedback. If the gain of the amplifier is sufficiently high, the deviation from unity gain can be made small enough to have no significant effect on the accuracy of the calculations. Even this error is eliminated if this amplifier is included in the setting up process.

This is shown in Fig. 6.

Fig. 7. Dividing by α



By arranging the switches S_2 and S_3 to co-operate with the input switches to the indicator the voltages corresponding to the two inputs X_1 and X_2 and the output 0_1 can all be obtained from the feedback winding of the amplifier output transformer. A further secondary winding provides 0_1 free from earth for further processing.

Fig. 6 is schematic, and in practice the first stage of the amplifier consists of a transistor stage with the input taken to the base and the feedback winding in the emitter circuit. In this way the collector current is proportional to the difference between the input and feedback voltage. This collector current is taken through further stages of gain to provide the magnetizing current for the transformer.

Such an amplifier can typically operate with a base-emitter voltage of less than 0·1 per cent of its output voltage, and if the output and feedback windings have the same number of turns the error between input and

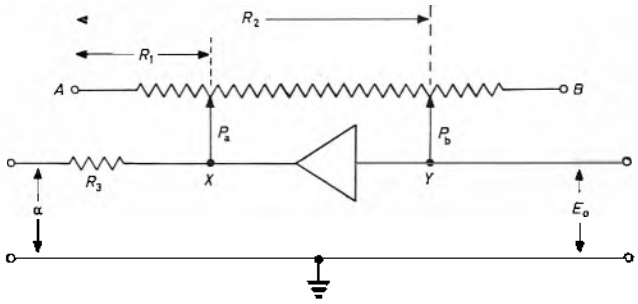


Fig. 8. Production of $|X_1 - X_2|$

output is less than 0·1 per cent. This error is removed from 0_1 however, by its inclusion in the setting of X_1 and X_2, as shown in Fig. 6.

In subsequent operations it is not always possible to eliminate the finite gain error in this way, but where such a technique is possible it can be exploited to advantage.

EQUATION (3)

$$0_3 = 0_1 + \left(\frac{1-\alpha}{\alpha} + L\right) 0_2$$

Only the part of equation (3) involving the division by α will be considered, as the rest is simple addition of voltages already produced in previous processes. Consider the circuit of Fig. 7.

Each winding of the transformer T_α has n turns. The magnetizing winding (1), and the feedback winding (2), are both tapped, each having taps αn selected by the ganged switch. The amplifier has high gain, so that its input voltage e is negligible compared with its output voltage, and will be ignored. It will be seen that by placing the feedback winding (2) in series with the input, the output developed across αn turns of winding (1) is E. It follows that the output across winding (3) is E/α.

The way the circuit is described infers that α is less than unity, as is the case in practice. For an α greater than unity it will be seen that the system operates equally well, the only difference being in the ratio between windings (1) and (3) of the transformer.

As in the first equation, the output of equation (2) is provided by a transformer. This transformer is tapped to give $(1 - \alpha)0_2$. This is used as the input to the circuit of Fig. 7. In this way the output is $\left(\frac{1-\alpha}{\alpha}\right) 0_2$.

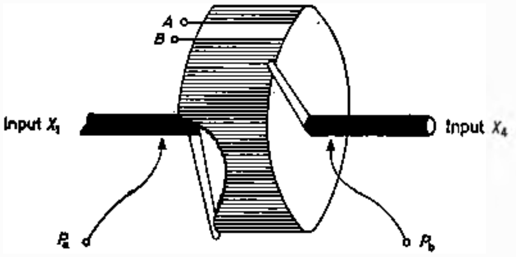


Fig. 9. Construction of potentiometer for modulus extraction

The output of equation (2) is provided with a further output winding tapped with the appropriate values of L to give $L\theta_2$.

This is added, in conjunction with an output from equation (1), giving θ_1, to the output of the circuit of Fig. 7.

In this way the total output amounts to

$$\theta_1 + L\theta_2 + \left(\frac{1-\alpha}{\alpha}\right)\theta_2$$

which gives

$$\theta_1 + \left(\frac{1-\alpha}{\alpha} + L\right)\theta_2$$

which was the requirement.

EQUATION (4)

$$\theta_4 = \alpha|X_1 - X_4| + (1-\alpha)X_5$$

Only the part of the equation involving the modulus will be considered in detail, as that part containing X_5 simply involves the process of addition. The modulus $|X_1 - X_4|$ demands that the result of this difference shall be positive, regardless of the relative superiority of either X_1 or X_4. A simple method for achieving this result with X_1 and X_4 as voltages was not apparent so a different approach was used.

Consider the arrangement of Fig. 8.

AB is a linear potentiometer carrying two independent wipers P_a and P_b. This is achieved by constructing a conventional winding on a cylinder, and having the two wipers P_a and P_b at opposite ends of the cylinder, as shown in Fig. 9.

If the resistance between end A and P_a is R_1 and the resistance between end B and P_b is R_2, the resistance between point X and Y is $|R_1 - R_2|$. This resistance is obviously always positive regardless of the relative magnitude of R_1 and R_2. Assuming the gain of the amplifier is sufficiently high for its input voltage to be regarded as negligible, the output E_o, with an input of α as shown, will be

$$E_o = \alpha\frac{|R_1 - R_2|}{R}$$

Thus if R_1/R is made equal to X_1 (and this is achieved by ganging P_a to P_1 in equation (1)) and R_2/R is made equal to X_4.

$$E_o = \alpha|X_1 - X_4|$$

X_4 can be set by switching the input voltage to unity, cancelling the α setting, and switching the wiper P_a to point A. This leaves R_2 as the feedback element, and enables X_4 to be set-up at the output of the amplifier. It is a simple matter to add $(1-\alpha)X_5$ to this output to provide θ_4.

EQUATION (6)

$$\theta_6 = L\theta_3 - \frac{L(L-1)}{2}\theta_2$$

The interest in this equation lies in the fact that several outputs are required from the θ_2 process. One is $L\theta_2$, used in equation (3), another is $\frac{L(L-1)}{2}\theta_2$ required in this equation. The transformer output of θ_2 has therefore to carry several secondaries, and these secondaries must have such a number of turns that taps can be provided to satisfy all requirements of α and L.

It was found that the simplest way instrumentally of forming this calculation was to take

$$\theta_2 + \left(\frac{1-\alpha}{\alpha}\right)\theta_2 + \left(\frac{L+1}{2}\right)\theta_3$$

The first term was obviously available, the second term was also available having been employed in equation (3) and the third term was available by using an extra winding on the θ_2 transformer.

This combination gives

$$\theta_3 = L\theta_2 + \left(\frac{L+1}{2}\right)\theta_2 = \theta_3 - \left(\frac{L+1}{2}\right)\theta_2.$$

Straightforward multiplication by L throughout then gives the required result. It will be seen then that the θ_2 output transformer has to provide three outputs, namely

(a) $(1-\alpha)\theta_2$

(b) $L\theta_2$

(c) $\left(\frac{L+1}{2}\right)\theta_2$

The range of values for α being 0·05, 0·1 by 0·1 to unity give the simplest configuration of a winding as 20 turns. The range of values for L are 0, 0·5,, 1·0, by 0·2 to 3·0, 3·5, 4·0 by 1·0 to 12. Having set the value of 20t for unity α, the values of $L\theta_2$ are thus given by 10, 20, 24, 28 etc. to 240 turns. It will be seen that

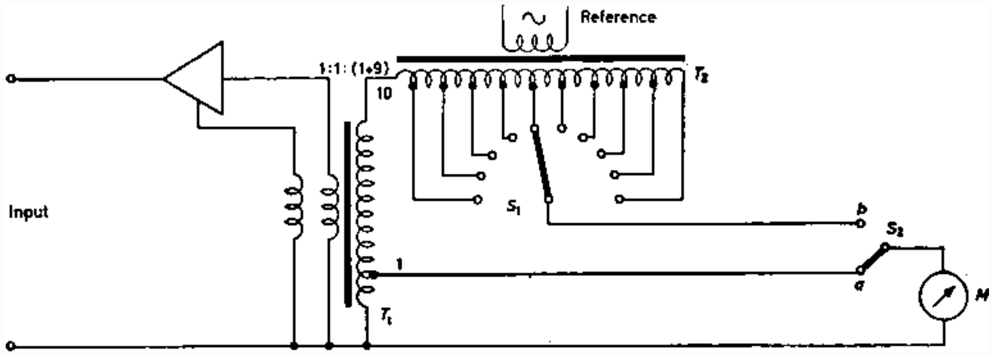
$$\frac{L+1}{2}\theta_2$$

is given by

0, 15, 20, 22, 24 etc. to 130 turns.

Thus a fairly simple transformer can provide quite a variety of output arrangements. In fact provided coefficients have reasonable numerical values quite simple transformers can give outputs involving first, second and third orders of a coefficient.

Fig. 10. System for enhanced reading accuracy



Indicators

In the setting up of the input data and in the extraction of the outputs some form of indicator is required. This could take the form of a simple voltmeter with a moving-coil indicator if only relatively low accuracy was required.

It will be seen that in some conditions the errors introduced in the output voltages by errors in setting the input data are considerably enhanced by the composite coefficients which appear in the formation of the equations. In some instances an output is the result of multiplying the difference between two large input quantities by a large factor. To give the required accuracy an input accuracy of better than 0·1 per cent f.s.d. is required, and a simple moving-coil voltmeter does not meet this requirement. A system similar to that employed with the self-balancing transformer bridge was adopted[2] giving first a quick reading, and secondly an accuracy to meet the need. The voltage to be measured is fed to a phase sensitive detector which, with a moving-coil meter, gives a two digit reading. The first digit of this reading is transferred to a decade system which performs two functions. The first is to subtract from the voltage to be measured an accurate reference, derived from the energizing voltage source, equivalent to the appropriate digit. The second function is to increase the sensitivity of the indicator by a factor 10.

The system is shown in Fig. 10.

The input is taken to a unity gain amplifier with a transformer output similar to those used in the formation of the equations.

The secondary of the transformer has two outputs giving an accurate gain of either 1 or 10.

Transformer T_2 is in fact a winding on the voltage source transformer. This provides accurate references in unit steps from zero to 10. To obtain a quick result S_1 is set so that none of this winding is in circuit. S_2 co-operates with S_1 in such a way that under these conditions S_2 is in position *a*. The meter *M* reads from zero to its full scale deflexion, 1. If the first digit is taken from the meter reading and S_1 operated to bring in the appropriate digit of the reference on T_2, S_2 changes automatically to position *b*. The signal now fed to the meter consists of the original signal multiplied by 10, from which has been subtracted the reference equivalent to the figure in the first place. This leaves the two subsequent significant figures to be read by the meter.

To take an example, suppose the input voltage is 0·643V, where the full scale deflexion is 1V. The reference decade selected by S_1 is initially inoperative and S_2 is in position *a*. The meter will have been calibrated to give a full scale indication for 1V, and thus indicates 0·64—the third digit being lost in the discrimination and accuracy of the meter.

If now the digit 6 is selected on S_1 the output is 6·43V on the 10 : 1 ratio of the transformer, and the reference subtracts 6V from this output. This leaves 0·43V to be fed to the meter, since S_2 has moved to position *b* by operation of the decade switch S_1. The decade dial indicates 6, and the meter indicates 0·43, giving the necessary three significant figures.

It will be seen that the meter can very conveniently be calibrated by a unit winding of T_2. This removes any systematic error from non-linearity in the meter and metering circuits.

Switching

The object of designing the computer was to speed the calculation of the various equations, and so permit a large number of stock lines to be reviewed in as short a time as possible. Operational ease, therefore, is a feature which requires a considerable amount of consideration.

For some applications not all the outputs may be required; perhaps a simple forecast is all that is needed. It is therefore desirable that any input or output can be selected at will, easily and quickly.

Push-button operation was chosen to provide the necessary speed and facility, the push-buttons operating micro-switches which in turn actuate relays. This makes for light operation and eases the task of the operator who may need to make 500 reviews per day, and also allows the switching to be designed for long life and reliability.

An input is selected by operating a button with the appropriate legend, and can be set with the potentiometer immediately below the button.

Selection of another input, or an output, automatically cancels the previous selection. The first digit of the indicated value can be transferred to the decade by simply pushing the appropriate button, leaving the two subsequent figures to be read on the indicator. Operation of another data button automatically cancels the decade, and resets the indicator to read the total quantity fed in.

Since the trends may be of either sign a button is provided for negative readings, but to avoid confusion on selection of a new data button the machine automatically resets to positive reading.

Accuracy

It is difficult to generalize on the subject of accuracy of the outputs since this depends so much upon the value of the constants α and *L* etc.

Each particular set of values has associated with it its own accuracy.

Four cases were taken in an attempt to drive the effect of errors to extremes. The equations were worked manually to give precise figures for the output data. Each example was used in the computer a number of times using several operators of widely different experience, with remarkably consistent results between operators.

The standard deviation for each set was calculated and it was found that in general the systematic error—that is the difference between the mean value obtained in practice and the calculated value—was less than one standard deviation.

The standard deviations obtained were referred back to the input and showed that a likely value for the general error is equivalent to errors in the input quantities of 0·05 per cent of the full scale values. Since the discrimination with which the inputs could be set was about 0·05 per cent, and the systematic error appeared to be no more than this figure, the system would seem to be satisfactory in practice.

Acknowledgment

The design of the system has been carried out in collaboration with Associated Industrial Consultants Ltd, and is the subject of British and foreign Patent Applications.

The authors wish to thank The Wayne Kerr Laboratories Ltd for permission to publish this article.

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The Meander Line Aerial

By T. S. M. Maclean*

A new form of endfire, travelling-wave aerial is described, in the form of a meander line. This consists of a conductor folded into elements each slightly longer than one half wavelength, using only right-angled bends. The endfire radiation exists over a bandwidth of approximately 12 per cent with a sidelobe level better than 6dB.

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

AN interesting example of two apparently different branches of electrical engineering where common structures have been used is that of microwave valves and radio aerials. Thus the concept of the helical aerial in both its circularly[1] and linearly[2,3] polarized forms was preceded by the invention of the travelling-wave tube using such helices. The reason for this is quite clear since both devices aim to produce electromagnetic waves which travel along the axis of the structure with a velocity less than that of light. However, it must not be thought that a direct transfer can be made between a unit designed for a microwave valve with the expectation of it performing optimally as a radio aerial. Two main differences arise. One is that the desired velocity in the valve case may be only one-tenth of that of light, whereas in the aerial case it will be approximately nine-tenths of the velocity of light. The other is that the aerial structure will be much larger than the valve one for good radiation at the same frequency. Nevertheless since the transfer has been successfully made in the cases cited above, consideration has been given to applying the meander line of valve usage as a radio aerial. This consists of a line folded in rectangular fashion as shown in Fig. 1, and can be expected to operate in the same manner as the zigzag aerial[4]. No easy prediction can be made theoretically of the bandwidth over which the aerial would be expected to work, but this has been examined experimentally and found to be approximately ± 6 per cent of its centre frequency of operation. Mechanically it offers some advantage in ease of construction over the zigzag aerial.

Simplified Theory of Operation

The coaxial cable feeding the meander aerial can be expected to launch a travelling wave of current along the structure. At the far end of the aerial this current must clearly fall to zero, and a reflected wave will thus be set up. As in the case of all other singly fed, endfire aerials, such as the Yagi or dielectric rod, this reflected wave is of small amplitude, and it is approximately true to say that radiation is predominantly due to the forward travelling wave alone.

Considering a single unit in the aerial as shown in Fig. 2, and replacing it by a pair of point sources, as shown, radiation will be in the endfire direction when the phase difference between these sources, at a distant receiving point, is:

$$\psi = \beta_1 (L + D) - \beta_2 D - \pi \quad \ldots\ldots\ldots\ldots \quad (1)$$

$$= 0^\circ \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

where the π arises because of the opposite directions of current flow in the two wires of the element. The phase constants β_1 and β_2 take account of the fact that the phase velocities along the wires and axially are not necessarily equal to the free space phase constant k. Then from equations (1) and (2):

$$L = \frac{\pi - (\beta_1 - \beta_2)D}{\beta_1} \quad \ldots\ldots\ldots\ldots \quad (3)$$

For $\beta_1 = \beta_2 = k$, this would give $L = \lambda_o/2$, where λ_o is the free-space wavelength. But though β_1, the phase constant for the wave along the wires may be assumed, for

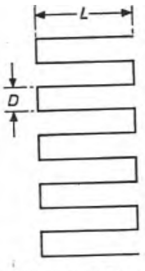
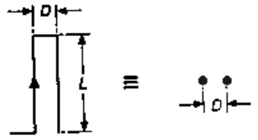
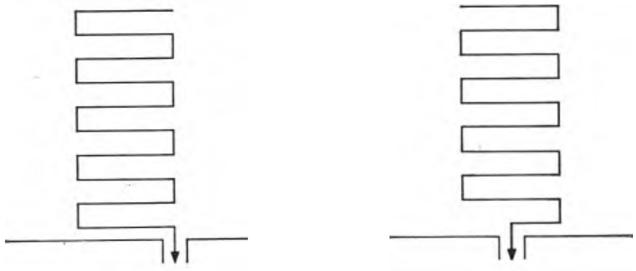


Fig. 1. Meander line aerial



(a) (b)

Fig. 2. (a) Single unit of meander line
(b) Pair of point sources equivalent to single unit



(a) (b)

Fig. 3. (a) End-fed meander aerial
(b) Centre-fed meander aerial

a TM mode, to be approximately k, this is not expected to be true for β_2. In this case it would be expected that the velocity axially would be less than the velocity of light, since the array of the elements L acts effectively as an artificial dielectric, as in the Yagi aerial[5]. Then from equation (3) L would be greater than $\lambda_o/2$. It should be noted, however, that this is just the opposite of the length of the director elements in a Yagi aerial, where in the absence of a conductive connexion between the elements, lengths greater than $\lambda_o/2$ would prevent operation of the aerial in its normal form. This length $L > \lambda_o/2$ is, however, found with both the actual and transverse projected lengths of the zigzag aerial.

Inspection of equation (3) suggests that the dimension D, the spacing between the arms of the radiator, may have only a minor influence in affecting the length L of these arms, since $(\beta_1 - \beta_2)$ is small. Even if D became relatively large, the axial phase velocity would be expected to increase, and thus $(\beta_1 - \beta_2)$ would decrease, tending to keep the product $(\beta_1 - \beta_2)D$ approximately constant. Were this simple relation true it would allow a great

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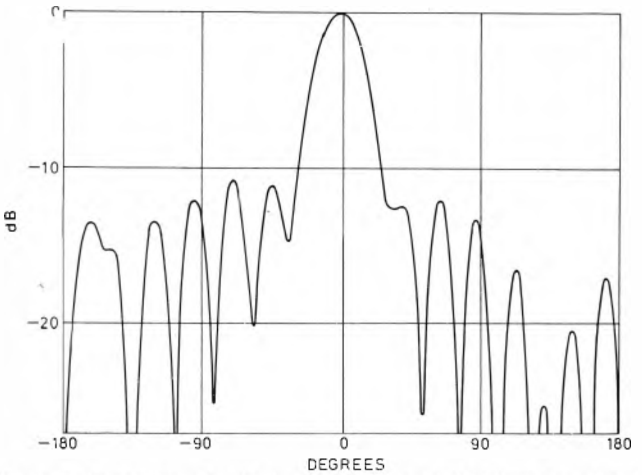


Fig. 4. E-plane pattern of end-fed meander aerial at 9.0kMc/s (L/λ_0 = 0·541)

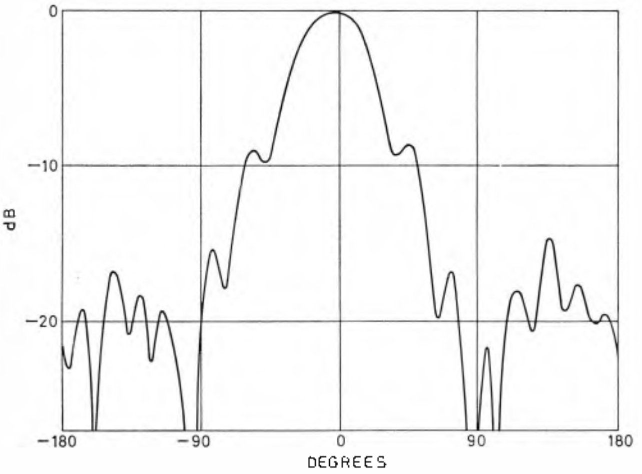


Fig. 5. H-plane pattern of end-fed meander aerial at 9.0kMc/s (L/λ_0 = 0·541)

measure of flexibility in the construction of the aerial. Clearly, however, D could not be allowed to increase so as to cause radiation at $\pm 90°$ equal to that in the endfire direction. Nevertheless use was made of this assumption in the first aerial tested.

Once the pattern for the unit element of Fig. 2 has been found, the total pattern for the aerial is readily calculated from the array factor for n such elements $\dfrac{\sin n\,(\psi/2)}{n \sin(\psi/2)}$, where ψ is now twice the value given by equation (1).

Construction of the Aerial

The aerial testing equipment available was in the 3cm band and the first model of length 6 unit sections was constructed as shown in Fig. 3(a) with the lengths L and D equal to 1·5 and 0·75cm respectively. The ratio L/D was deliberately kept small since it simplified the mechanical construction of the aerial, and the simplified considerations of the previous section suggested that the length D might not be of great importance. An alternative method of centre feeding the aerial as shown in Fig. 3(b) was also used. In both cases the return current of the aerial flows along a square ground plane of side 4cm, back to the outer of the coaxial N-type connector.

For reasons which are referred to later it became necessary to modify the ratio L/D and a new value for this of 4·5 was adopted. In this case also the length L was increased to 1·8cm, but this change was made for the purely practical reason of operating over a more sensitive range of the equipment. The overall axial length in this case was 8cm, or approximately 2·5 wavelengths.

Experimental Results

Radiation Patterns

(a) $L/D = 2$

The usefulness of any new travelling-wave endfire aerial can be judged by comparing its power gain or radiation pattern with the corresponding values for already existing aerials, such as the Yagi or zigzag. In both these cases the radiation pattern consists of a main central lobe with first sidelobe level approximately 8 to 9dB below the main lobe maximum. This general behaviour exists over a frequency range ± 5 per cent from the centre frequency. As judged by these criteria the meander aerial with L/D equal to 2 is not good. The best pattern had a sidelobe level only 7dB below the main lobe and this fell rapidly with change of frequency. This relatively poor performance applied to both the centre-fed and end-fed aerials. Consequently a new model was constructed with L/D equal to 4·5.

(b) $L/D = 4{\cdot}5$

In this case there was a marked difference between the centre-fed and end-fed aerials. When the aerial was centre-fed, as in Fig. 3(b), a marked asymmetry in the level of the sidelobes on each of the main beams existed. With the end-fed aerial, however, as in Fig. 3(a) considerable

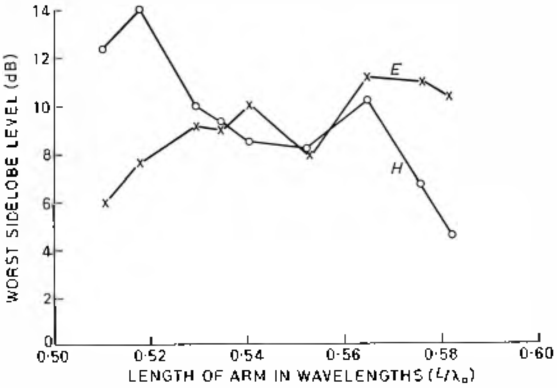
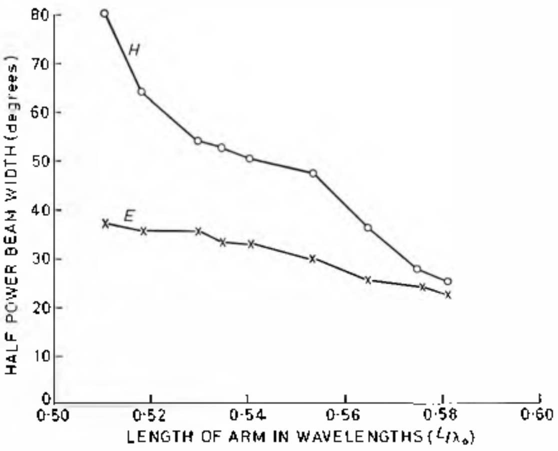


Fig. 6. Worst sidelobe level as a function of frequency for end-fed meander aerial

Fig. 7. Half power beamwidth as a function of frequency for end-fed meander aerial



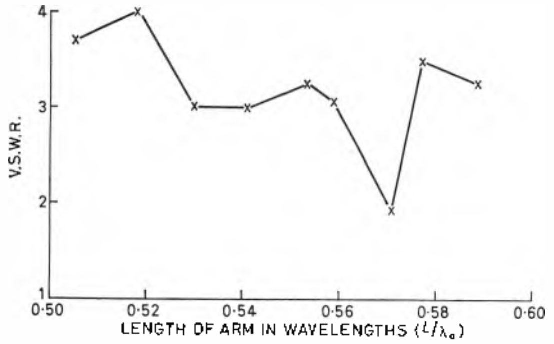


Fig. 8. V.S.W.R. of end-fed meander aerial on 50Ω line as a function of frequency

improvement in radiation pattern was evident, and typical results for the E and H planes are shown in Figs. 4 and 5 at a frequency of 9·0Gc/s. If a usable frequency range for the aerial is defined as the range for which the worst sidelobe is at least 6dB below the main beam, then over this range the variation of worst sidelobe level and also of half power bandwidth are as shown in Figs. 6 and 7. With this value of L/D, therefore, and with end feeding, the aerial compares well with both the Yagi and zigzag aerials.

Standing Wave Measurements

The voltage standing-wave ratio of an aerial as measured on the transmission line feeding it is in general not the determining factor in ascribing a bandwidth to a travelling-wave aerial. Nevertheless a large v.s.w.r. does reduce the power gain available and consequently it is necessary to find the variation of this on a typical transmission line. In this case a line with a characteristic impedance of 50Ω was used, and the variation of v.s.w.r. with frequency is shown in Fig. 8. As would be expected from physical considerations this follows closely the corresponding curve for the simple zigzag aerial[6].

Conclusions

The meander aerial has been shown to be an acceptable alternative to the Yagi and zigzag aerials. It offers a slightly better bandwidth (12 per cent) than the Yagi (10 per cent) and a greater ease of construction than the zigzag, particularly for large conductor diameters.

Acknowledgments

The author wishes to acknowledge conversations with A. E. Siegman of Stanford University which led to the investigation of this aerial.

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Numerically Controlled Milling Machines

The Maxemill range of tape controlled milling machines for machining all types of material including high tensile steel, cast iron, and light alloys, are among the first machines of this type that have been designed explicitly for numerical control.

There are two basic machines in the range: a bridge type machine carrying a twin vertical head, and a vertical column machine with horizontal head. With each, there are three table sizes: 5ft × 3ft, 8ft × 3ft and 8ft × 5ft.

Each type will be fitted with Ferranti three axis control with eight channel punched paper tape for automatic straight line milling and co-ordinate positioning; or with magnetic tape for continuous path milling. Other types of control will be available to meet customers' requirements.

Each axis is controlled by closed loop servo systems and measurements are taken from optical grating spars, positioned near to the ball screws. Infinitely variable hydraulic servo motors driving precision gear boxes provide the motive power.

An infinitely variable hydraulic motor drives the head and three gear positions give a speed range from 20 to 4 000 rev/min. Thirty horse power is available from 36 rev/min to top speed, reducing proportionately to 20 h.p. below 36 rev/min.

The machines will cater for a maximum component weight of 4½ tons and at low feed speeds a net table thrust of more than 1½ tons is available. At higher feed speeds, up to the maximum of 100 or 120in/min, the thrust is maintained with a proportionate increase in gross power.

In the vertical Maxemill with paper tape control the cross slide spans the table and is supported on both sides of the machine by large columns containing all the electronic equipment and the controls, including the tape reader. The head includes twin spindles as standard, mounted at fixed centres, each rising and falling together under z axis control. One of these spindles can cover the whole width of the table and can be used alone. For work pieces of less than half the table width, both spindles can be used so that two workpieces can be machined simultaneously.

Cutter changes can be made between any movement or at planned stops, so that all types of operation in any of the three axes (such as milling, boring, drilling, copying and centre boring) can be covered in one programme without elaborate tooling and transfer of work.

The machine can also be operated by dial control using decade switches and call buttons. This is an accurate and economical method of producing 'one off' components and carrying out additional machining operations not catered for on the tape.

Automatic clamps on each axis operate as soon as a movement is completed and remain on until that axis is again called. Decelerations are automatically applied so that overrun is eliminated. For positioning, simultaneous movement in either x and y or x and z axes, can be supplied to meet individual customer's requirements.

With the vertical Maxemill using magnetic tape continuous path control (360° profile with depth control) the basic dimensions and layout differ little from the paper tape controlled vertical miller. The differences are in the ratio of the servo gear boxes and the absence of axis clamps. The machine is continuously under servo control in all three axes simultaneously.

The machine is provided with a retaining clamp on the z axis which operates when the hydraulic supply to that axis is switched off preventing the head from lowering under its own weight.

Once initial settings have been made, the machine requires virtually no attention for reproducing components in two dimensions to any defined shape and in three dimensions, providing that the programming is an economical proposition using feed speeds and rapid traverses predetermined by the planner.

The operator sets the spindle speed according to planner's instructions. Different cutters can be inserted at tape stop points pre-arranged in the programme.

On the horizontal Maxemill with paper tape control a large vertical column and slide carry the head unit. The head slide is horizontal and the open side feature of the machine allows awkward workpieces to be handled and machined.

A single horizontal spindle is fitted but in other respects the head is similar to the vertical machine. As an optional extra, a wider table with outrigger support can be fitted for milling wide components.

Normal operation is by automatic tape control of position, feeds, and machine function. The control systems available are identical to those of the vertical miller, but variations do occur in the amount of movement available on the y and z axes. The horizontal machine can be regarded as a vertical machine but with a y/z axis interchange. The same wide field of application and capability from high speed routing of light alloys to the lower speed machining of hard ferrous materials apply to both.

A Two-Transistor Stimulator for Physiological Purposes

J. Opalinski*, P. Eng.

This article describes a compact two transistor physiological stimulator for classroom application with electromagnetic coupling through a Zener diode. A separate unijunction transistor timing generator is included.

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

THE physiological stimulators used at present for student class work are in general quite obsolete. In recent years, only two articles[1,2] in *Electronic Engineering* have reported circuits using transistors for generating the single and repetitive pulses necessary for this type of stimulator. The circuit to be described is that of a simple, reliable monostable blocking oscillator, triggered by a unijunction relaxation oscillator which is used as a timing device. Using this technique, it has been possible to make a stimulator showing the following characteristics:

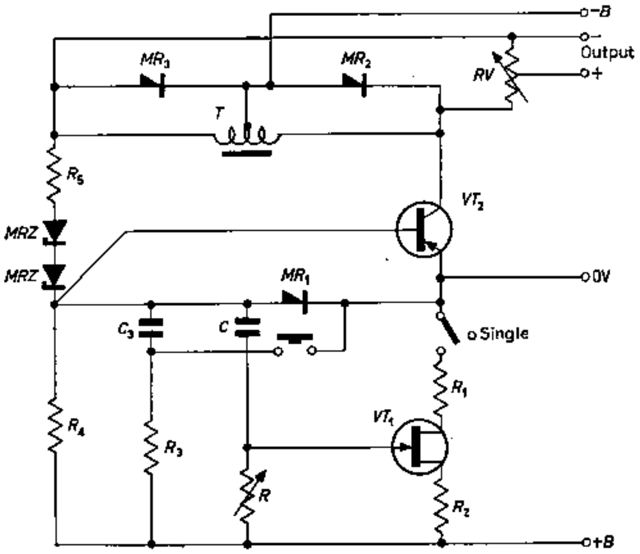


Fig. 1. Basic circuit of stimulator

Pulse duration:	Variable, or may be fixed at 1msec.
Repetition Rate:	Single pulse, or variable from less than 1c/s to over 100c/s.
Output Voltage:	Variable from 0 to 120V.
Pulse Shape:	Rectangular, with very short rise and fall time. Output unaffected by loads down to 1kΩ.

By suitable design changes, it is possible to obtain higher frequencies, longer pulse duration, or to increase the output voltage. The output power is limited only by the dissipation of the transistor which is used, and the rise and fall time depend on the transistor characteristics.

Circuits

Fig. 1 shows the circuit of the stimulator. Its operation may be explained as follows: The path $B+$, R_4, MR_1, 0V conducts a small current ($R = 47\text{k}\Omega$, $E = 15\text{V}$) which causes a voltage drop of about 0·3V across MR_1. The positive bias on the base of VT_2 thus keeps it normally cut off. The Zener path does not conduct because the Zener voltage is greater than the voltage supplied by the battery,

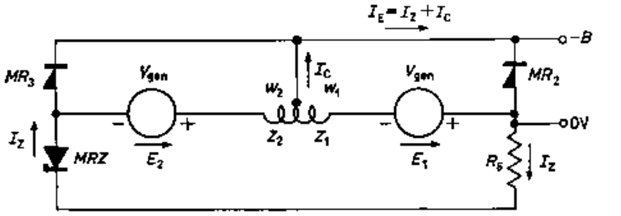


Fig. 2. Equivalent branch circuit. On time

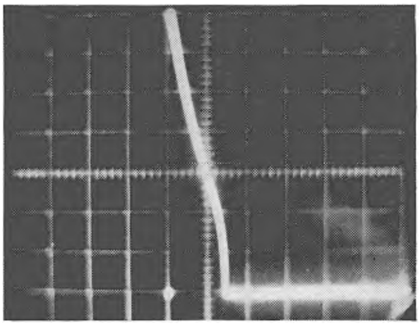
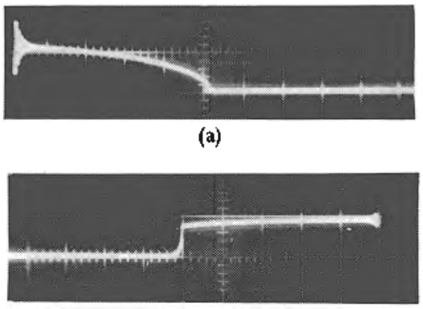


Fig. 3. Trigger pulse
Ordinate—1V/div Abscissa—10μsec/div



(a)

(b)

Fig. 4. (a) Zener current
Ordinate—10mA/div Abscissa—0·4msec/div
(b) Zener voltage
Ordinate—1V/div Abscissa—0·4msec/div

plus the voltage drop across MR_1. If now VT_2 is triggered, either by a short spike from VT_1, or from the push-button trigger, it temporarily conducts and its collector current I_c starts to flow through the transformer winding W_1. The transformer current induces a counter e.m.f., E, which opposes the rise in current. A proportional e.m.f., E_2, is induced in W_2 of the transformer.

The equivalent branch circuit is shown in Fig. 2, and

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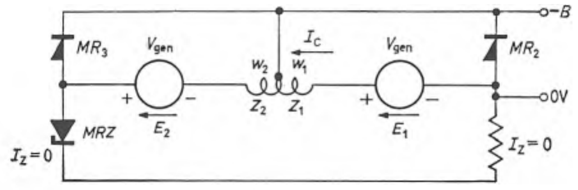


Fig. 5. Equivalent circuit. Off time

the trigger pulse and Zener current are shown in Figs. 3 and 4. The voltage on the Zener is now increased by E_2, and thus the sum of voltages on the Zener is greater than the Zener voltage, causing the Zener diode to start conducting. The Zener current cannot pass through MR_1, but rather flows through VT_2, and maintains VT_2 in a conducting state as long as the voltage is sufficiently high. No current flows through MR_3. However, the magnetic flux of the transformer cannot rise indefinitely, and after a time proportional to L/R, the time-constant of the transformer feedback path, E_2 becomes too small to keep the Zener conducting. It stops suddenly and the collapse of the transformer field induces currents, as shown in Fig. 5, which speed up the cut off of VT_2. The diode MR_1 keeps VT_2 cut off until the next trigger.

The induced voltages, E_1 and E_2, in the collapsing magnetic field in the transformer are shorted by diodes MR_2 and MR_3 and the energy is dissipated.

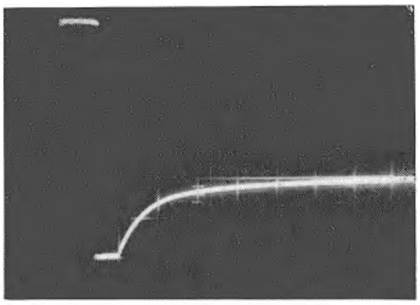
Fig. 6 shows the full swing of the Zener voltage, with the recovery of the zero line to the d.c. level.

The output pulse taken from W_1 and W_2 (the outside terminals of the transformer) is shown in Fig. 7.

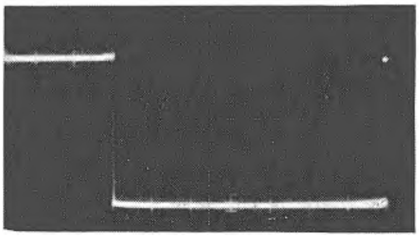
The photographs in Fig. 7(a to e) show the influence of varying the load. The top falls off slightly, because of the voltage drop due to the resistance of the transformer windings.

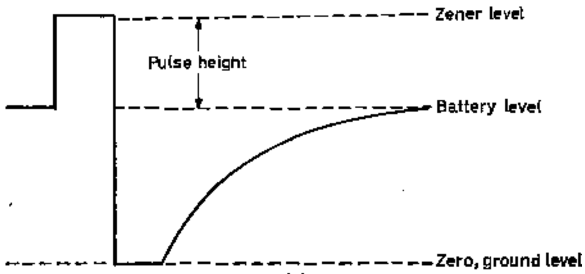
If the output voltage is taken from the centre-tap of the transformer, and the collector of VT_2, the pulse is not affected by the load, and the output voltage equals the battery voltage. The transformer does not deliver any power in this case, and it can be as small as the time-constants of the feedback path may require, i.e. L/R (Fig. 8). The leading and trailing edge of the pulse are shown

(a)



(b)





(c)

Fig. 6. Zener voltages

(a) Diode across W_1

(b) Diodes across W_1W_2

Ordinate—20V/div Abscissa—1msec/div

(c) Voltage levels for (a) and (b)

in Fig. 9. These results are obtained with the values of the components as shown in Fig. 1.

A complete description of the stimulator requires the mention of two more circuit elements. The relaxation

Fig. 7. Output pulse, from transformer

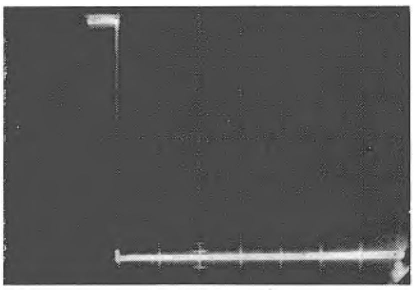
(a) Two diodes across both sections of transformer
Ordinate—20V/div
Abscissa—1msec/div

(b) As (a), only Abscissa—0·1msec/div

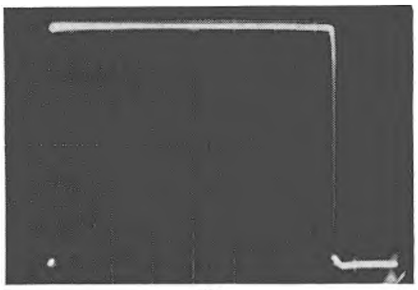
(c) One diode only
Ordinate—20V/div
Abscissa—1msec/div

(d) Output loaded by voltage divider
Ordinate—20V/div
Abscissa—1msec/div

(e) Ordinate—20V/div
Abscissa—0·1msec/div

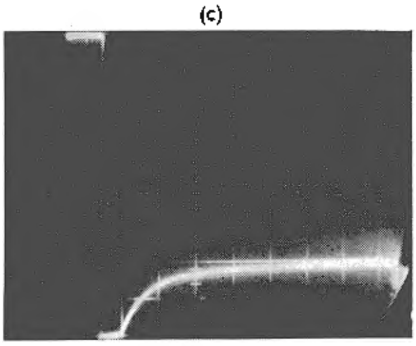


(a)

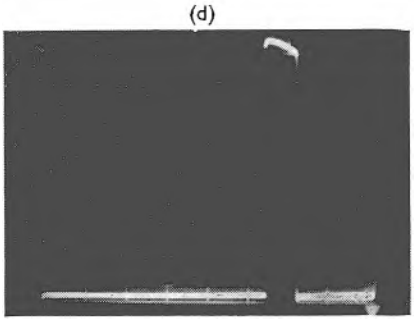


(b)

(c)



(d)



(e)

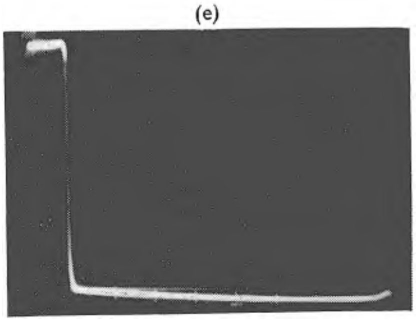
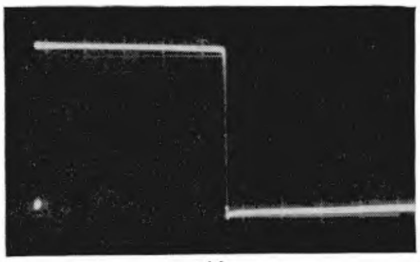


Fig. 8. Load independent output
(a) *Two diodes*
(b) *One diode*
Ordinate—10V/div
Abscissa—0·2msec/div


(a)

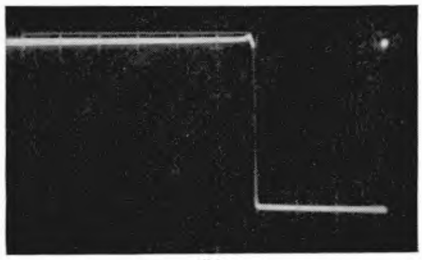
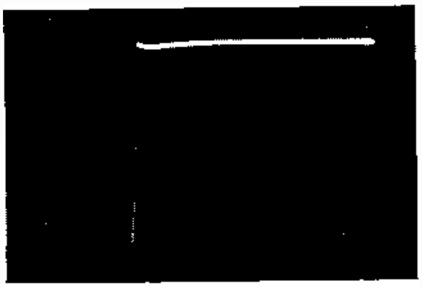

(b)

Fig. 9. Rise and fall time
(a) *Rising edge*
(b) *Trailing edge*
Ordinate—20V/div
Abscissa—0·2msec/div


(a)

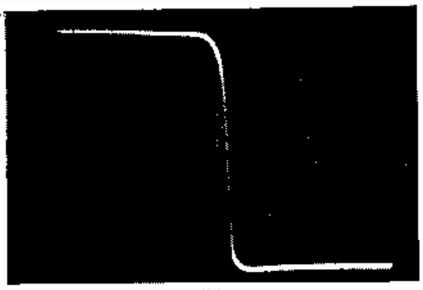
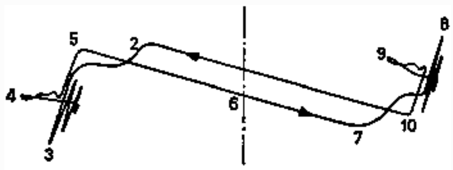
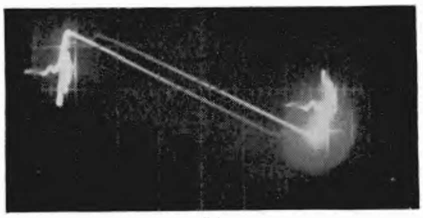

(b)

Fig. 10. Voltage-current characteristics
(a) *Schematic drawing.*
(b) *Photograph of c.r.t. display*
Ordinate—50mA/div
Abscissa—20V/div



(a)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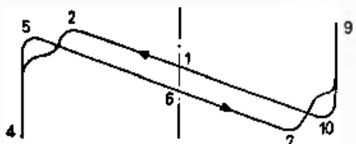


(a)


(b)

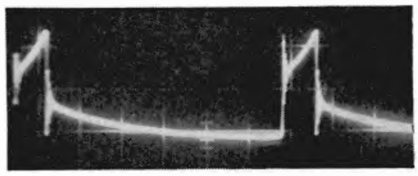
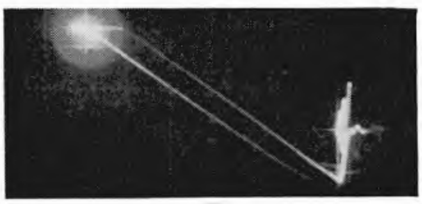
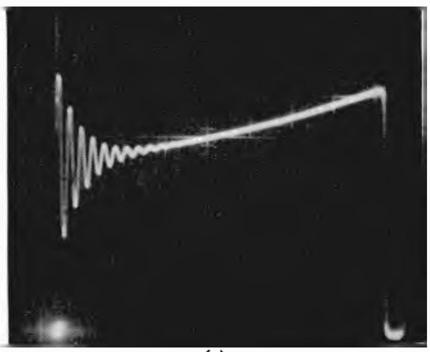

(c)

Fig. 11. Voltage/current and current/time characteristics
(a) *Schematic drawing*
(b) *Photograph of c.r.t. display*
Ordinate—50mA/div *Abscissa—20V/div*
(c) *Photograph of c.r.t. display*
Ordinate—50mA/div *Abscissa—1msec/div*


(a)


(b)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(d)

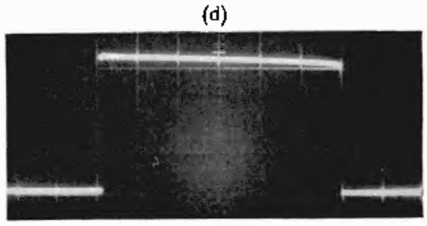
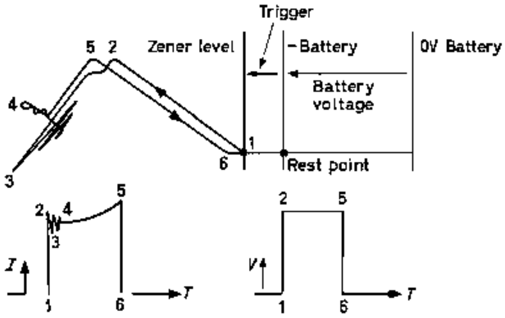


Fig. 12. Circuit characteristics
(a) *Ordinate—current (20mA/div)* *Abscissa—time (1msec/div)*
(b) *Ordinate—current (50mA/div)* *Abscissa—voltage (20V/div)*
(c) *Ordinate—current (10mA/div)* *Abscissa—time (100µsec/div)*
(d) *Ordinate—voltage (2V/div)* *Abscissa—time (100µsec/div)*

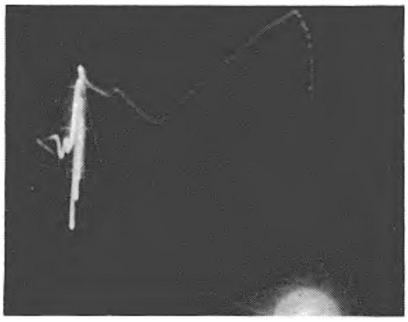
oscillator (Fig. 1) is comprised of VT_1 with base resistors R_1 and R_2, and C and R. When switch S is closed, the oscillator circuit is completed, and the frequency of oscillation (i.e. the repetition rate) is governed by the setting of potentiometer R.

With the switch S open, the stimulator can be triggered for single pulses by operating the push-button. This connects the positive lead of C_3, charged through R_3, to the emitter of VT_3 and thus the single pulse is initiated.

Consideration of the circuit shows that it is adaptable to other modes of operation. If the battery voltage is greater than the Zener voltage, a free running mode of operation with variable frequency is possible. The frequency depends on the d.c. core saturation of the transformer. This can be introduced through the secondary of the transformer from a battery, through a diode and a potentiometer. If the unijunction transistor VT_1 be eliminated, the circuit may be triggered by coupling any transient into the Zener diode path.



(a)



(b)

Fig. 13. Switching trajectories
(a) Schematic, with I/T (lower left) and V/T (lower right) plots
(b) Photograph of c.r.t. display
Abscissa—10 V/div   Ordinate—20mA/div

The unit has been tested in physiological experiments with frogs, turtles and rabbits to show vagus inhibition and vagus escape, as well as ordinary nerve-muscle stimulation. Heart sounds, driving a magnetic earphone acting as an electromagnetic transducer, triggered the unit so that one pulse of constant duration and amplitude was delivered for each heart beat. This output could be used to drive a simple heart-rate meter.

The conventional description of the stimulator with characteristic plots of currents and voltages as functions of time is satisfactory for a general understanding of the circuit, but a more complete analysis yields data of much more interest.

The output for stimulation is only a half-cycle, not a full cycle. A plot of the current-voltage characteristic, obtained from one section of the transformer windings reveals an N type curve. Fig. 10(a) shows a drawing of the curve, Fig. 10(b) a photograph of the c.r.t. trace.

With a slightly different setting of the oscilloscope, a simplified trace appears on the c.r.t. and this is shown in Fig. 11. The double trace is due to a phase shift between current and voltage. The electrical events are as follows: If for any reason (e.g. an external triggering voltage) the voltage passes the Zener break-down voltage (point 5, Fig. 11(a)), the Zener becomes conducting and switches to diode MR_2 (see Fig. 1) and the current is driven through point 10 back to point 1. The cycle is closed, or it can be repeated indefinitely (if the Zener voltage is exceeded).

Fig. 12 shows more details of the half cycle used for stimulation. These plots were taken from the output of the stimulator.

The traces in Fig. 13 show the switching trajectories of the output pulse. Switching is very fast and takes place between points 1 and 2, but is slowed between points 5 and 6 (see Fig. 13(a)). The plot may have a very irregular shape, depending on reactances in the feedback path (see Fig. 13(b)). Strong damped oscillations are visible between points 3 and 4.

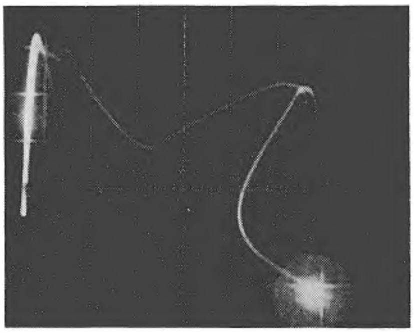


Fig. 14. Switching trajectories
Photograph of c.r.t. display
Ordinate—20mA/div   Abscissa—10V/div

Fig. 14 shows the switching from the start point where the Zener voltage has not been reached in the first cycle. After a few loops about the feedback path the voltage reaches the Zener voltage, and an output pulse is formed. The frequency of the loops increases until the phase shift around the feedback loop brings the impedance into the negative region, where it increases until it equals the total positive impedance of the circuit. The maximum amplitude of oscillation is then reached, but the oscillations are not sustained because the overall gain and energy available from the feedback are not sufficient.

The heavily damped oscillations vanish in the smooth top of the Zener pulse. (Compare 2,3,4 with 4,5 of the current against time plot of Fig. 13(a). This corresponds to 4,5 of the voltage/time plot of Fig. 13(a). Voltage changes on the diode are similar, but not identical. The circuit is monostable until the next trigger voltage, added to the battery voltage, is greater than the start voltage.

Acknowledgments

The co-operation of Dr. C. H. Best of the Banting and Best Department of Medical Research is gratefully acknowledged. Dr. D. W. Clarke assisted in the preparation of the manuscript.

The use of the stimulator in actual physiological experiments has been checked by Mr. A. Cronin and Mr. K. Bush.

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

By S. N. Bhattacharyya*, M.Sc., M.E.E. A.M.I.E.(Ind.) and J. K. Choudhury*, M.Sc., M.I.E.E.

The bridged-T network has certain definite advantages over the conventional bridges, particularly due to its comparative immunity from the effects of stray capacitive couplings. In the present investigation a configuration of the bridged-T network has been used, which essentially represents a compensated series-resonant circuit at balance. Test results are given to illustrate the application of the network for magnetic testing under various conditions of excitation of the sample. Measurements have been made over frequencies ranging from 50c/s to 40kc/s

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

ALTERNATING current bridges constitute by far the most versatile means of a.c. magnetic testing since they are suitable for measurements over a wide range of frequency and under varied conditions of magnetization in the test-sample. Also, a.c. bridges are particularly useful for magnetic measurements on small samples and at low flux-densities under which conditions the dissipation in the core is very small. Conventional bridges of the Owen[1-4], Campbell[4,5,6], Maxwell[4,7,8], Hay[4,9] etc. type are widely used for the purpose.

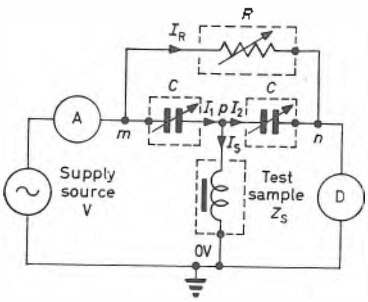

Fig. 1. Arrangement of the present bridged-T network

Most of the conventional bridges mentioned above constitute a four-arm network of the Wheatstone bridge type structure with the supply-source connected to a pair of opposite junctions and the detector connected across the other pair. For operation at audio and higher frequencies, one junction of the bridge, which is very often one terminal of the supply-source also, is connected to earth. Elaborate multiple screening of the bridge components is then necessary for the purpose of minimizing the effects of earth-capacitances at the other three junctions of the bridge. Alternatively, a complicated earthing device such as the Wagner earth has to be used in addition to elaborate screening of the components, for reducing the earth-capacitance effects.

The bridged-T network[4,10-14] has a new type of structure which incorporates the advantage of having one terminal of the supply-source and one terminal of the detector jointly connected to earth. This leaves only two major junctions of the bridge likely to be influenced by stray earth-capacitances. Simple screening of the bridge components is, however, sufficient to minimize the effects of earth-capacitances of these two junctions even at high frequencies[13,14].

The bridged-T is extensively used as a frequency selective network for the purpose of blocking a particular frequency. Its application for such purposes in harmonic analysers and wave analysers is well kown[12]. Recently, the bridged-T network has been advantageously used by Choudhury and Sen[13,14] for magnetic testing. In the configuration used by them, the test magnetic coil was used to bridge the two capacitors of a symmetrical *CRC* T-network. In the present investigation, the authors describe a different configuration of the bridge-T network for magnetic measurements, in which two equal capacitors form the series arms, and the test magnetic coil the shunt arm of a symmetrical T-network, whereas a resistance is used as the bridge across the two capacitors. The present configuration has certain definite advantages over the former one.

Description and Theory of the Network

The circuit diagram of the bridged-T network used by the present authors is given in Fig. 1. The test-sample Z_S and a pair of variable standard capacitors form a symmetrical T-network (C-Z_S-C) with the common junc-

SYMBOLS

a	=	area of cross-section of sample (m²)
B_m	=	maximum flux-density (Wb/m²)
C	=	capacitance (F)
f	=	frequency (c/s)
I	=	r.m.s. current (A)
l	=	mean length of magnetic circuit in sample (m)
L	=	self-inductance (H)
N	=	number of turns in exciting coil of sample
P	=	core-loss in sample (W)
P_S	=	specific core-loss (mW/kg)
R	=	resistance (Ω)
V	=	potential difference (V)
Z_S	=	gross impedance of sample (including resistance of coil winding) (Ω)
Z_S'	=	impedance of sample after allowing for ohmic resistance of its exciting coil (Ω)
(Z_t)	=	transfer impedance (Ω)
$\mu_r\angle\theta$	=	relative complex permeability
μ_r	=	modulus of relative complex permeability
θ	=	angle of relative complex permeability (degrees)
μ_0	=	permeability of free space $= 4\pi\times10^{-7}$ H/m
$\omega = 2\pi f$	=	angular frequency (rad/sec)

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tion at p. The capacitors are bridged over by a variable non-inductive resistor R. The positions of the a.c. supply-source and the detector are as indicated. Detector null is obtained by varying R and the C's together. The two capacitors are given equal values.

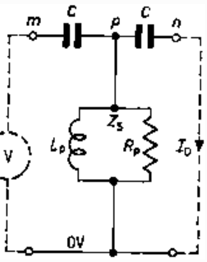


Fig. 2. The C-Z_S-C T-network having transfer impedance $(Z_t)_T$

It is well known that the bridged-T network is derived from the twin-T network. The basic twin-T network consists of two T-sections connected in parallel; the bridged-T is a special case of the twin-T network in which the shunt arm of one of the parallel T-sections is infinitely large. Tuttle[10] has shown that, for balance in either the twin-T or its special version—the bridged-T, the sum of the transfer impedances of the two networks must be zero.

The balance conditions of the configuration of the bridged-T network (Fig. 1) used in the present investigation may be developed by first considering the transfer impedance $(Z_t)_T$ of the T-network (C-Z_S-C) of Fig. 2, in which the test-sample Z_S has been considered equivalent to a parallel combination of a self-inductance L_p and a resistance R_p. This transfer impedance is given by:

$$(Z_t)_T = V/I_0 = -\frac{1}{R_pC^2\omega^2} + j\left[\frac{1}{L_pC^2\omega^3} - (2/\omega C)\right]$$
$$= R' + j\,[\omega L' - (1/\omega C')] \quad \ldots\ldots\ldots\ldots\ (1)$$

where

$$\left.\begin{aligned} R' &= -\frac{1}{R_pC^2\omega^2} \\ L' &= \frac{1}{L_pC^2\omega^4} \\ \text{and} \quad C' &= C/2 \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (1a)$$

with the asumption that the output terminals of the T-network are short-circuited. This assumption will evidently hold good, in effect, under the balanced condition of the network of Fig. 1, for which the two detector terminals are at the same potential.

The transfer impedance $(Z_t)_T$ will evidently reduce to a pure negative resistance R' if the parameters are so proportioned that:

$$\omega L' = 1/\omega C'$$

which gives:

$$L_p = 1/2\omega^2 C \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (2)$$

Under this condition, the transfer impedance is given by:

$$(Z_t)_T = R' = -\frac{1}{R_pC^2\omega^2} \quad \ldots\ldots\ldots\ldots\ (3)$$

Again, the transfer impedance $(Z_t)_R$ of the bridge arm R (Fig. 1) is obviously given by:

$$(Z_t)_R = R \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ (4)$$

Now, for complete bridge balance:

$$(Z_t)_T + (Z_t)_R = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (5)$$

Hence, from equations (3) and (4):

$$R_p = 1/\omega^2C^2R \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (6)$$

The balance conditions of the bridge are thus given by equations (2) and (6), from which the values of L_p and R_p are obtainable in terms of the values of bridge-components C and R at balance and the a.c. supply frequency.

The vector diagram of the voltages and currents of the balanced bridge of Fig. 1 is indicated in Fig. 3 which is self-explanatory.

Incidentally, an inspection of equation (1) shows that the transfer impedance $(Z_t)_T$ of the C-Z_S-C T-network consists, in general, of a series circuit comprising a self-inductance L', a capacitance C' and a negative resistance R'. Equation (2) represents a special case of this transfer impedance in which the above series circuit has been made resonant by making $\omega L' = 1/\omega C'$.

Calculation of Core-loss and Complex Permeability

Calculation of Core-loss

If V_S is the voltage across the test-sample, the total loss in the sample is given by:

$$P_{\text{Total}} = V_S^2/R_p = \omega^2C^2RV_S^2 \quad \ldots\ldots\ldots\ (7)$$

from equation (6). The loss given in equation (7) comprises of the core-loss in the sample and the copper-loss in its exciting coil. If R_w be the resistance of this coil, the core-loss P in the sample will be given by:

$$P = \omega^2C^2RV_S^2 - I_S^2R_w \quad \ldots\ldots\ldots\ldots\ (8)$$

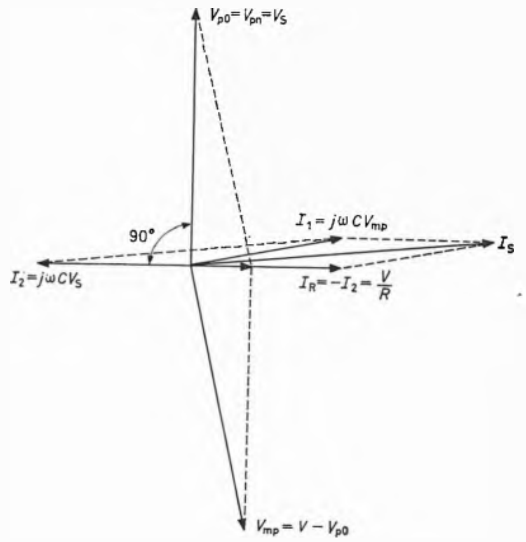


Fig. 3. Vector diagram of the voltages and currents of the balanced bridge

V = Reference vector
I_R = V/R in phase with V
I_2 = $-I_R$
V_{pn} = $I_2/j\omega C$ = V_{pO} = V_S
V_{mp} = V-V_{pO}
I_1 = $j\omega CV_{mp}$
I_S = $I_1 - I_2$ = $I_1 + I_R$

where I_S is the r.m.s. current through the sample. It may be noted that at balance of the bridge, I_S becomes equal to the total input current of the bridge, which may be measured with an r.m.s. ammeter A connected as shown in Fig. 1. The voltage V_S across the sample is to be measured with a high impedance valve voltmeter (not shown in Fig. 1).

Calculation of Complex Permeability

Complex permeability[15] $\mu\angle\theta$ of a material may be expressed as:

$$\mu\angle\theta = \mu_0\mu_r\angle\theta = \mu_0\left[\mu_r\cos\theta - j\mu_r\sin\theta\right] \quad \ldots\ldots (9)$$

where μ_r is the modulus and θ is the angle of the relative complex permeability, μ_0 being the permeability of free space ($4\pi \times 10^{-7}$H/m).

It may easily be shown[1,2,15] that for a sample having a closed magnetic circuit of mean length l metre, uniform cross-sectional area a metre^2 and having N turns in the exciting coil:

$$\mu_r = \frac{Z_S' l}{\mu_0\omega a N^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (10a)$$

and

$$\theta = \arctan R_s'/L_s\omega \quad \ldots\ldots\ldots\ldots (10b)$$

where Z_S' is the effective impedance of the sample, duly corrected for the coil winding resistance R_w, and R_s' and L_s are respectively the series resistance and self-inductance equivalent to Z_S'.

R_s' and L_s may be expressed as:

$$R_s' = \frac{R}{\omega^2C^2R^2 + 4} - R_w \quad \ldots\ldots\ldots\ldots (11)$$

$$L_s = \frac{2}{\omega^2C\,(\omega^2C^2R^2 + 4)} \quad \ldots\ldots\ldots\ldots (12)$$

Equations (11) and (12) easily give:

$$Z_S' = \left[R_s'^2 + (\omega L_s)^2\right]^{\frac{1}{2}} = 1/\omega C \left[\frac{1 - 2RR_w\omega^2C^2 + R_w^2\omega^2C^2\,(\omega^2C^2R^2+4)}{\omega^2C^2R^2 + 4}\right]^{\frac{1}{2}} \quad \ldots (13)$$

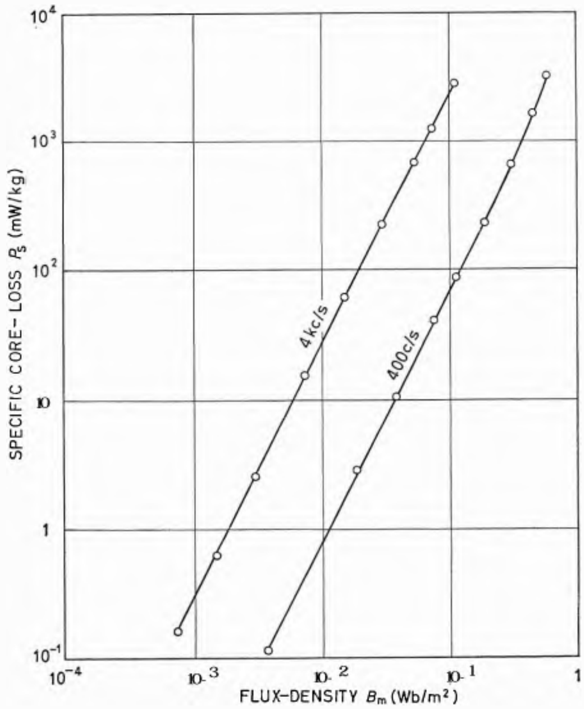
Equations (10b), (11) and (12) give:

$$\theta = \arctan \tfrac{1}{2}\,\omega C\left[R - R_w\,(\omega^2C^2R^2 + 4)\right] \quad \ldots (14)$$

The Apparatus Used

Precision grade laboratory type decade boxes with very low residuals of known values were employed as the components C and R. An RC oscillator followed by a power amplifier constituted the main supply source. For 50c/s measurements, the supply mains followed by a voltage stabilizer[16] was employed. A cathode-ray oscilloscope was used as detector. Voltages were measured with a high impedance valve-voltmeter.

Fig. 4. Specific core-loss P_s as a function of peak flux-density B_m at different frequencies (400c/s and 4 000c/s)



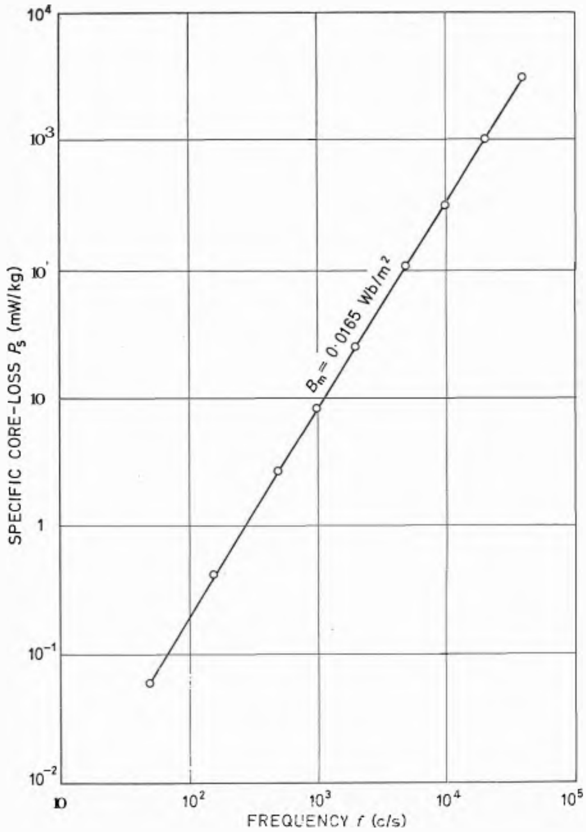


Fig. 5. Specific core-loss P_s as a function of frequency f at constant flux-density (B_m = 0·0165 Wb/m²)

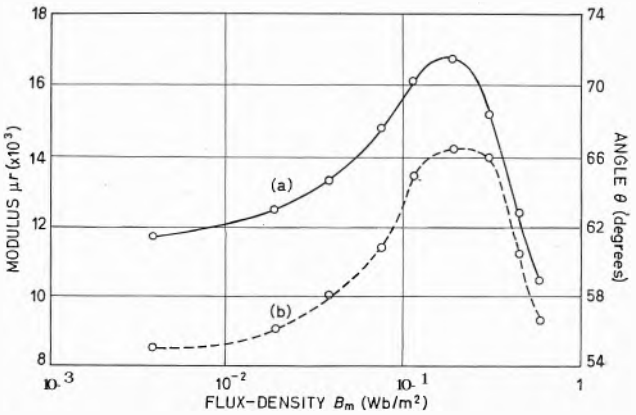


Fig. 6. Relative complex permeability as a function of peak flux-density at constant frequency (f = 400c/s)
(a) Curve of μ_r/B_m
(b) Curve of θ/B_m

Examples of Measurements

Tests were carried out on a sample consisting of a clock-spring type ring core of 12·5mm wide Mumetal tape of thickness 0·375mm and weighing about 0·25kg. The following measurements were made:

(a) Specific core-loss P_S (mW/kg) as a function of maximum flux density B_m (Wb/m²) for two different frequencies ($f = 400$c/s and 4 000c/s)—Fig. 4 illustrates the relation P_S/B_m.

(b) Specific core-loss as a function of frequency for a fixed B_m (0·0165Wb/m²)—Fig. 5 illustrates the relation P_S/f.

(c) Relative complex permeability (modulus μ_r and angle θ) as a function of B_m at constant frequency (400c/s). Fig. 6 illustrates the relation μ_r/B_m and θ/B_m.

The testing of core-loss and permeability of a magnetic material sets up extremely stringent conditions on the measuring equipments, since the method must be capable of measuring very small losses at low flux-densities and low frequencies and also very high losses at high flux densities and high frequencies. It will be observed that with the present bridged-T network, the magnitude of measured core-loss varies over six decades involving variations of frequency and flux-density over four decades each. Essentially the same set-up has been used for all these measurements. The method thus proved to be extremely flexible.

Conclusion

Bridged-T networks possess, as pointed out earlier, the remarkable feature that the supply-source and the detector have one point in common. Earthing of this common point eliminates the need of elaborate multiple screening of components used and also of any special form of earthing, e.g. Wagner earthing. The earth-capacitance of the screens of the resistor R and the right-hand capacitor C shunt the detector, that of the screen of the left-hand capacitor shunts the supply-source, whereas the screen of the sample Z_S is directly earthed. Thus, none of the earth-capacitances affects the balance conditions of the bridge. This feature enables the bridge to be employed up to quite high frequencies. In fact, high frequency limitations of the components themselves, rather than any shortcoming of the bridge arrangement, set the upper limit of operation of the bridge.

A consideration of the present bridged-T network reveals the fact that, at balance, the network behaves as a compensated series-resonant circuit shunted across the supply-source. Thus, near balance, the bridge behaves almost as a resistance of low value. This significantly reduces the effects of capacitive couplings between the source and the detector and between these and the body of the operator and other neighbouring conductors.

As has already been mentioned, the bridged-T configuration used in the present investigation is basically a compensated series-resonant circuit. In it, the positions of the sample and the variable resistor have been interchanged with their respective positions in the bridged-T set-up used earlier by Choudhury and Sen[13,14]. This earlier bridge was essentially a compensated parallel-resonant circuit. By using the compensated series-resonant circuit in preference to the compensated parallel-resonant circuit, the following two distinct advantages have been realized in the present investigation:

(a) Only one-quarter as much capacitance is required for balancing the bridge.

(b) The value of the variable resistance required for balance is very much lower.

The use of lower capacitance and lower resistance evidently minimizes to a great extent the errors due to residuals in the components and stray earth admittances.

It is thus seen that the bridged-T network presented herein offers several advantages for a.c. magnetic measurements particularly at high frequencies. It has also certain definite advantages over the alternative set up used earlier by Choudhury and Sen.

Acknowledgements

The authors wish to thank the authorities of the Jadavpur University, Calcutta, for providing laboratory facilities for the investigation. Their thanks are also due to Prof. H. C. Guha, Head of the Electrical Engineering Department, Jadavpur University, for his keen interest in their researches, and to Mr. N. C. Dey for helpful services in the laboratory.

APPENDIX

DERIVATION OF EXPRESSIONS FOR R_s' AND L_s

To derive equations (11) and (12) giving R_s' and L_s respectively, let Z_S be equivalent to a resistance R_s and a self-inductance L_s in series. In this case, the transfer impedance $(Z_t)_T$ will be given by[10]:

$$(Z_t)_T = -(1/\omega^2C^2)\frac{R_s}{R_s^2 + \omega^2L_s^2} + j\left[\frac{\omega L_s}{\omega^2C^2}\,\frac{1}{R_s^2+\omega^2L_s^2} - (2/\omega C)\right] \ldots (15)$$

Now, applying the same considerations as for deriving equations (2) and (6), the following relations are obtained:

$$1/\omega^2C^2\,\frac{\omega L_s}{R_s^2 + \omega^2L_s^2} - 2/\omega C = 0 \quad \ldots\ldots\ldots\ldots (16)$$

and

$$1/\omega^2C^2\,\frac{R_s}{R_s^2 + \omega^2L_s^2} = R \quad \ldots\ldots\ldots\ldots (17)$$

or:

$$\frac{\omega L_s}{R_s^2 + \omega^2L_s^2} = 2\omega C \quad \ldots\ldots\ldots\ldots (18)$$

and

$$\frac{R_s}{R_s^2 + \omega^2L_s^2} = \omega^2C^2R^2 \quad \ldots\ldots\ldots\ldots (19)$$

Squaring and adding equations (18) and (19):

$$\frac{1}{R_s^2 + \omega^2L_s^2} = \omega^2C^2(\omega^2C^2R^2 + 4) \quad \ldots\ldots (20)$$

The expression for L_s now follows immediately from equations (18) and (20). The expression for R_s' also follows from equations (19) and (20) if it is noted that:

$$R_s' = R_s - R_w \quad \ldots\ldots\ldots\ldots (21)$$

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Direct Reduction of a Multi-Mesh Circuit to a Four-Terminal Network

By D. W. W. Rogers*, A.M.I.E.R.E.

A method is described for obtaining the four-terminal equations of a multi-mesh network. The method is based on a technique whereby the impedance (or admittance) parameters of any network may be written down directly in terms of reflected impedance. A similar treatment employing signal flow analysis has been carried out by Robichaud[1].

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

In general when a network is to be analysed in such terms as say voltage gain, and input impedance, the first approach is to give due consideration as to whether any simplification is possible by means of Thevenin's or other network theorems. Such an approach is economical in effort because these theorems deal immediately with the topology of the network, whereas the process of setting up circuit equations and obtaining their solution by determinants or matrices is lengthy and ineffective, because the topological information relating to the particular network in question has been discarded. In the method to be described there is a direct one to one correspondence between the network configuration and the impedance (or Z) parameters, when expressed in terms of reflected impedance. Further it has been found from experience that this method is particularly effective, when used in conjunction with the Unit Current technique[2] for obtaining the Z parameters of the individual meshes.

In order to clarify the presentation of the method and the unit current determination of the Z parameters, the concept of reflected impedance is first reviewed in relation to the input impedance of a common base equivalent-T transistor network from a tutorial point of view.

Unit Current Determination of Z Parameters

The network equations relating to Fig. 1 are

$$V_1 = i_1(r_e+r_b) - i_2 r_b \quad \dots\dots\dots\dots \quad (1)$$

$$V_2 = -i_1(r_b+\alpha r_c) + i_2(r_b+r_c) \quad \dots\dots\dots \quad (2)$$

Equations (1) and (2) may be formally written as

$$V_1 = i_1 Z_{11} + i_2 Z_{12} \quad \dots\dots\dots\dots \quad (3)$$

$$V_2 = i_1 Z_{21} + i_2 Z_{22} \quad \dots\dots\dots\dots \quad (4)$$

and define the well known four terminal impedance parameters Z_{11}, Z_{12}, Z_{21}, and Z_{22}. An alternative definition of these parameters is as follows:

'Z_{mn} is numerically equal to the voltage drop induced in the m^{th} mesh by a Unit Current in the n^{th} mesh when all the voltage sources are short-circuited.' The latter definition enables the impedance parameters of a given mesh to be found by inspection. Consider Z_{12} for example in Fig. 1 A current of 1A in mesh 2 induces a voltage drop of $-r_b$ in mesh 1. It should be noted in Fig. 1, that both the currents and the voltages have a clockwise sense.

First Order Reflected Impedance Term

If the relation $V_2 = -i_2 r_L$ is substituted in equation (2), and i_2 in turn, substituted in equation (1), then the input impedance

$$Z_i = V_1/i_1 = r_e + r_b - \frac{r_b(r_b + r_c\alpha)}{r_b + r_c + r_L}$$

$$= Z_{11} - (Z_{12}Z_{21}/Z_{22}) \quad \dots\dots\dots\dots \quad (5)$$

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Here $Z_R = Z_{12}Z_{21}/Z_{22}$ is the reflected impedance due to the loading effect of mesh 2. In particular it should be noted that for the simple first order case under discussion, the reflected impedance is equal to the product of the forward and reverse coupling impedances between the meshes in question, divided by the self impedance of the mesh being reduced.

The General Multi-Mesh Reflected Impedance Term

For a multi-mesh network, where m and n are the two

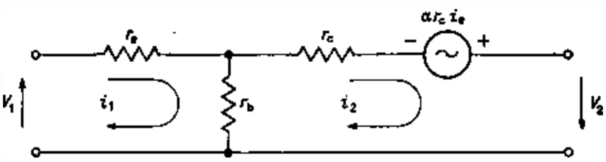


Fig. 1. Common base equivalent T transistor network

four terminal meshes under consideration, the reflected impedance Z_R of Z_{mn} has the form

$$\sum \frac{Z_{mn} \,.\, \Delta_{mn}}{\Delta}$$

where Z_{mn} = the product of the individual coupling impedances along a possible path taken from the m^{th} to the n^{th} mesh, divided by $Z_{rr} \cdot\cdot\cdot . \cdot\cdot\cdot . \cdot\cdot Z_{uu}$, the product of the self-impedances between the m^{th} and the n^{th} mesh for the particular path taken. When the reflected impedance of an impedance parameter such as Z_{mm} or Z_{nn} is being determined, paths through the other n^{th} or m^{th} mesh respectively must be excluded.

$$\Delta = 1 - \sum_{\mu} P_{\mu 1} + P_{\mu 2} + P_{\mu 3} + \cdots$$

$P_{\mu\theta}$ = a dimensionless factor which is the product of the θ^{th} possible combination of coupling impedances not forming part of the m^{th} or n^{th} mesh, divided by immediately adjacent self-impedances of the form

$$\frac{Z_{rs} \,.\, Z \cdots . Z_{qr}}{Z_{rr} \ldots . Z_{qq}}$$

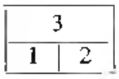
Δ_{mn} = the value of Δ for that part of the network not touching the m^{th} or n^{th} path. The general reflected impedance term is similar in form to the general gain formula of S. J. Mason[3] used in signal flow graphs.

Examples of Multi-Mesh to Four-Terminal Network Reduction

The general reflected impedance formula may appear formidable to use at first glance, but its actual application is quite straightforward. In all the following examples the four-terminal representation to be determined is taken between meshes 1 and 2. The multi-mesh network is drawn schematically, the forward and reverse coupling impedances between two meshes being represented by a straight line. The examples are given in general Z parameter form, as once the four-terminal reduction has been carried out; the actual values of the Z parameters may be substituted directly by means of the Unit Current technique.

EXAMPLE 1

$$Z_{11(\text{Total})} = Z_{11} - \frac{Z_{13}Z_{31}}{Z_{33}},\quad Z_{12(T)} = Z_{12} - \frac{Z_{13}Z_{32}}{Z_{33}},$$

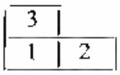


$$Z_{21(T)} = Z_{21} - \frac{Z_{23}Z_{31}}{Z_{33}},\quad Z_{22(T)} = Z_{22} - \frac{Z_{23}Z_{32}}{Z_{33}},$$

Here Z_{r11} for example is the simple first order coupling between mesh 1 and mesh 3, mesh 2 being excluded as $m = 1$. No other paths are possible, and $\Delta_{11} = 1$ as there are no other parts of the network which do not touch the 1st and 2nd meshes, in this simple case. In the case of Z_{R12}, the coupling path is from mesh 1 to mesh 2. Again no other paths are possible, and there are no other parts of the network which do not touch the meshes 1 and 2.

EXAMPLE 2

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}}{Z_{33}},\quad Z_{12(T)} = Z_{12},$$

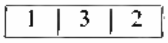


$$Z_{21(T)} = Z_{21},\quad Z_{22(T)} = Z_{22},$$

Here Z_{R11} has only one possible path, i.e. from mesh 1 to mesh 3, and again it is a simple first order coupling.

EXAMPLE 3

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}}{Z_{33}},\quad Z_{12(T)} = -\frac{Z_{13}Z_{32}}{Z_{33}},$$



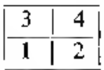
$$Z_{21(T)} = -\frac{Z_{23}Z_{31}}{Z_{33}},\quad Z_{22(T)} = Z_{22} - \frac{Z_{23}Z_{32}}{Z_{33}},$$

Here simple couplings are possible from mesh 3 to both 1 and 2, but not directly between meshes 1 and 2, so $Z_{12} = Z_{21} = 0$. However, the reflected impedances $Z_{13}Z_{32}/Z_{33}$ and $Z_{23}Z_{31}/Z_{33}$ both exist.

EXAMPLE 4

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}\,.\,Z_{31}/Z_{33}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}},\quad Z_{12(T)} = Z_{12} - \frac{Z_{13}Z_{34}Z_{42}/Z_{33}Z_{44}}{1 - \dfrac{Z_{34}\,.\,Z_{43}}{Z_{33}Z_{44}}}$$

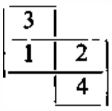
$$Z_{21(T)} = Z_{21} - \frac{Z_{24}Z_{43}Z_{31}/Z_{33}Z_{44}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}},\quad Z_{22(T)} = Z_{22} - \frac{Z_{24}Z_{42}/Z_{44}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}}$$



Here the reflected impedance $Z_{R(mn)}$ is in each case similar to the previous examples, except that a path through meshes 3 and 4 must involve three coupling impedances. Furthermore the coupling impedances Z_{34} and Z_{43} do not touch the m^{th} and n^{th} paths. Accordingly, these terms are included in Δ.

EXAMPLE 5

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}}{Z_{33}},\quad Z_{12(T)} = Z_{12},$$



$$Z_{21(T)} = Z_{21},\quad Z_{22(T)} = Z_{22} - \frac{Z_{24}Z_{42}}{Z_{44}},$$

In this case the reflected impedances are all simple first order terms.

EXAMPLE 6

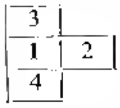
$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}/Z_{33}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}},\quad Z_{12(T)} = -\frac{Z_{13}Z_{34}Z_{42}/Z_{33}Z_{44}}{1 - \dfrac{Z_{43}Z_{34}}{Z_{33}Z_{44}}},$$

$$Z_{21(T)} = -\frac{Z_{24}Z_{43}Z_{31}/Z_{44}Z_{33}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{44}Z_{33}}},\quad Z_{22(T)} = Z_{22} - \frac{Z_{24}Z_{42}/Z_{44}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{44}Z_{33}}},$$

This example should be compared with example 4, except in this case there is no direct coupling between meshes 1 and 2, and hence Z_{12} is zero in example 6, and also Z_{21} of course.

EXAMPLE 7

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}}{Z_{33}} - \frac{Z_{14}Z_{41}}{Z_{44}},\quad Z_{12(T)} = Z_{12},$$

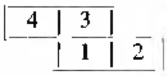


$$Z_{21(T)} = Z_{21},\quad Z_{22(T)} = Z_{22}$$

This example should be compared with that of example 2, except in this case the addition of mesh 4 has contributed the extra term $Z_{14}Z_{41}/Z_{44}$.

EXAMPLE 8

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}/Z_{33}}{1 - \dfrac{Z_{43}Z_{34}}{Z_{33}Z_{44}}},\quad Z_{12(T)} = Z_{12},$$

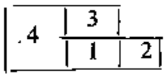


$$Z_{21(T)} = Z_{21},\quad Z_{22(T)} = Z_{22},$$

This example should be compared with that of example 2. The additional mesh 4 coupled only to mesh 3 introduces the extra coupling impedances Z_{34}, and Z_{43}, not touching meshes 1 and 2; and hence $\Delta = 1 - Z_{43}Z_{34}/Z_{33}Z_{44}$.

EXAMPLE 9

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{31}/Z_{33} - Z_{14}Z_{41}/Z_{44}}{1 - \dfrac{Z_{43}Z_{34}}{Z_{33}Z_{44}}}$$



$$-\frac{Z_{14}Z_{43}Z_{31}/Z_{44}Z_{33} - Z_{13}Z_{34}Z_{41}/Z_{44}Z_{33}}{1 - \dfrac{Z_{43}Z_{34}}{Z_{33}Z_{44}}},\quad Z_{12(T)} = Z_{12}$$

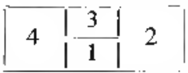
$$Z_{21(T)} = Z_{21},\quad Z_{22(T)} = Z_{22}$$

Here the reflected impedance $Z_{R(11)}$ has two possible simple coupling paths to be considered, and in addition the two coupling paths via meshes 1, 3 and 4. Coupling impedances Z_{34} and Z_{43} are not touching meshes 1 and 2, and hence $\Delta = 1 - Z_{43}Z_{34}/Z_{33}Z_{44}$.

EXAMPLE 10

$$Z_{11(T)} = Z_{11} - \frac{Z_{13}Z_{34}Z_{41}/Z_{33}Z_{44} - Z_{14}Z_{43}Z_{31}/Z_{33}Z_{44} - Z_{13}Z_{31}/Z_{33} - Z_{14}Z_{41}/Z_{44}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}}$$

$$Z_{12(T)} = Z_{12} - \frac{Z_{13}Z_{32}/Z_{33} - Z_{14}Z_{43}Z_{32}/Z_{33}Z_{44}}{1 - \dfrac{Z_{43}Z_{34}}{Z_{33}Z_{44}}}$$



$$Z_{21(T)} = Z_{21} - \frac{Z_{23}Z_{34}Z_{41}/Z_{33}Z_{44}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}}$$

$$Z_{22(T)} = Z_{22} - \frac{Z_{23}Z_{32}/Z_{33}}{1 - \dfrac{Z_{34}Z_{43}}{Z_{44}}}$$

In the case of $Z_{11(T)}$, there are four possible paths to be considered, and the coupling impedances Z_{43} and Z_{34} which do not touch meshes 1 and 2. $Z_{R(21)}$ for example has only one possible path via mesh 3, and mesh 4 to mesh 1.

EXAMPLE 11

In this last example a five mesh network is reduced in order to illustrate the case when Δ_{mn} is not equal to unity as in the previous ten examples—

$$Z_{11(T)} = \frac{Z_{11} - \dfrac{Z_{13}Z_{31}}{Z_{33}}\left[1 - \dfrac{Z_{54}Z_{45}}{Z_{55}Z_{44}}\right] - \dfrac{Z_{15}Z_{51}}{Z_{55}}\left[1 - \dfrac{Z_{34}Z_{43}}{Z_{33}Z_{44}}\right] - \dfrac{Z_{13}Z_{34}Z_{45}Z_{51}}{Z_{33}Z_{44}Z_{55}} - \dfrac{Z_{15}Z_{54}Z_{43}Z_{31}}{Z_{55}Z_{44}Z_{33}}}{\Delta}$$

$$Z_{21(T)} = \frac{-\dfrac{Z_{24}Z_{45}Z_{51}}{Z_{44}Z_{55}} - \dfrac{Z_{24}Z_{43}Z_{31}}{Z_{44}Z_{55}}}{\Delta},$$

$$\text{where } \Delta = 1 - \frac{Z_{54}Z_{45}}{Z_{44}Z_{55}} - \frac{Z_{34}Z_{43}}{Z_{33}Z_{44}}$$

$$Z_{12(T)} = \frac{-\dfrac{Z_{24}Z_{45}Z_{51}}{Z_{44}Z_{55}} - \dfrac{Z_{24}Z_{43}Z_{31}}{Z_{44}Z_{55}}}{\Delta},$$

$$Z_{22(T)} = \frac{Z_{22} - \dfrac{Z_{24}Z_{42}}{Z_{44}}}{\Delta},$$

1	3	
5	4	2

In this case it is possible to take simple reflected impedances for $Z_{R(11)}$ and still have factors in Δ which are not in the meshes immediately adjacent. Hence Δ_{mn} for $Z_{13}Z_{31}/Z_{33}$, and $Z_{15}Z_{51}/Z_{55}$ respectively is $1 - Z_{54}Z_{45}/Z_{55}Z_{44}$ and $1 - Z_{34}Z_{43}/Z_{33}Z_{44}$.

Conclusions

A method has been presented whereby a multi-mesh network may be reduced directly to the impedance parameters of a four-terminal network, by means of reflected impedance terms. As these terms relate directly to a physical mesh or meshes, the effect of a particular mesh may be separately evaluated by inspection. The method is particularly efficient, when used in combination with the Unit Current technique[3] for obtaining the Z parameters of the meshes.

By writing a given impedance parameter in the form of reflected impedance terms, it is possible to reverse the reduction process and synthesize a suitable network.

Acknowledgment

The author is indebted to Mr. R. E. Hardman, Senior Lecturer, Oldham Technical College for stimulating discussions in connexion with network reduction.

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A Numerically Controlled Draughting Machine

At the Toulouse factory of Sud Aviation a large numerically controlled draughting machine is being used on the Concord supersonic airliner project. Built by Cramic Engineering Co. Ltd in their factory at Bridge Road, Southall, Middlesex, this machine is one of the largest automatic draughting machines in operation anywhere in the world. It is equipped with a draughting table measuring 160in by 80in and can draw lines to an accuracy of ±0·002in/ft or ±0·005in overall across the full 160in length of the board. Actually these figures are the performance figures defined in the machine specification. In practice, tests have confirmed that the machine draws lines even more accurately than this. The machine can thus draw contours much more precisely than an experienced loftsman.

The machine supplied to Sud Aviation was fitted with a Ferranti Mk.IV magnetic-tape numerical control system. Control data defining the shape of the lines to be drawn are derived by computer and recorded on magnetic tape by a curve generator. When the tape is played back on the console attached to the draughting machine it operates the various machine drives and causes the required lines to be accurately drawn.

Drawings can be made on paper, linen or stable-base plastic drawing materials in ink. A diamond scribing head is provided enabling the machine to be used for accurately marking out sheet metal components.

Power to traverse the column is provided by a hydraulic servo-motor operating through a rack and pinion drive. Two pinions, one pre-loaded against the other to eliminate backlash, engage with the rack. To reduce frictional effects to a minimum, the column is supported on Dexter linear recirculating ball bearings.

The saddle is also supported on Dexter linear bearings and is moved by means of a Rotax recirculating ball nut and leadscrew driven by a hydraulic motor.

Dexter linear bearings support the drawing head on the saddle.

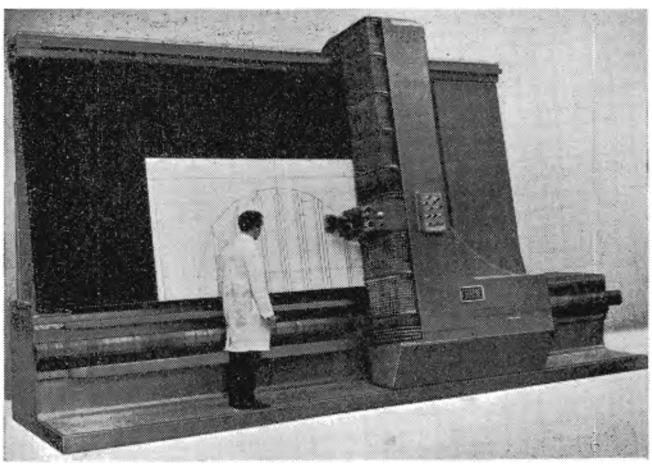
A turret fitted to the drawing head holds four pens plus a spring-loaded diamond scriber for marking metal. The pen to be used is selected manually before a drawing sequence is initiated. 'Pen on' and 'pen off' positions can be selected by program, or manually. Uniform dotted (0·4in. constant pitch) as well as continuous lines can be drawn.

Under numerical control, lines can be drawn at any speed up to a maximum of 120in/min along each axis. Under manual control, 'fast' or 'slow' speeds can be selected by push-buttons fitted in a control panel on the machine column. Fast is 120in/min—slow can be adjusted to suit the application, or even be made infinitely variable.

A projector microscope (×5 power) is fitted on the drawing head turret for setting the machine to a datum point. Vernier manual controls are provided to enable the operator to position the microscope cross-wires precisely over the datum point.

The machine is also equipped with a Ferranti industrial measuring system fitted to the x and y axes. Using the projector microscope, the operator can position the machine over any point of a drawing and read off a digital numerical display its precise co-ordinate displacement from a pre-set datum to the nearest 0·001in or 0·02mm. This information may also be printed out and punched in paper tape if required. Sequential reference numbers are recorded to identify every co-ordinate reading made.

The Cramic numerically controlled draughting machine



A Diode Phase-Sensitive Detector

By B. S. Leela* and R. Shankar*

A diode phase-sensitive detector circuit is analysed in detail from the point of view of non-linearity. Computer solution of the equations concerned leads to new criteria for selection of component values. In particular, it is shown that considerably lower non-linearity figures are obtainable through comparatively minor restrictions on signal level.

Results are presented in the form of a 'quick-look' Table.

(Voir page 705 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 712)

THE effect of loading on phase-sensitive detectors has been studied in detail by Krishnan and Chidambaram[1].

Fig. 1 shows the circuit of a diode phase-sensitive detector. E_1 and E_2 are the amplitudes of the reference and signal voltages assumed to be in phase. The sum of the voltages E_1 and E_2 is applied to one of the diodes MR_1 and the difference to the diode MR_2. R_1 is the effective resistance of each of the diodes which also includes any resistance connected in series with them. CR_2 is assumed to be very large compared to the period of the applied voltage so that the output voltages E_A and E_B may be assumed to be steady.

If a load resistance R is connected across the output, due to the current flowing from node A to node B, the effective resistance R_2 of the two diodes changes causing

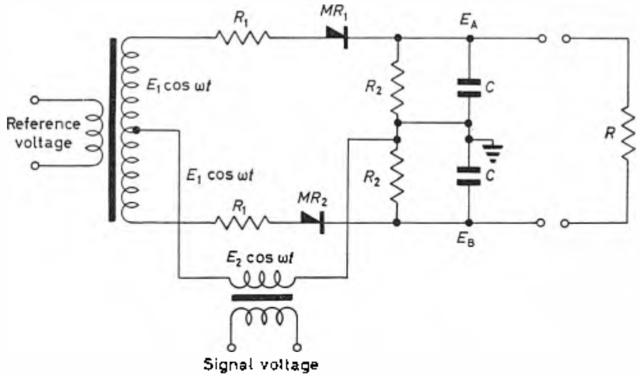


Fig. 1. Arrangement of the diode phase-sensitive detector

the transfer ratios (k_1 and k_2) of the two diodes (defined as the ratio of the d.c. output voltage to the peak value of the input a.c.) to vary considerably with the signal, causing non-linearity at the output.

The output of the detector increases with the increase in signal (or with the increasing of x which is the signal to reference voltage ratio). With increase in signal, there is a value of x, say x_{crit}, where the diode MR_2 stops conducting.

This happens when the voltage transferred from node A to node B through R equals the peak value of the voltage applied to the diode MR_2. For still higher values of signal, the sensitivity falls rapidly making the detector highly non-linear.

The authors of the earlier paper, have calculated the total percentage of non-linearity N_o of the detector at x_{crit}, where N_o is given by the equation[1].

$$N_o = \frac{\alpha_o - \alpha_{crit}}{\alpha_o} \times 100 \quad \ldots\ldots\ldots\ldots (1)$$

where α_o and α_{crit} are the sensitivity of the detector at zero signal and x_{crit} respectively.

It is comparatively easy to calculate the sensitivity α_o of the detector for small values of the signal voltage ('zero signal') and at x_{crit} and calculate the total percentage non-linearity from 0 to x_{cirt} by using equation (1). It can be shown that[1]:

$$\alpha_o = \frac{2k_o}{1 + \dfrac{2n_2}{n_1(\theta_o/\pi) + 1}} \quad \ldots\ldots\ldots\ldots (2)$$

and:

$$\alpha_{(crit)} = \frac{2k_{1(crit)}}{n_2 + 1 - n_2 k_{1(crit)}} \quad \ldots\ldots\ldots (3)$$

where θ_o is the semi angle of conduction of the diodes: $k_o = \cos\theta_o$, $k_1 = \cos\theta_1$, $k_2 = \cos\theta_2$, $n_1\ R_2/R_1$, $n_2 = R_2/R$

θ_1 and θ_2 are equal to θ_o for the two diodes when the signal is zero. With an increase in signal, θ_1 the semi angle of conduction of the diode MR_1 increases and θ_2, that of the diode MR_2, decreases.

θ_o is given by the equation[1]:

$$\pi/n_1 = \tan\theta_o - \theta_o \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

At x_{crit} it is known that MR_2 stops conducting and $\theta_2 = 0$. At this point the resistance R_2 becomes effectively paralleled by $(R + R_2)$. θ_{1c} at x_{crit} of the diode MR_1 is calculated in a similar manner to the calculation of θ_o except that R_2 is replaced by:

$$\frac{R_2(R + R_2)}{2R_2 + R}$$

θ_{1c} is given by the equation[1]:

$$\pi/n_1 \frac{1 + 2n_2}{1 + n_2} = \tan\theta_{1c} - \theta_{1c} \quad \ldots\ldots\ldots\ldots (5)$$

Equation (5) can also be arrived at by substituting $\theta_2 - 0$ in either of the equations (7) or (8), since $\theta_2 = 0$ at x_{crit}.

The non-linearity at x_{crit} however does not give any idea about the behaviour of the detector at different signal levels. It appeared probable that the non-linearity of the detector at lower signal voltages may be considerably less and hence it was considered worthwhile to calculate the non-linearity at different signal levels. The calculation of the non-linearity N_x at different values of x involves calculating α_x which in turn involves calculating the angles of conduction.

α_x, the sensitivity at different values of x is given by the equation[1]:

$$\alpha_2 = k_2 + k_1 - \frac{k_2 - k_1}{x} \quad \ldots\ldots\ldots\ldots (6)$$

Due to the difficulties involved in numerically calculating θ_1 and θ_2 from the two transcedental equations (7) and (8) the calculation of N_x has not been done before. It was found possible to calculate the non-linearity of the

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detector at different signal levels on a computer by Newton's approximation method described below:

Calculation of θ_1 and θ_2

The two transcedental equations involving semi angles of conduction of the two diodes for a given value of x, n_1 and n_2 are given by[1]:

$$n_1/\pi\,(\sin\theta_1 - \theta_1\cos\theta_1) = \cos\theta_1\,(1+n_2) - wn_2\cos\theta_2 \quad \ldots\ldots\ldots\ldots \quad (7)$$

$$n_1/\pi\,(\sin\theta_2 - \theta_2\cos\theta_1) = \cos\theta_2\,(1+n_2) - (n_2/w)\cos\theta_1 \quad \ldots\ldots\ldots\ldots \quad (8)$$

where $w = \dfrac{1-x}{1+x}$

θ_1 and θ_2 cannot be calculated directly from these equations. Hence a graphical method for solving θ_1 and θ_2 was adopted by the authors of the earlier papers. The accuracy obtainable by such a method is low. The method described below provides sufficiently accurate solutions for θ_1 and θ_2 when programmed on a computer.

Equations (7) and (8) can be rewritten as:

$$f_1(\theta_1, \theta_2) = (n_1/\pi)\sin\theta_1 - \{1+n_2+(n_1/\pi)\,\theta_1\}\cos\theta_1 + wn_2\cos\theta_2 = 0 \quad \ldots\ldots\ldots\ldots \quad (9)$$

$$f_2(\theta_1, \theta_2) = (n_1/\pi)\sin\theta_2 - \{1+n_2+(n_1/\pi)\,\theta_2\}\cos\theta_2 + (n_2/w)\cos\theta_1 = 0 \quad \ldots\ldots\ldots \quad (10)$$

TABLE 1

The percentage non-linearity for the values of x equal to $0{\cdot}25x_c$, $0.5x_c$, $0.75x_c$, $0{\cdot}9x_c$ and x_c for various values of n_1 and n_2

n_1 \ n_2	100	30	10	3	1	0·3	0·1	0·03	0·01
	0·1	0·1	0·1	0·1	—	—	—	—	—
	0·5	0·5	0·4	0·3	0·2	0·1	—	—	—
1	1·4	1·4	1·3	1·0	0·6	0·2	—	—	—
	2·8	2·7	2·6	2·1	1·4	0·5	0·1	—	—
	5·2	5·1	4·9	4·3	3·2	1·7	0·7	0·2	0·1
	0·3	0·3	0·2	0·2	0·1	—	—	—	—
	1·2	1·1	1·0	0·8	0·4	0·1	—	—	—
3	3·1	3·0	2·8	2·2	1·3	0·4	0·1	—	—
	5·5	5·4	5·0	4·0	2·5	0·9	0·2	—	—
	8·6	8·5	8·0	6·8	4·7	2·3	0·9	0·3	0·1
	0·4	0·4	0·4	0·2	0·1	—	—	—	—
	1·9	1·8	1·6	1·1	0·5	0·1	—	—	—
10	4·8	4·6	4·0	2·9	1·5	0·4	0·1	—	—
	8·0	7·7	6·9	5·0	2·7	0·9	0·2	—	—
	11·8	11·4	10·4	8·0	4·9	2·1	0·8	0·3	0·1
	0·5	0·5	0·4	0·2	0·1	—	—	—	—
	2·3	2·1	1·6	0·9	0·4	0·1	—	—	—
30	5·7	5·2	4·2	2·5	1·1	0·3	0·1	—	—
	9·3	8·6	7·1	4·4	2·0	0·6	0·1	—	—
	13·5	12·6	10·8	7·2	3·8	1·5	0·5	0·2	0·1
	0·6	0·5	0·3	0·1	—	—	—	—	—
	2·4	2·0	1·3	0·6	0·2	—	—	—	—
100	6·0	5·0	3·5	1·6	0·6	0·2	—	—	—
	9·7	8·3	5·9	2·9	1·2	0·3	0·1	—	—
	14·1	12·4	9·3	5·0	2·3	0·8	0·3	0·1	—
	0·5	0·4	0·2	0·1	—	—	—	—	—
	2·3	1·6	0·9	0·3	0·1	—	—	—	—
300	5·7	4·2	2·4	0·9	0·3	0·1	—	—	—
	9·4	7·1	4·1	1·7	0·6	0·2	—	—	—
	13·7	10·9	6·9	3·0	1·2	0·4	0·1	—	—
	0·4	0·3	0·1	—	—	—	—	—	—
	1·9	1·1	0·5	0·2	0·1	—	—	—	—
1000	4·9	2·9	1·4	0·5	0·2	—	—	—	—
	8·2	5·0	2·4	0·8	0·3	0·1	—	—	—
	12·3	8·1	4·2	1·6	0·6	0·2	0·1	—	—

Arbitrary values for θ_1 and θ_2 are assumed, both θ_1 and θ_2 lying between 0 and $\pi/2$.

Since θ_1 and θ_2 deviate from actual solutions of the equations (9) and (10) these equations may be written as:

$$f_1\,[(\theta_1+\alpha, \theta_2+\beta)] = 0 \quad \ldots\ldots\ldots\ldots \quad (11)$$

$$f_2\,[(\theta_1+\alpha, \theta_2+\beta)] = 0 \quad \ldots\ldots\ldots\ldots \quad (12)$$

where α and β are corrections for θ_1 and θ_2 respectively.

By Taylor's theorem equations (11) and (12) are written as:

$$f_1 + \alpha\,(\partial f_1/\partial\theta_1) + \beta\,(\partial f_1/\partial\theta_2) = 0 \quad \ldots\ldots\ldots \quad (13)$$

$$f_2 + \alpha\,(\partial f_2/\partial\theta_1) + \beta\,(\partial f_2/\partial\theta_2) = 0 \quad \ldots\ldots\ldots \quad (14)$$

Let

$(\partial f_1/\partial\theta_1) = f_{11} = A\sin\theta_1$, where $A = 1+n_2+(n_1/\pi)\,\theta_1$

$(\partial f_1/\partial\theta_2) = f_{12} = -wn\sin\theta_2$

$(\partial f_2/\partial\theta_2) = f_{21} = -(n_2/w)\sin\theta_1$

$(\partial f_2/\partial\theta_2) = f_{22} = B\sin\theta_2$, where $B = 1+n_2+(n_1/\pi)\,\theta_2$

Equations (13) and (14) may be written as:

$$\alpha f_{11} + \beta f_{12} + f_1 = 0 \quad \ldots\ldots\ldots\ldots\ldots \quad (15)$$

$$\alpha f_{21} + \beta f_{22} + f_2 = 0 \quad \ldots\ldots\ldots\ldots\ldots \quad (16)$$

If $f_{12}f_2 - f_1f_{22} = C$

$f_1f_{21} - f_{11}f_2 = D$ and

$f_{11}f_{22} - f_{12}f_{21} = E$

then:

$$\alpha/C = \beta/D = 1/E,$$

that is:

$$\alpha = C/E \text{ and } \beta = D/E$$

If $\theta_{1(p)} = \theta_{2(p)}$ are the values of the θ_1 and θ_2 assumed, α, β the corrections, then the second approximations to θ_1, θ_2 are given by:

$$\theta_1 = \theta_{1(p)} + \alpha$$

$$\theta_2 = \theta_{2(p)} + \beta$$

With the new value of θ_1 and θ_2, the same process is repeated until α and β are sufficiently small to provide the required accuracy.

Thus θ_1 and θ_2 are calculated for a given value of x, n and n_2.

Calculations of θ_o

In a similar manner to the calculation of θ_1 and θ_2, θ_o can be calculated by Newton's method of approximation. If $\theta_{o(p)}$ is any value assumed $\theta_o = \theta_{o(p)} + \delta$ is a better approximation where:

$$\delta = \left[\frac{(\pi/n_1) + \theta_{o(p)} - \tan\theta_{o(p)}}{\tan^2\theta_{o(p)}}\right]$$

The process is repeated for a new value of θ_o until δ is sufficiently small.

Calculation of θ_c

θ_c is calculated in a similar manner. The correction δ to $\theta_{c(p)}$ is:

$$\delta = \left[\frac{\pi/n_1\,\dfrac{1+2n_2}{1+n_2} + \theta_{c(p)} - \tan\theta_{c(p)}}{\tan^2\theta_{c(p)}}\right] \text{from equation (5)}$$

Calculation of Non-linearity

Thus θ_1 and θ_2 of the diodes are calculated for a given value of x, n_1 and n_2. α_x and α_o are calculated using equations (6) and (2).

Percentage non-linearity of the detector at any value of x can be calculated by the equation.

$$N_x = \frac{\alpha_o - \alpha_x}{\alpha_o} \times 100 \quad \ldots\ldots\ldots\ldots\ldots \quad (17)$$

which is the general case of equation (1).

Results and Conclusions

Actual calculation of angles of conduction was made by the above method on a SIRIUS computer. The total time taken for completing the entire range of values shown in the Table 1, was about two hours. The error in the calculated angles of conduction is believed to be not more than a few seconds of arc and the non-linearity values are probably accurate to 0·1 per cent.

It was found that the maximum non-linearity for signal levels lower than x_c was much lower than for x_c itself. In all cases, the non-linearity is reduced from its value at x_c by a factor of not less than six for a value of $x = x_c/2$.

Thus the table provides new criteria for the design of a phase sensitive detector circuit.

Table 1 shows the performance of the detector at different values of x (viz) $0{\cdot}25\,x_c$, $0{\cdot}5\,x_c$, $0{\cdot}75\,x_c$, $0{\cdot}9\,x_c$ and at x_c itself. (The value at x_c is the same as given in the earlier paper referred to, thus providing a check on the present calculations).

Acknowledgment

The authors' thanks are due to Dr. S. Krishnan for suggesting the problem and for his ideas on interpreting the results.

REFERENCE

1. Chidambaram, R., Krishnan, S. Diode Phase—Sensitive Detectors with load. *Electronic Engg.* 31, 613 (1959).

A Take-off and Overshoot Director System

An aircraft instrumentation system to give aircraft pilots assistance during the take-off period has been produced by Elliott-Automation Ltd, and is the first stage in the development of automatic take-off and overshoot.

Designed to comply with all current aircraft procedures, as specified by A.R.B., F.A.A., and I.C.A.O., the system is intended to reduce the load on the pilot by indicating the optimum profile to fly under all conditions, and is now undergoing flight trials in an R.A.F. Vulcan.

Variations in aircraft weight, air temperature, wind velocity and airfield altitude all combine to alter the safety characteristics demanded by each take-off. This information is not available to the pilot from any other single source.

The system incorporates a computer which is programmed with the characteristics of the particular aircraft and is then simply up-dated for the existing take-off conditions. If engine failure occurs at any point in the take-off or climb, the system automatically computes a new optimum flight path for the pilot to follow.

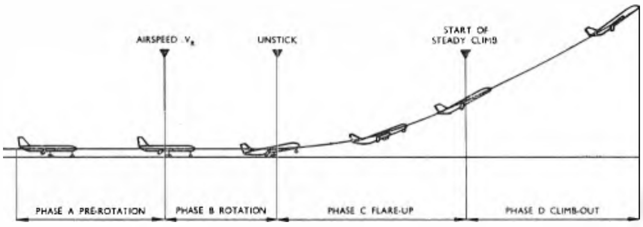
Another important facility is a 'noise abatement' climb path which can be selected by the pilot when operating from airfields in thickly populated districts. This is likely to be particularly useful in the case of supersonic aircraft.

The system now flying in the Vulcan directs the aircraft in the pitch plane only. Under development is a system giving directions to the pilot in all three axes and operating to the same safety standard as that required for automatic landing.

The procedure during a take-off is as follows:

While the aircraft is accelerating to the rotation speed, the pilot is directed to hold the nosewheel on the runway, by a 'nose-down' demand on the director display. As the rotation speed approaches he is given warning of this, as the computer linearly reduces the 'nose-down' demands so that the demand is nil at V_R. At rotation a chosen increase of incidence is now required. This is demanded rapidly so that the runway consumed during rotation is as little as possible. The directed rate is compatible with the rates currently 'pulled' by pilots without instrument direction, a 'nose-up' pitch angle is demanded which is backed off by pitch rate and actual pitch angle. The angle chosen and gains used are carefully chosen with the object of obtaining lift-off with a small but finite positive pitch rate which will allow the aircraft and director to move smoothly through into the flare-up. During this phase the aircraft will continue to accelerate.

At unstick the aircraft leaves the ground as indicated by the full extension of the oleo legs, this being electrically signalled by multiple micro-switches. This signal, which initiates the flare-up stage, is held for a short time after the attainment of V_R to cover the case of the aircraft 'hopping' on runway. The aircraft incidence is maintained at a moderately high value to give upward acceleration and is later reduced as the speed approaches V_{TOC} to give a smooth transition to take-off climb. The demand is for a constant pitch rate this being necessary since the airspeed will still be relatively near the stall. During the 'climb-out' a phase advanced speed control law is used, where the demanded speed is continuously computed to take account of any deterioration or marked in-



Take-off sequence

crease in performance over the normal standard of the aircraft in question. This speed has an absolute minimum value of V_2, the Take-Off Safety speed, but is normally significantly higher. This is achieved by adding to the V_2 speed demand, a rate of climb term with a preset bias which must be exceeded before the term becomes operative. On any take-off the rate of climb must exceed the preset bias before the term becomes effective, but if the rate of climb is very high the demand speed is correspondingly raised, thus avoiding undesirably high pitch angles during the take-off climb. In the event of engine failure under marginal operational conditions, particularly on a large jet aircraft, the rate of climb may not reach the preset bias and the aircraft will climb out at V_2. In all engine failure cases before V_2, simulator results show an improvement, when using the rate of climb term, in the distance to the 35-foot point. This automatic compensation for decreased performance is considered an essential part of the 'climb-out' law.

The system copes automatically with speed changes resulting from flap angle changes on the climb. Furthermore, the system has been designed to be unaffected by the position errors introduced into the pitot-static system due to ground effect and to the aircraft attitude changing.

In 'Overshoot' when the pilot opens the throttles the system will demand the pitch rate to achieve the correct flare which checks the rate of descent and places the aircraft on the optimum climb-out path. This will apply whether the aircraft has already started to flare or not. A simple extension of the control manoeuvre demanded for the flare-up and climb out phases of the take-off climb is used and the same automatic allowances are included for such effects as flap changes and engine failures. This system may be switched on prior to approach to land or else left on continuously for the complete flight from take-off to landing.

Short News Items

A one-day Symposium organized by the Tees-side Section of the Society of Instrument Technology and entitled 'The Future Pattern of Research in Instrument Technology, is to be held on 14 October at the Constantine College of Technology, Middlesbrough, Yorks.

Further details are obtainable from G. F. Shute, 10 Mayberry Grove, Linthorpe, Middlesbrough, Yorkshire.

The Science Research Council has made a grant of approximately £29 000 to the University of Essex for research on plasma heating by absorption of laser light by Professor A. F. Gibson. The grant will run from 1 October 1965 to 31 July 1967.

The research programme, of considerable interest both in the fields of lasers and plasma physics, will be concerned with the further development of high power laser techniques and their application to the production of plasmas. The mechanism of plasma generation, particularly using solid targets, will be investigated by several methods to find out how best to use lasers to produce really hot plasmas.

The BBC Research Department has produced an electronic store unit for delaying a television video signal by 20msec—the time taken to scan a complete 405 or 625 line field. The equipment will be used in experimental projects including work on standards conversion and film recording. In addition, the design experience gained will be utilized in the future construction of similar delay equipment.

The broad bandwidth required and the comparatively long delay is obtained by using quartz ultrasonic delay lines. Eight 2·5msec lines have been supplied by The M.E.L. Equipment Company Ltd, and are being used in cascade by the BBC to obtain the total delay of 20msec.

Each line is enclosed in a double skin oven whose temperature is maintained accurately at 70°C. The elevated temperature, in addition to being easier to control, tends to lower insertion loss and improve the bandwidth characteristic.

The equipment will be completely transistorized and will incorporate carrier modulators, demodulators, intermediate delay line amplifiers, variable mercury delay lines for fine adjustment and other features designed and built by the BBC.

The Port of Bristol will, early next year, become the first port in Britain, and possibly in Europe, to operate an on-line traffic control system.

Between the Avonmouth Docks and the Port of Bristol Authority's head offices in Queen Square, Bristol, there is to be operated an automatic data transmission link by which details of every ship's cargo arriving at Avonmouth and its subsequent discharge will be controlled on an NCR 315 computer equipped with random access memory files.

As soon as each vessel arrives at Avonmouth, details of the cargo are punched into paper tape which is then fed into a special transmitter unit. This transmits data over a GPO line to the Port's computer centre in Queen Square, Bristol, where an identical tape is created. The tape is then fed into the NCR 315 computer which stores the information on its random access memory files.

James Scott (Electronic Engineering) Ltd of Glasgow have designed and developed on behalf of the Ministry of Aviation a complete automatic equipment for analysing the effect of rocket motor flame on radar signals.

Radar signals transmitted from the ground to a missile or vice versa will in most cases have to pass through the rocket motor flame path. This will have the effect of both attenuating the signal and also inducing spurious noise modulations. The characteristics of different types of motor will vary and it is therefore essential that the radar system's designers be provided with this information.

The 'Allscott' equipment consists basically of a transmit/receive system with facilities for accurately determining the attenuation of the appropriate signal frequencies by the flame, combined with a multi-channel high speed recorder and also incorporating a specially designed analyser having extremely high sensitivity capable of detecting low order noise modulations induced by the flame.

Interest has been shown in this equipment by US rocket motor design authorities.

The British Scientific Instrument Research Association (SIRA) has established a new enquiry service to be known as SIRAID covering the whole range of British commercially available instrument equipment.

The service is designed to cover enquiries on:

Thin film technology and microelectronics
Electronics and electromechanical systems
Optical systems and optical component design
Precision mechanics
Electron microscopy
Scientific and industrial measurement techniques
Industrial applications of measurement and control
Data processing
Performance evaluation of industrial instruments

Enquiries should be sent to SIRAID, South Hill, Chislehurst, Kent. (Tel.: IMP 0055.)

Smiths Aviation Division announce that the 15 Hawker Siddeley Trident 2 aircraft recently ordered by BEA will be fitted with their Series 5 Flight Control System. This is the Autoland System designed by Smiths to provide fully automatic blind landing facilities for all the BEA Trident fleet.

The Series 5 system, already fitted to the BEA Trident 1 aircraft, was recently certificated by the Air Registration Board for Autoflare and subsequently made the world's first passenger carrying automatic landing at London Airport on a scheduled flight from Le Bourget.

Valradio Ltd of Feltham, Middlesex, have received an order valued at £25 000 from Singapore for filtered power supplies, for use in conjunction with the Cossor v.h.f. radio telephone equipment Type CC. 15 ABWX6.

The application is for 12V mobiles to operate as fixed stations using filtered 12V d.c. direct from a.c. source with Type 25-RU/12/10.

The Rectifier Equipment Division of Standard Telephones and Cables, Harlow, Essex, has just shipped to Algeria nineteen transformer/rectifier equipments to be used to prevent corrosion of a 500 mile oil pipeline nearing completion.

The new pipeline, owned by the Algerian Government Agency SONATRAH, runs from a reception tank farm at Haoud El Hamra, over a mainly desert route, to the coast terminal at Arzew, near Oran.

Five of the equipments for the new pipeline provide 100 amperes and fourteen have 50 amperes outputs at voltages between 7-48V. Coarse, medium and fine controls on the equipments give 63 steps of output voltage control. The equipments are of standard design electrically, but have extra oil cooling convection tubes to prevent overheating in tropical conditions.

The equipments will not be positioned at equal distances along the pipeline. Most will be operating along the northern one third of the route which is part-mountainous and in the rainier areas. Two thirds of the route is desert terrain.

Care has been taken to avoid electrical cross-intereference with the existing pipelines installed earlier by French companies which are already protected by cathodic protection systems.

Skanska Cementaktiebolaget, Sweden's largest cement producer, has awarded a contract worth substantially more than £500 000 to the G.E.C. Ltd of England, for comprehensive computer control and centralized monitoring and control systems for its plant at Limhamn near Malmo. International System Control Limited, Wembley, the G.E.C.'s computer automation subsidiary will supply the major part of the scheme.

The order covers the entire automation and control sections of a major modernization programme for this cement plant which will ultimately produce over 1 million tons of cement per year.

Equipment will be provided for two control rooms one of which will control the cement production plant and the other the quarrying area and new raw materials blending plant some 2 kilometres away from it.

A particular feature of the contract is a computer control system for on-line regulation and optimization of the five wet process kilns and also control of the materials blending operation which will incorporate an on-line X-ray analysis.

The BBC has awarded a contract to The M.E.L. Equipment Company Ltd for the supply of four 'fill-in' u.h.f. television translator equipments for operation on the BBC-2 service. This follows an earlier order for prototype equipment which it is proposed to install in the Guildford area.

The translators handle both the video and sound components together as one broadband signal, and retransmit at a peak synchronizing level of 1kW. The equipment is entirely transistorized with the exception of the final klystron amplifier.

Siemens-Schuckert have recently opened their Research Centre at Erlangen, Bavaria. Built at a cost of approximately £10M, it becomes Europe's largest private enterprise research establishment covering 82 acres.

The main emphasis at the centre is in the fields of semiconductors, superconductors, plasma physics, direct conversion of energy (fuel cells, magnetohydrodynamic generators, etc.), automation, control engineering and reactor development.

The Ghanian Corvette GNS Keta, built by Vickers Armstrongs (Ship Builders) Ltd, has been fitted with a wide range of Redifon communications equipment.

A special feature of the installation is the Redifon circuit control exchange which allows all communications equipment to be remotely controlled from the wheelhouse, bridge or operations room, as well as from the wireless office.

The principal item of the installation is a 602E m.f./h.f. main/emergency transmitter but Redifon also supplied an MR.604 MF receiver, an R.50M main receiver and an R.141 combined emergency and loudspeaker watch receiver. U.H.F. facilities are provided by an A43R 2W transmitter/receiver.

Elliott Traffic Automation Ltd have received a £30 000 contract from the Stevenage Development Corporation for the supply of a Central Car Park Information System to be installed in Stevenage Town Centre.

Motorists approaching the town centre, under this new system will observe internally illuminated signs giving directions to parking areas with parking space availability clearly shown.

Vehicles entering and leaving the seven car parking areas will be monitored and the net totals of vehicles in the car parks will be transmitted in binary code to a Central Processor which incorporates computing technology which will, in turn, control the signs. The signs themselves, will give information in a positive fashion only, making it easy for the motorist to understand the direction he is to follow if he is to find parking space without delay.

Digital read-out facilities will be provided in a central Control Room, for the continuous checking of cars parked in all parking areas at any one time. In addition, for statistical and survey purposes, cumulative totals will also be given separately for each car park, over a period that can be pre-set, according to the requirements.

Elliott-Automation Ltd have received two orders, together worth £300 000 for their ARCH automation systems, one from Russia and the other from East Germany.

The Russian order is for the largest ARCH 8000 system ever supplied by the Company. It is to be installed in a large existing ammonia plant in South Russia and will provide on-line computer control for the complete production process. The central processor of the system is an Elliott 803B computer, with 8 192 words of core-store backed by two magnetic film memory units. The ARCH system as a whole will be capable of handling five hundred and twelve analogue plant signals as well as a large number of digital signals and signals from a number of chromatographic instruments. The system has both line printer and graph-plotter output channels. The physical control of the processes in the plant will be implemented through thirty-five control loops, using set-point adjustment techniques incorporating ARCH LINK equipments. The complete system is to be shipped to Russia early next year.

The order from East Germany is for an ARCH 2000 system, of which the central processor is an Elliott 4120 computer with a 16 000 word core-store memory. The entire system is to be mounted in a mobile trailer so that it can be moved from place to place to work on a number of projects in the Schwedt oil refinery.

The ARCH 2000 systems, which makes exclusive use of silicon solid-state devices, will handle four hundred analogue and three hundred digital plant input signals and provides twenty analogue, one hundred digital and twenty stepping motor output channels for implementing the control functions.

An operational Research Techniques course starting on 11 October and lasting nine weeks is being organized by C.E.I.R. Ltd. The course itself, will last for six weeks but will be preceded by three weeks of background work on mathematics and statistics.

Among the subjects studied will be: linear and non-linear programming; dynamic programming; critical path methods; queueing theory; simulation; forecasting and time series analysis; data processing; computer; costing and accounting; stock control.

Several allied subjects will also be discussed, and considerable time will be devoted to detailed case studies and exercises.

Copies of the programme booklet are obtainable on application to S. Batten, Training Manager, C-E-I-R Ltd, 30/31 Newman Street, London, W.1.

The Plessey Company Limited of Ilford, Essex and the Bissett-Berman Corporation of Santa Monica, California, have announced a distribution and manufacturing cross licensing agreement in the oceanographic and meteorological fields.

The agreement will extend the expanding product range of each company and provide oceanographers and meteorologists with an advanced range of instruments.

Plessey will be the exclusive distributor of the Bissett-Berman "Hytech" oceanographic and meteorological sensors, instruments and systems in the British Commonwealth, Holland, South Africa, Norway and Sweden. Bissett-Berman will be the exclusive distributor in the United States of Plessey oceanographic and meteorological products.

The Printed Circuit Group of the Institute of Metal Finishing is to hold a one day symposium on printed circuit techniques at Kimbells, Osborne Road, Southsea on 7 October.

Full details are available on application to P. G. L. Vivian, 20 Bowling Green Lane, London, E.C.1.

LETTERS TO THE EDITOR

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

A Velocity Controlled Servo System Using Stepping Motors

DEAR SIR,—Many laboratory problems require the use of mechanical servo control, in an electronic or physics laboratory usually the stumbling block is obtaining mechanical gear boxes and motors which are compatible. Mullard Ltd manufacture a number of stepping motors with a very wide range of gear-boxes to which the motors can be fitted. The following describes a method of using stepping motors for servo control.

A circuit diagram of the control system is given in Fig. 1; it consists of three basic parts, a standard operational amplifier (Fenlow Electronics Ltd, type A2), a comparator and monostable resetting circuit and the motor drive circuit (Mullard Ltd).

A method of control which is stable for many mechanical loading conditions is to use the error signal to control the velocity of the output shaft. To control the stepping motor in this way the error signal must be used to produce a train of pulses, the frequency of which depends upon the magnitude of the error and the pulses should appear on one terminal for a positive error and another terminal for a negative error. This is achieved in the control system described by using the operational amplifier as an integrator, the error signal causes the amplifier output to increase linearly in either a positive or negative direction according to the sign of the error, on reaching a given level, one or other monostable is triggered and the integrator is reset (by the transistor gates, either VT_1 or VT_2 according to the sign of the error). Waveforms at the output of the amplifier are shown in Fig. 2.

The motor drive circuits are arranged to drive the motor in one direction when triggered on input *A* and the other when triggered on input *B*, by triggering these points from the monostables the desired velocity control is obtained.

The effective gain of the system can be varied by altering the integrating capacitor *C*.

Using the Fenlow amplifiers type A2 and A3 accurate servo control systems have been worked from thermocouple outputs of 2·5mV.

Yours faithfully,

M. BAKER,

Weybridge, Surrey.

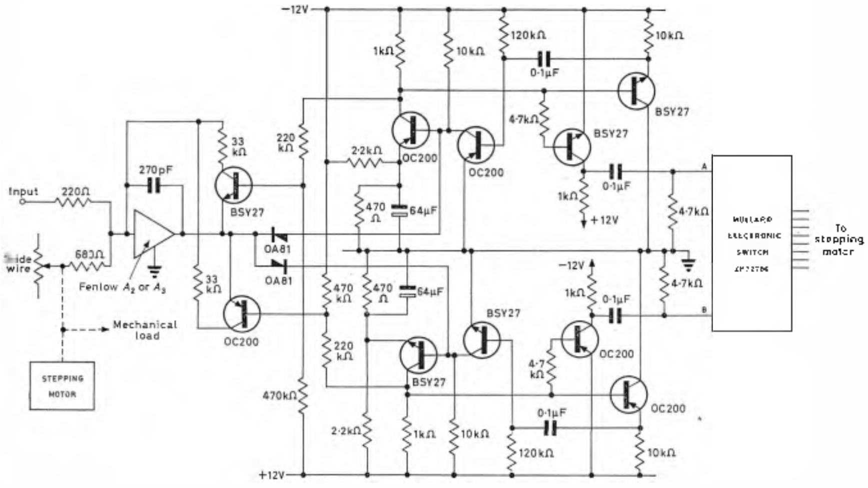


Fig. 1. The control system

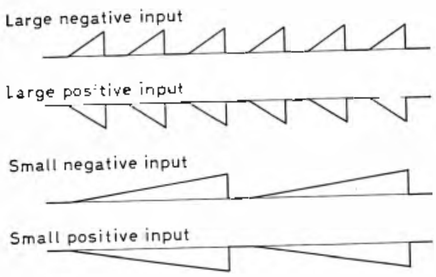
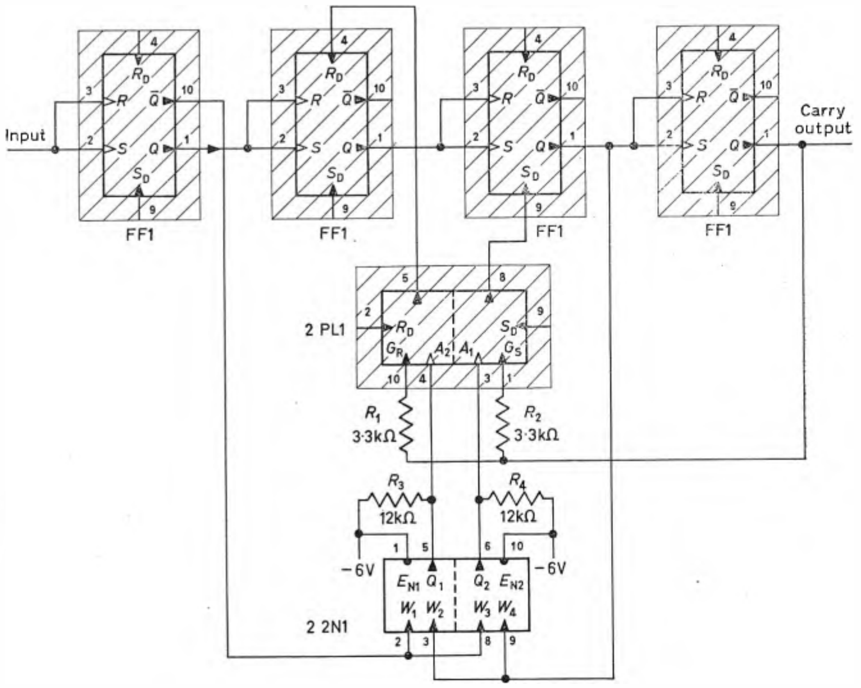


Fig. 2. Waveforms at the amplifier output

A Reversible Decade Counter

DEAR SIR,—As suppliers in the U.K. of the Philips Circuit Blocks mentioned under reference 3 in Mr. I. Scollar's article 'A Reversible Decade Counter with Symmetrical 1242 Coding' in the July 1965 issue we are interested in the ideas and principles put forward but would like to indicate some ways of making the detailed circuit more reliable.

Fig. A. Symmetrical 1242 decade



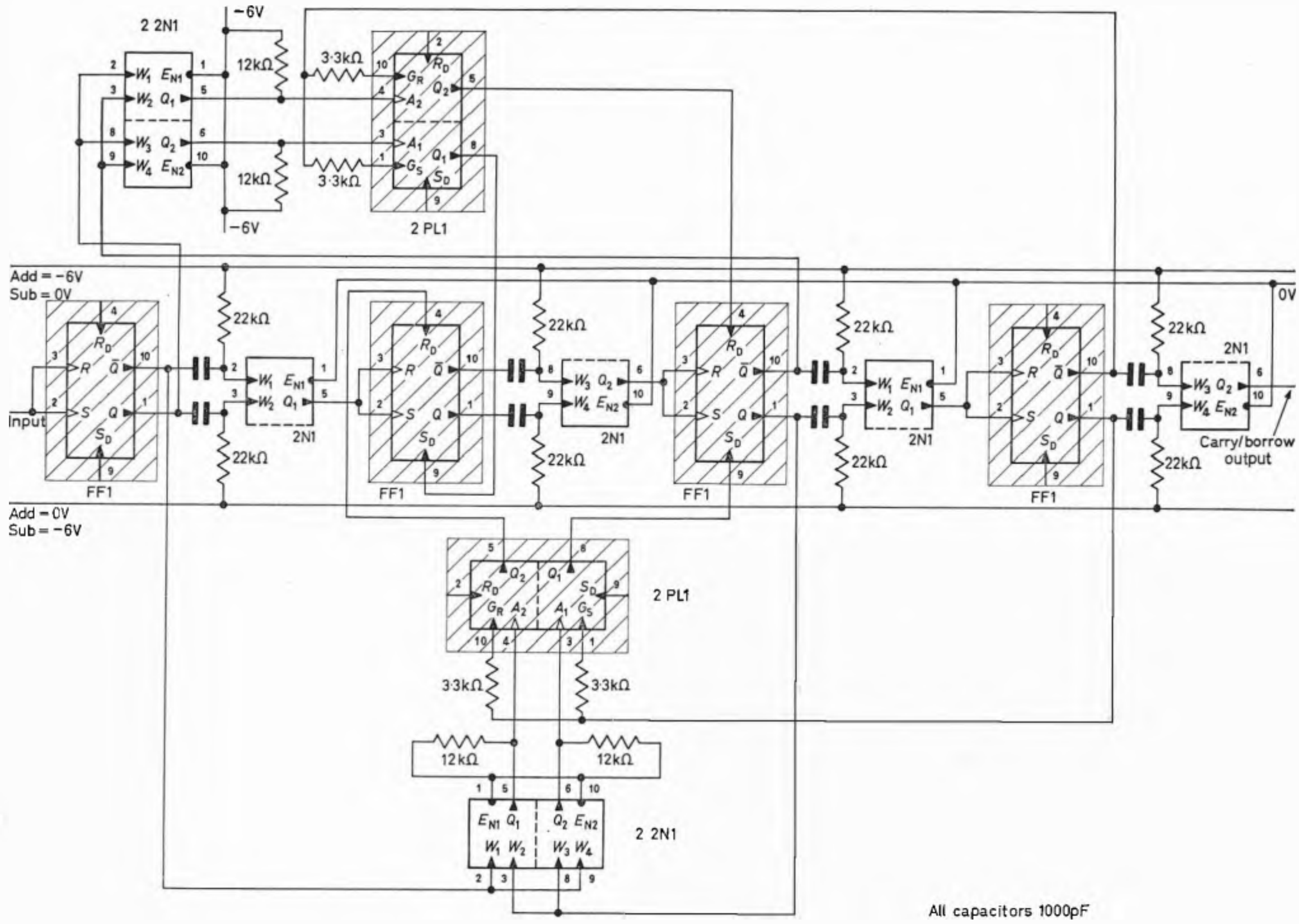


Fig. B. Reversible decade counter

Fig. 1. In this circuit a 12kΩ source impedance (3N1) is feeding two flip-flop d.c. inputs both requiring a minimum of 0·5mA for reliable operation. Also, if transistor 6 happened to be a deal faster than 3 due to production spreads or different wiring capacitance, etc., then the driving signal would be removed before 3 had switched properly. This type of circuit is unreliable and should never be used. (R. K. Richards[3] copes with this situation by calling for delays in all lines feeding the AND gate.) A reliable circuit to produce the same code is shown in Fig. A. This can be used at 30kc/s without R_3 and R_4 and 100kc/s with them included.

The maximum counting rate for these blocks is 100kc/s under worst case conditions, thus claims of higher rates can only occur because of the built-in safety margin. However, for reliable operation of circuits under production conditions, this figure should not be exceeded.

Fig. 2. The combination N-P-EFI-FFI (a.c. output) is not normally to be recommended particularly when loaded by the matrix in Fig. 3 and a more reliable reversible counter embodying the circuit from Fig. A is shown in Fig. B. This cannot drive the matrix in Fig. 3.

The changeover speed is 60μsec in the case shown. This can be decreased to 12μsec by using the more complex coupling circuit shown in Fig. C.

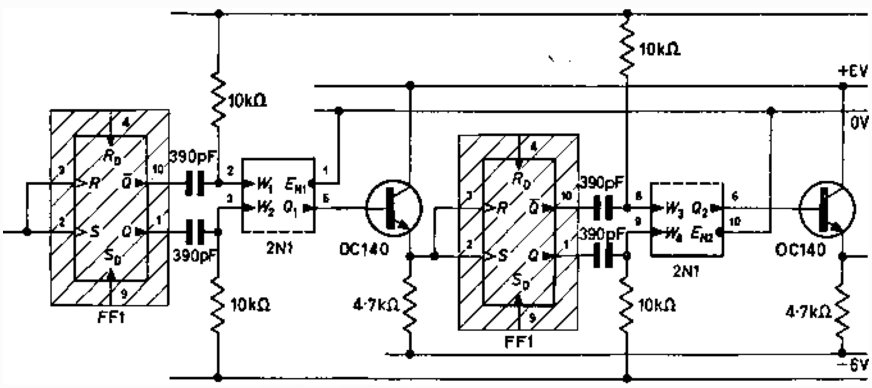
Therefore, while we agree with the ideas and principles put forward in the article, we cannot recommend the use of the actual circuits except possibly under laboratory conditions. It is also worth mentioning that although limited to 30kc/s, the new MEL range of circuit blocks known as the '10 series' can be used for this type of application with practically no external components at all.

Yours faithfully,

B. G. STARR,

The M.E.L. Equipment Co. Ltd.

Fig. C. Fast reversible counter



The author replies:

DEAR SIR,—The writer is grateful to the Philips organization and to Mr. Starr for calling attention to several weaknesses in the design of reversible counters when using the components supplied by them.

In all counters, 3·9kΩ resistors between the negative supply and the output of each 3N1 gate were inadvertently omitted in the drawings. These provide the current required as stated in the second paragraph of Mr. Starr's letter. With regard to the production tolerances in the switching times of the flip-flops and the dangers of premature gate closure, perhaps Mr. Starr would be good enough to provide this data in statistical form. However, to the switching time must also be added that of the gate itself and associated wiring. Experimentally, it was necessary to place 2 800pF of additional capacitance at the base of the third transistor in order to produce malfunction due to delay. The value is far in excess of the usual wiring capacitance. Hence the undersigned cannot agree with Mr. Starr's conclusion that the proposed configuration "should never be used".

Mr. Starr is quite correct in pointing out that the combination of N-P-EF1 plus additional resistor used between the flip-flops in the reversible counter cannot drive the matrix reliably. An additional set of four emitter-followers has to be interposed to prevent excessive loading. This, unfortunately, was not clearly expressed in the text. It may be here pointed out that fifteen decades of the type described have been in service in five field instruments on three continents for two years without error, even under considerable extremes of temperature. However, the reversible counter shown by Mr. Starr is, in some respects, superior in that it uses fewer elements. When using the extra amplifiers of his Fig. C. it is also capable of more rapid reversal. The matrix presumably could be connected via isolating emitter-followers to one set of flip-flop collectors though, as noted in the undersigned's article, negatives will then be indicated in complement form. Because a.c. coupling is used by Mr. Starr, switching automatically to the opposite collector set requires eight additional gates driven from the sign recognition circuit, increasing the number of elements to one more than that used by the undersigned.

It is not every company which takes such great interest in the application of its products and Philips is to be congratulated for the concern which has been shown in this instance.

Yours faithfully,

IRWIN SCOLLAR,

Rheinisches Landesmuseum,
Bonn.

D.C. Voltage Level Detection

DEAR SIR,—In recent circuit investigations into three-state (ternary) logic circuits, the need was encountered for circuits which would respond sharply to preset levels of an input d.c. voltage. A typical problem was for a circuit whose output voltage was maximum when the input was below a preset level and zero when the input became just greater than this level. The circuit arrangement produced is as detailed below and, as this is thought not to have been previously published, may therefore be of interest to your readers.

For the above typical problem, it requires that a common-emitter transistor be held fully 'off' when the input

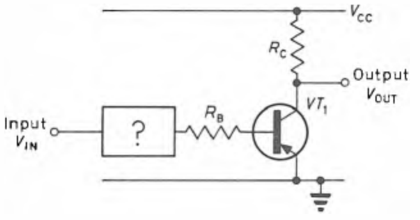


Fig. 1. Circuit to satisfy the following conditions:
$V_{OUT} \simeq V_{CC}$ *when* $V_{IN} < V_X$
$V_{OUT} \simeq 0$ *when* $V_{IN} > V_X$
Where V_X *is some preset voltage between 0 and* V_{CC}

voltage is below the preset value, and switched fully 'on' when the input voltage exceeds this level (see Fig. 1).

A conventional Schmitt trigger circuit as the input voltage detector is unsatisfactory due to the inherent backlash between the switching 'on' and switching 'off' input voltage levels, and also because the circuit d.c. output voltage level swings are unsuitable to switch 'on' and 'off' the final transistor VT_1 without resorting to an extra d.c. supply plus potential divider network.

A series Zener diode is also impractical, both operationally and also from the point of view of inconvenience when modifying the preset level voltage. Its operational impracticability stems from

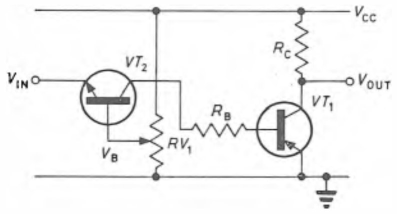


Fig. 2. Practical circuit arrangement

the fact that there is a rather wide range of input voltage during which the output circuit is neither fully 'off' nor fully 'on', the input voltage minus the Zener breakdown voltage being inadequate to provide the base input current required to saturate the output transistor VT_1.

The circuit solution found satisfactory is as shown in Fig. 2, and utilizes one transistor only of opposite polarity to that of the output transistor.

The action of this circuit is very simple but possesses several attractive properties on closer investigation. The complementary input transistor VT_2 will, of course, be fully cut off as long as V_{IN} is less than the base potential V_B controlled by RV_1, but as soon as V_{IN} becomes slightly greater than V_B, VT_2 will conduct. For a very small rise in V_{IN} above V_B, VT_2 will become fully saturated, and thus in effect virtually the full input voltage V_{IN} is transferred to the collector of VT_2. Provided R_B, R_C, and the large signal current gain of VT_1 are appropriately proportioned, this appearance of V_{IN} on R_B will be sufficient to fully saturate VT_1 and hence provide the rapid collapse of output voltage from V_{CC} to zero.

If V_{IN} continues to rise above the initial conducting level, it will take the base voltage V_B with it, with no effect on VT_1. RV_1 can be a high resistance potentiometer, say 50kΩ, but if preset for a low voltage V_B a series resistance between RV_1 and VT_2 may be desirable.

While silicon transistors may be advantageous in position VT_1 to maintain minimum 'off' collector current without any extra biasing circuit, germanium transistors may conveniently and advantageously be used in position VT_2, as this transistor is automatically biased 'off' when negligible current flow is required through it. The use of germanium transistors in this position is also economically attractive, being less costly, for example, than Zener diodes which might otherwise be employed. The gain performance required in position VT_2 is minimal, which again adds to the financial attractiveness of the circuit.

In general, the performance of the circuit is such that the full supply voltage output swing is obtained from VT_1 for a change of input voltage ΔV_{IN} of less than 0·2V about the critical preset value. No backlash effects are, of course, present, the circuit switching off at the same voltage level at which it switched on. With the ease of control of the preset level and this small value of ΔV_{IN}, various applications other than the one originally required can be suggested, for example, over-voltage or under-voltage detection circuits, analogue-to-digital conversion systems, and the like.

One of the author's colleagues has also suggested that RV_1 may be supplied from the collector of VT_1 instead of from the d.c. supply rail V_{CC} with advantageous results in over-voltage detection applications. When this is done, ΔV_{IN} is reduced from 0·185V in one particular circuit, to $<$10mV for full swing of output voltage V_{OUT}. The circuit is now a 'latching' type circuit, as V_{IN} has to be reduced to a low value before VT_1 switches off again. Multivibrator arrangements of this circuit are also possible when suitable resistor-capacitor networks are placed in the input to VT_2, but these do not appear to offer any practical advantages over conventional multivibrator circuits.

Yours faithfully,

S. L. HURST,

Bristol College of Science and Technology.

BOOK REVIEWS

Data Transmission

By William R. Bennett and James R. Davey. 356 pp. Med 8vo. McGraw-Hill Book Co. 1965. Price 116s.

THE authors establish in this book the principles of data communications and the basic concept of digital data messages. Various transmission methods are described in detail, with particular emphasis upon the shaping of data signals to achieve maximum tolerance against noise. Theoretical error rates are given for specific kinds of modulation. The effects of transmission impairments commonly encountered are described, with particular attention to the limitations of the telephone noise channel. Systems are compared as to their bandwidth efficiency, their tolerance to noise and to other transmission impairments. The practical problems of coherent detection, synchronization, line equalization and error control are all discussed. A section is included on various means for measuring the operational performances of systems, and the concluding chapters deal with optimization theory and statistics of digital signals.

The need has been great for some exposition of data transmission and for an unbiased comparison of the various methods of modulation, with their own inherent limitations under practical conditions. The authors, through their professional capacities at the Bell laboratories, have largely been able to fulfil this need with their book. It is felt, however, that the contents would have found far wider appeal if the practical approach to the subject had been employed earlier in the book.

The mathematical analyses, which form the major part of the book, will be beyond many readers; but with all due respect to the authors they do state that the contents are "intended to be comprehensible to those who have the equivalent of a bachelor's degree in electrical engineering", and that "some mathematical maturity is expected". The material for the book was first presented as a series of lectures and this is apparent while reading, for a fair amount of repetition of the same subject matter does occur in various chapters. Anyone employing the mathematics expounded, or comparing these against established expressions, should watch the definitions carefully.

As a book attempting to cover all aspects of data transmission it is a pity that some attention has not been given to the useful work that has been carried out elsewhere than in the U.S.A. The book would have been more complete if it had

(a) Extended its considerations of system performances in the presence of intersymbol interference. This is touched on but is then hastily dropped.

(b) Devoted more space to line error control, because regardless of which optimum modulation method used these line errors will still occur.

(c) Devoted at least a chapter to the additional transmission impairments encountered on h.f. radio links.

Despite these shortcomings, this book could become a standard reference for data transmission.

W. F. S. CHITTLEBURGH

Design Theory and Data for Electrical Filters

By J. K. Skwirzynski. 701 pp. Med. 8vo. D. Van Nostrand & Co. Ltd. 1965. Price £8 10s.

THIS book, perhaps with some loss of rigour, introduces the reader to the fundamentals of insertion-loss filter design technique, studies the behaviour of various types of filters, the effects of dissipation in the reactive components comprising the filters and group delay characteristics of the filters. However, it is basically written to introduce the tables of design parameters which enable one to design filters to specifications of some given forms.

In his preface, the author points out that for an immediate use of the tables it is sufficient to read only chapter 10. The reviewer was interested to find that an engineer with some general ideas about communication networks and filters could learn to use these tables to design filters by just reading chapter 10 and studying the examples given in chapter 11. As chapter 10 comprises only 32 pages, this facility should appeal to engineers who design filters only occasionally. As such, chapter 10 is well presented (as is also the remainder of the book) and quite readable.

In trying to provide generalized tables for insertion loss design, the treatment in the book imposes two limitations on the range of filters considered.

(1) Filters with different minimum-loss requirements in different parts of their stop-bands are not considered.

(2) As the band-pass and band-stop filters are developed by means of frequency transformations from a basic low-pass filter, the design data for these types of filters apply only to filters having loss specifications which have geometrical symmetry along the frequency axis.

Within these limitations the book should be a useful handbook to non-specialist filter designers, while serving as a useful introduction to the topic for engineers intending to specialize on filter design.

There are a few minor points which seem worthy of mention:

(1) 'Discrimination of a filter' is normally used to indicate the difference between the losses introduced by the filter at the unwanted and wanted frequencies. It may have been better to have used a different expression for the so-called 'discrimination loss', which is a completely different concept.

(2) 'Insertion loss' is normally defined on the basis of voltage or current ratios. Especially in a text including a study of the effects of open- and short-circuit terminations (as well as finite resistive loads) in filters, it might have been less ambiguous to avoid the use of the insertion power loss concept in the derivations, particularly in chapter 10.

(3) Perhaps more serious is the statement (p. 250) that 'when R_2 becomes infinitely large . . . the discrimination power ratio is then replaced by the square of the corresponding voltage ratio, $|E_0/V_2|^2$. This statement is valid, surely, for the *insertion* power ratio.

H. O. BERKTAY

Electricity and Magnetism

By W. J. Duffin. 452 pp. Demy 8vo. McGraw-Hill Publishing Company Ltd. 1965. Price 45s.

THIS basic textbook, covering material for the first two years of an honours degree physics course, contains much of value to electrical engineering students and parts will interest school sixth-formers.

Basic concepts of charge and current are first established. Four chapters are then devoted to electric forces and fields and three to corresponding magnetic phenomena. The network theory presented goes further than usual in a book of this standard because of its relevance to electronics. An introduction to charged-particle dynamics shows modern applications of a classical subject and illustrates the confirmation of laws established for induction currents. Conduction, dielectrics and magnetic materials are treated in detail. The laws established earlier in the book are gathered together towards the end and generalized in Maxwell's equations, thus providing the starting point for more advanced study. Finally, measurements, standards and units are discussed. A chapter-by-chapter conversion from m.k.s. to c.g.s. units is included.

Material is frequently presented which is relevant to other studies, particularly atomic, nuclear and solid-state physics.

Each chapter starts with an outline of its aim and subject matter and ends, where appropriate, with a summary and discussion of results obtained. Early chapters have appendices on mathematical topics which may not have been covered at school, e.g. vectors, polar co-ordinates, partial differentiation, solid angles. Supplementing the text are 239 carefully chosen problems with answers and hints. References for further reading and to original works are given.

Many experiments establishing or confirming basic laws are described, new concepts are clearly introduced and defining equations are so labelled. Points frequently glossed over in elementary texts are examined; some standard treatments are discarded in favour of more satisfactory ones. An excellent choice of material has been made but there are omissions, e.g. thermoelectric and chemical effects and electrical machinery receive scanty attention while electronic circuits are omitted.

The stimulating and reasonably-priced book can be highly recommended to the students for whom it is intended and will probably become a well established textbook.

F. A. Benson

Quantum Electron Theory of Amorphous Conductors

By Aleksandr Ivanovich Gubanov. 277 pp. Med 8vo. Consultants Bureau Enterprises. 1965. Price $17.50.

The main task the author has set himself is the presentation of theoretical models which give qualitative explanations of the electrical properties of liquid and amorphous conductors. Attention is concentrated specifically on bulk effects associated with the distribution of electron energies; ionic conduction and the defect state are expressly excluded. The models are developed from the quantum electron theory of crystalline solids, and for completeness a concise treatment of this is presented in chapter 2.

Chapters 1 and 3 respectively are reviews of the relevant electrical properties of liquid and amorphous conductors and of their structure. In chapter 4 a liquid is considered as a one-dimensional chain of atoms in which there is short-term but not long-term order: it is shown that bands of electron energies can be predicted, as for crystalline solids. In the next chapter a more complicated three-dimensional model is developed. Chapters 6 and 7 are devoted to electron-scattering by long-range atomic disorder and by phonons respectively. The results are applied in chapter 8 to the discussion of electrical conductivity, thermoelectric power and galvanomagnetic effects. In chapters 9 and 10 an electron theory of liquid metals and its extension to electrical conduction are considered. In the final chapter a quasi-classical theory of amorphous ferromagnetism is presented.

The book is written at an advanced level for the specialist. The author presents his own point of view, but is fair in drawing attention to divergent views and in his comments on them. It will undoubtedly be many years before a universally-accepted theory evolves.

Considering that the text is translated from the Russian it reads very well. In the main, the mathematics, of which there is plenty, is well presented. There are, however, some disparities between symbols used in equations and in the text and there are some inappropriate choices of founts.

It seems unlikely that this book can have appeal to electronic engineers or indeed to many applied physicists in view of its theoretical bias. It is nevertheless a valuable contribution to scientific literature as it is the first monograph to deal with the physics of amorphous semiconductors.

F. J. Hyde

Electronic Transformers

By H. M. Nordenberg. 298 pp. Med. 8vo. Reinhold Publishing Corporation and Chapman and Hall Ltd. 1964. Price 108s.

Directed to electronic-equipment engineers and transformer designers this book is intended to be a complete guide to design, construction and application of electronic transformers. The author claims that readers will find the latest transformer design, testing and manufacturing techniques. Military transformers have been chosen for exemplification. The title is misleading for the book deals mainly with audio power and pulse transformers for use in electronic equipment.

Over 20 per cent of the book is concerned with characteristics, advantages and limitations of various types of transformer materials including data on dielectric strength, dielectric constant and losses, corona, impregnants, encapsulation, insulation to ground, core shapes and optimum dimensions and special environmental effects such as very high temperatures, high pressures and ionizing radiation. Specific types of transformer fabrication are recommended. Hermetically-sealed and oil-filled transformers are described. Relative merits of gas and liquid filling materials are discussed including the use of fluorochemical liquids.

A completely worked-out design procedure is presented for power transformers. Other chapters are devoted to general information on transformers, mechanical, environmental and electrical testing, specification and selection and faults. Experimental transformers for operation at very high ambient temperatures (350°C and 600°C) are described. Future developments resulting from improvements of present-day materials and processes are briefly discussed. References are given with each chapter.

The interests of the author, who is Head of the Electronics Parts and Assemblies Unit, U.S. Bureau of Ships, seem to have been closer to specification and testing than to design for a considerable time. As a result, probably, the chapter on general transformer information cannot be recommended. It is occasionally misleading and there are obvious omissions and misstatements. C.G.S. units are used with a short badly-explained statement on the m.k.s. system. Transformer designers will find some valuable information in this up-to-date book, which contains a wealth of data (also some printing errors), but they should look elsewhere for guidance in understanding its basis and will discover that much of the material presented is treated superficially.

F. A. Benson

Traveling Wave Antennas

By C. H. Walter. 429 pp. Med. 8vo. McGraw-Hill Book Co. 1965. Price 156s.

Dr. Walter is an Associate Professor at Ohio State University. He has written this book as part of a post-graduate course in aerial design and therefore assumes that the reader is conversant with field theory up to degree standard. Engineers slightly less familiar with mathematics will still find much of interest, thanks to the clear writing and ample references.

The scope of the book is narrower than the title suggests. Certain classes of travelling wave aerial, such as dielectric rods and surface wave structures, are covered in considerable detail. Others, such as resonant slot arrays, are given little space in spite of their greater practical value. Logarithmic aerials are dismissed in two pages and Yagis in three.

The first chapter opens with definitions and a discussion of reciprocity. Chapter two deals with basic radiation theory, starting from Maxwell's equations and including near field, Fresnel region and far field approximations. Chapter three discusses the synthesis of source distributions to produce given radiation patterns and chapter four theoretical aspects of the choice of radiating structure. Chapters five and six review the design theory of linear and planar sources respectively. Chapter seven deals with methods of exciting or feeding travelling wave aerials. Chapter eight, the longest chapter, is titled 'Practical Traveling Wave Antennas' but in fact stops well short of the practical design stage.

The theoretician will find in this book a useful collection of material from many sources, not all readily accessible in this country. The practical designer will gain insight, but will probably find it more convenient to cut and try than to compute the optimum parameters for his aerial from the formulae arrived at the design chapters. This, however, could be said of far too many other books.

M. F. Radford

Problems in Electrical Engineering for Technical Students

By B. Hamill. 197 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 25s.

The problems, together with answers, contained in this book are intended to cover the Ordinary National Certificate course in Electrical Engineering (A & B syllabuses) and the Electrical Science Course C (Electrical Installation Work).

Proceedings of the International Conference of Magnetism

876 pp. Med. 4to. Institute of Physics and The Physical Society. 1964. Price £8 8s.

This volume contains the 230 papers and discussions presented at the International Conference on Magnetism held at Nottingham University in September 1964.

This conference was arranged by the Institute of Physics and the Physical Society on behalf of the British National Committee for Physics of the Royal Society with the support of the International Union of Pure and Applied Physics.

Plasma Physics

By J. L. Delcroix. 266 pp. Demy 8vo. John Wiley. 1965. Price 50s.

Special emphasis on the relationship between microscopic and macroscopic descriptions of plasma is given in this book which is intended for the research student in plasma physics.

A quick introduction to this subject can be obtained by omitting the sections printed in smaller type which contain the more difficult aspects.

Soviet Research in New Semiconductor Materials

By D. N. Nasledov and N. A. Goryunova. 121 pp. Demy 4to. Consultants Bureau. 1965.

This book is a report based on seventeen separate papers on the various problems related to new semiconductor materials.

Statistical Dynamics of Linear Automatic Control Systems

By V. V. Solodovnikov. 576 pp. Crown 4to D. Van Nostrand Co. Ltd. 1965. Price £7 10s.

This book, the second edition, has been completely revised and now consists of 13 chapters dealing with the whole field of the statistical dynamics of linear automation control systems.

The author discusses the synthesis of servo systems, variable parameter, self-adaptive and sampled-data systems, the construction and use of computers for the correlation function, and methods of evaluating data.

First International Conference on Electron and Ion Beam Science and Technology

By R. Bakish. 945 pp. Med. 8vo. John Wiley and Sons. 1965. Price 185s.

This book contains the papers presented at the First International Conference on Electron and Ion Beam Science and Technology held at Toronto, Canada, in May 1964 sponsored by the Electrothermics Society together with the Metallurgical Engineers.

It consists of seven sections as follows: (1) Physics of Electron and Ion Beams, (2) Electron and Ion Beams in Microelectronics, (3) Electron Beams—Recording and Information Storage, (4) Electron Beams in Material Processing, (5) Electron Beams in Materials Joining, (6) Electron Beams in Microanalysis, and (7) Ion Propulsion.

Advances in Cryogenic Engineering

By K. D. Timmerhaus. 438 pp. Vol. 1. 524 pp. Vol. 2. Crown 4to. Plenum Press. 1965. Price $17.50

This publication is divided into two sections, Section A-L containing 62 and Section M-U 42 of the papers presented at the tenth National Cryogenic Engineering Conference held at the University of Pennsylvania in August 1964 and covering low temperature research in mechanical properties of materials, insulation, refrigeration methods, catalysis and recreations at low temperature and temperature measuring and control devices.

Transistor Electronic Organs for the Amateur

By A. Douglas and S. Astley. 76 pp. Demy 8vo. Isaac Pitman. 1965. Price 18s.

This small book contains a detailed design of a full scale organ for construction by the amateur.

Medical Electronics

By F. H. Bostem. 976 pp. Royal 8vo. Imprimerie S.A. 1965. Price 1.250FB

This book contains the proceedings of the Fifth International Conference on Medical Electronics held at the University of Liege in July 1963 at which 58 papers were presented under the following subject headings: (a) The Physical Transducer. (b) Analyzers and Computers for Physiological Signals. (c) Analyzers and Computers for E.E.G. (d) Simulators for Circulation Research. (e) Automation of Chemical Data Handling. (f) Foetal Heart Investigation Methods.

Guide to Radio Technique—Volume 1

By E. Julander. 238 pp. Med. 8vo. Philips Technical Library. 1965. Price 37s. 6d.

This is the first of three volumes and serves as an introduction to the fundamentals of electronics and radio techniques. Volume II "Passive Elements and Fundamental Circuits" and Volume III "Technique of AM Radio Reception, Transmitters, Measuring Instruments and Servicing" are to follow.

International Telemetering Conference 1965 Washington D.C.—Volume 1

Sponsored by International Foundation for Telemetering. 802 pp. Crown 4to. Aero-Space-Tech-Tronic Back Files Co. 1965. Price $20.00 USA $22.50 foreign

This book contains the proceedings of the conference held at Washington D.C. in May 1965 sponsored by the International Foundation for Telemetering at which some 55 papers were presented.

Structural Synthesis of High-Accuracy Automatic Control Systems

By M. V. Meerov. 341 pp. Demy 8vo. Pergamon Press. 1965. Price £5

The aim of this book is to establish a method of synthesis of control systems as distinct from the usual analysis of specific types. Stable systems with high gain are given prominence.

Matrix Algebra for Electronic Engineers

By P. Hlawiczka. 216 pp. Pott 4to. Iliffe Books Ltd. 1965. Price 45s.

The level of the subject matter in this book has been set to provide a basic course in matrix methods for students of electronic engineering. The book is divided into two parts, Part 1 serving as an elementary introduction requiring only a basic knowledge of algebra.

Part 2 deals with the eigenvalue problem and the differential equations of linear networks including the matrix form of the Laplace transformation.

Elements of Transistor Pulse Circuits

By T. D. Towers. 172 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 35s.

This book is mainly descriptive, detailed analysis and mathematics are kept to a minimum.

Among the subjects included are linear pulse amplifiers, astable and monostable multivibrators and their applications, generating time delays and gating control circuits. One chapter is given to the Eccles-Jordan bistable multivibrator and another to the general design features of diode, transistor and Miller integrator pumps and the Schmitt trigger circuit.

ELECTRONIC EQUIPMENT

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

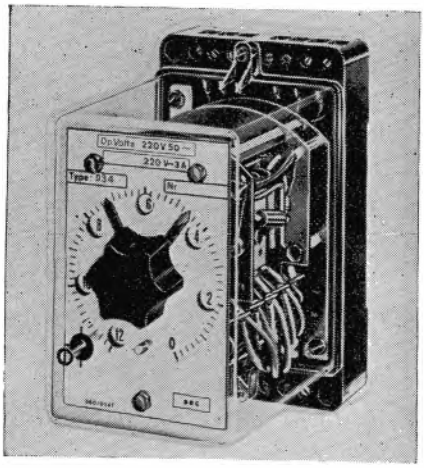
(Voir page 699 pour la traduction en français; Deutsche Übersetzung Seite 706)

PROCESS TIMERS

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustrated below)

A new range of three synchronous motor-driven process timers, specially designed for industrial applications, has been introduced by Magnetic Devices Ltd. All three incorporate a new feature, allowing the alternative of having timer return to the initial position on failure of supply, or to remain in the interrupted position till resumption of supply when it will complete the unelapsed time.



Series 934 can perform the majority of functions required in process control and automation, and has three ranges, any one of which can be chosen by positioning a selector. Within the scale of the time range selected the delay can be infinitely adjusted.

Control is by means of an external impulse switch, or on-off switch, and it is adjustable for three time delay ranges of 0 to 12, 0 to 60 and 0 to 120sec; 0 to 24, 0 to 120 and 0 to 240sec; 0 to 6, 0 to 30 and 0 to 60min; 0 to 3, 0 to 15 and 0 to 30h. Four microswitches are used, two operating instantaneously, and two after time-delay period.

Series 931 and 932 have one load and two load contacts respectively, with changeover micro switches used as load contacts. Within the scale of the time range selected the delay can be infinitely adjusted. Control is by means of an external on-off switch, and time delay ranges are 0 to 12, 0 to 24, 0 to 60, 0 to 120, 0 to 240sec; 0 to 6, 0 to 30, 0 to 60min; 0 to 3, 0 to 15, 0 to 30h.

Design styles include base and flush mounting with screw terminal connexions and plug in connexions.

Operation is by synchronous motor and electromagnetic clutch which are simultaneously energized by closing an external control switch. At any instant the unelapsed time is indicated by the position of a black pointer.

Setting accuracy is approximately 1·5 per cent of complete cycle, with repetitive accuracy better than 1 per cent. Operating voltage is 240V, 110V or 24V a.c., and operating power approximately 13VA. Flash test voltage is 2 500V.

All live and moving parts are protected by a transparent dustproof cover, with connexion terminals recessed. Ambient temperature range is −40°C to +60°C. Weight of the Series 931 and 932 is 1 lb 14 oz, of the Series 934 2 lb.

EE 86 751 *for further details*

SILICONE RUBBER ENCAPSULATED CAPACITORS

The Plessey Co. Ltd, Components Group, Kembrey Street, Swindon, Wiltshire

A new range of silicone rubber encapsulated capacitors, believed to be the first manufactured in Britain, is now available from the Components Group of The Plessey Co. Ltd.

Silicone rubber encapsulated units combine ease of manufacture with exceptional temperature resistance, while avoiding the packing and insulation difficulties associated with metal cans. They also offer special advantages in many applications.

In the case of large capacitors designed to carry heavy currents and operating at high ambient temperatures, for example, it is desirable to divide the current among several component units, often leading to a difficult shape for encapsulation. This presents no problem where silicone rubbers are used, since the units may be taped together into a block and suspended in a suitably shaped moulding box. The mould is then filled with a catalysed silastomer, which after curing provides a hard and durable encapsulation capable of operating without degradation at temperatures up to 250°C.

A typical 30μF capacitor designed to handle 115V a.c., with surge voltages up to 200V r.m.s. at 400c/s will operate at up to 100°C ambient temperature, and measures 4¾in by 2in by 3in.

Another important application of silicone rubber encapsulation is in mica capacitors designed for use at temperatures of 150°C and above. If encapsulation rather than open construction is required, silicone rubber is virtually the only satisfactory material. A 1μF mica unit measuring 2in by 1¾in by 1¼in and designed to operate at 350V d.c. is safe in service at temperatures up to 250°C.

Using silicone rubber encapsulation, conventional terminals are normally replaced by flying leads or tinned copper wires. For use at temperatures above 150°C pure silver wires are used. Although the process cannot provide the same degree of humidity protection as a metal can owing to the difficulty of sealing the leads, this is seldom important where rubber encapsulation is otherwise desirable.

EE 86 752 *for further details*

INTERKAMA

ELECTRONIC ENGINEERING will occupy Stand 4254 in Hall D2 at INTERKAMA (International Congress and Exhibition for Instrumentation and Automation) to be held in Düsseldorf, Western Germany, from 13 to 19 October 1965, and visitors will be welcome.

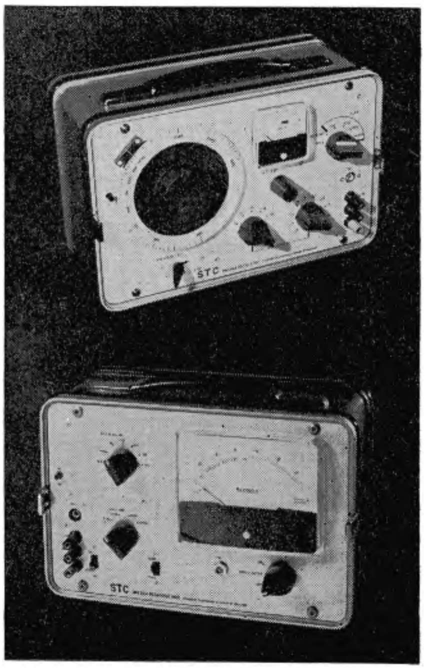
OSCILLATOR AND LEVEL MEASURING SET

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustrated below)

The Testing Apparatus Division of Standard Telephones and Cables Ltd has announced a complementary pair of portable battery-operated equipments for generating signals and measuring levels over the range 300c/s to 300kc/s. They are intended primarily for use on audio, broadcast and multi-circuit telephone systems.

The 74254-A oscillator has an output level variable in 1dB steps from 0 to −49dBm into 75, 125, 140 and 600Ω circuits. The attenuator has two controls,



0 to 40dB in 10dB steps and 0 to 9dB in 1dB steps. A set level control is incorporated in association with a meter calibrated ±0·25dBm. The output level accuracy is ±0·1dB at 0dB output (20°C) over the whole frequency range, although at other attenuator settings an extra 0·2dB error may arise. Between 10°C and 40°C, changes in ambient temperature will cause errors no greater than ±0·2dB. The harmonic content of the output will not exceed 1 per cent. The 5in diameter frequency dial has one engraved scale only reading between '300' and '3 000', and a range switch having factors of ×1, ×10 and ×100 produces frequency ranges of 300 to 3 000c/s, 3 to 30kc/s and 30 to 300kc/s. The frequency accuracy is ±1 per cent. The oscillator consumes about 50mA from a 12V supply made up of eight 1·5V dry cells. It weighs 12 lb (5·4kg) with batteries and measures 12¼ by 8 by 7⅛in (311 by 203 by 181mm).

The 74255-A level measuring set has been designed for use in conjunction with the 74254-A oscillator and has the same frequency range. The measuring range of the instrument is +21dBm to −51dBm and it can be used on 75, 125, 140 and 600Ω balanced and unbalanced circuits. Both through and terminated levels can be measured, and an accuracy of ±0·25dB can be expected at the temperature of calibration. The use of negative feedback not only gives a flat response/frequency characteristic, but also provides a high impedance feed to the meter so that an almost linear shape is obtained across the 4½in scale. The meter is calibrated from +1dB to −11dB in 0·2dB steps and works in conjunction with a switched control that enables an input level range of 72dB to be covered.

The level measuring set weighs 12½ lb and measures 12¼ by 8 by 7⅛in. Like the oscillator, it uses eight 1·5V dry cells from which it consumes about 40mA.

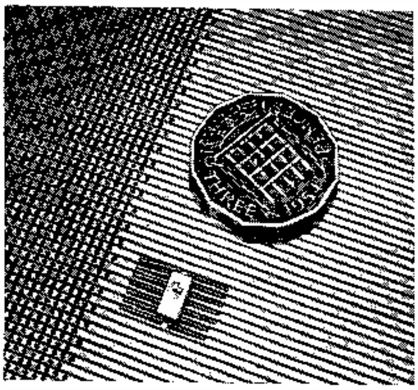
EE 86 753 *for further details*

PRINTED CIRCUIT BOARDS

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(Illustrated below)

'Veroboard' with a pitch of 0·05in is now being produced following demands by a number of electronic equipment manufacturers, anxious to obtain a 'mother' board for integrated circuits which does not involve the extremely high costs and delays that are arising in producing conventional printed circuit boards to the very close tolerances essential when using micro miniature components. This new range of Veroboard is manufactured from 0·031in thick s.r.b.p. with 0·0015in thick copper or cupro/nickel strips on one side 0·035in wide, which are pierced with holes 0·025in diameter on a matrix of 0·050in × 0·050in. It can be obtained as 6in square panels or as plug-in cards to suit 0·050in pitch edge connectors manufactured by Ultra Electronic (Components) Ltd.



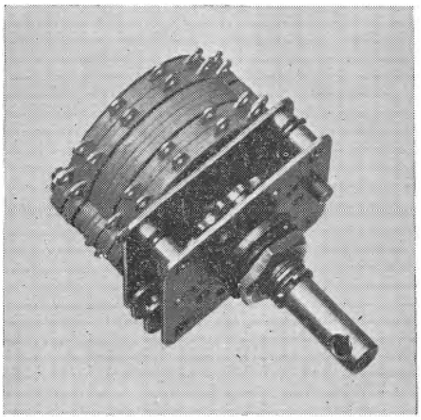
EE 86 754 *for further details*

ROTARY SELECTOR SWITCHES

I.D.M. Electronics Ltd, 1 Arkwright Road, Reading, Berkshire

(Illustrated below)

This range of manually operated rotary selector switches has been designed to meet the requirements of high quality instrument manufacturers who must have switches which are reliable and consistent in their performance over the life of the instrument.



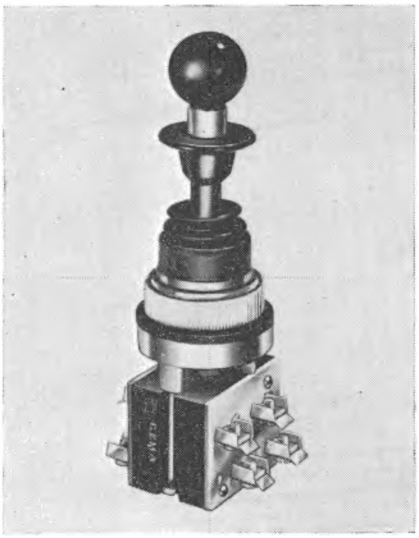
The switch wafers are constructed of glass-filled alkyd mouldings to provide good electrical and mechanical strength, positional control and anti-humidity qualities. A special design of wiper, coupled with hard silver contacts and wiper, provides low and stable contact resistance and corrosion protection. Shorting and non-shorting types are available and up to twelve positions per wafer can be provided. The contacts are enclosed, giving protection from dust and other contaminants.

EE 86 755 *for further details*

LEVER SWITCHES

Shawford Control Gear Co. Ltd, 715 Tudor Estate, Abbey Road, Park Royal, London, N.W.10

(Illustrated above right)



Available from Shawford Control Gear is a new version of two- and four-way lever ('joystick') selector switches. The new model incorporates a locking sleeve that must be lifted manually before the selector lever can be moved, thus preventing accidental operation. The sleeve is biased in the locked position, and unlocking is effected by lifting a disk concentric with the lever, after which any desired selection can be made. The closed contact, thus obtained, may be temporary or maintained as required.

On this model, as in all later models of existing lever selectors, cam operation of the contact block plungers has been replaced by operators moving vertically, so eliminating side load, and consequent risk of sticking.

The switch is fitted with standard two pole contact blocks and, if required, a further pair may be added, thus doubling the number of contacts. Rating is 4A at 220V a.c. with a maximum of 550V a.c.

EE 86 756 *for further details*

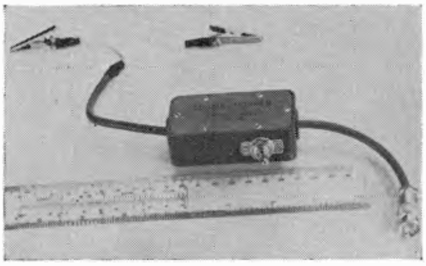
IMPEDANCE CHANGING PROBE

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

(Illustrated on page 694)

A self-contained impedance changer probe enabling low-impedance test gear such as voltmeters or oscilloscopes to be used without loading error is available from Kynmore Engineering Co Ltd. Manufactured by the Swindon firm of Adams-Norken Ltd, and designated the EPO-1, the device consists of a unity-gain amplifier contained, together with its on-off switch and two replaceable batteries, in a sealed diecast body completed with double-screened leads.

A feature of the unit is its high input impedance of 8MΩ in parallel with only 1pF. The output impedance is 200Ω in series with 50μF, giving a ratio of 40 000:1. The dynamic range is 4V peak-to-peak and bandwidth is 10c/s to 3Mc/s. The probe thus provides a convenient matching link for general laboratory use, suitable for any application where a high input impedance and low output impedance are required. Linearity is better than 1·0 per cent and noise is 20μV peak-to-peak when measured with an impedance of 100kΩ. Harmonic distortion measured at 1V, 10kc/s, is less than 0·25 per cent. The



unit has undergone climatic tests to DEF 133.

The probe body measures $3 \times 1{\cdot}75 \times 0{\cdot}875$in, the two replaceable batteries being Mallory type TM175 connected in series and giving a working life of 80 hours in continuous operation. The probe lead is normally fitted with a double-screened BNC plug, but modified units with different leads or specifications are available.

For wiring into equipment, or for plug-in applications, a standard encapsulated version (EPO-2) is available, contained in a ¾in diameter × 1½in long aluminium can without battery or coupling capacitors. It is available with solder pins, printed circuit pins or valve base pins.

EE 86 757 *for further details*

INTEGRATED CIRCUIT CONNECTOR

Ferranti Ltd, Kings Cross Road, Dundee, Scotland

(Illustrated below)

Ferranti Ltd has introduced a new S.8 connector which has been specially designed to accept integrated circuits enclosed in eight-lead TO-5 encapsulations.

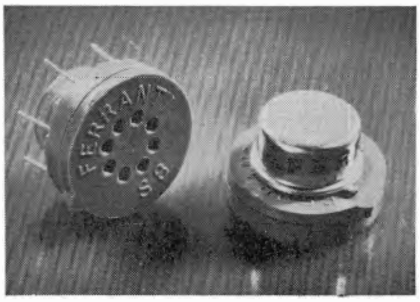
The S.8 connector is intended primarily for use in printed circuit boards but can prove of value when used in prototype equipments, development projects and test procedures.

Two of the features of the S.8 connector are the weight, which is only 1g, and the use of low-rate spring contacts which enable the contact force to be accurately controlled. These contacts are of beryllium copper, hard gold plated 0·0002in. (5 microns) thick.

To provide dimensional stability and excellent electrical properties the S.8 connector body is moulded in glass-filled nylon and to facilitate the correct insertion of the TO-5 a projection on the rim of the body is provided to which the tongue of the TO-5 is aligned.

Typical performance details of the S.8 connectors are:-

Insulation resistance: 1×10^7MΩ.



Contact resistance: 40mΩ (to which the integrated circuit leads contribute approximately 10mΩ).

Self inductance of each contact: 0·001μH.

Capacitance between adjacent contacts: 0·2pF.

Insertion force: 600g.

Temperature range: -55°C to +120°C.

EE 86 758 *for further details*

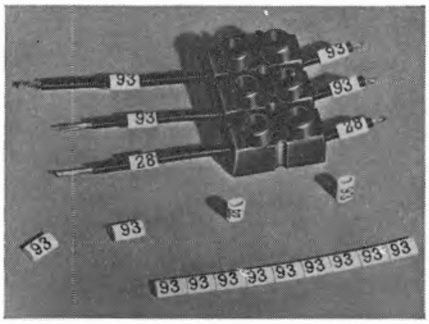
CABLE MARKING SYSTEM

Label Processes Ltd, 17-19 Crescent, Salford 5

(Illustrated below)

A rapid, easy, economic way to identify cables and wires has been introduced by Label Processes Ltd. The system consists of plastic tubing in roll form, which is printed and half cut through. Half cutting ensures that the printed sections stay in roll form until required yet enables each section of printed tubing to be easily detached.

The detached section is simply slipped around the cable or wire and shows



quickly and clearly what each cable or wire is. Numbers, symbols or any type of information can be applied to the tubing and by using different colours of plastic tubing a combined information/colour coding system is easily achieved. The tubing is printed and cut on the LP70 machine (one of a number of such machines made by Label Processes Ltd) which automatically prints and cuts sections at 3600 per hour. The hot stamping method of printing is used which gives a completely dry, smudge-free, scuff-resistant print which is usable immediately.

The LP70 can be operated by unskilled operatives and is capable of printing consecutive numbering or variable information, in areas up to 3in by 2in.

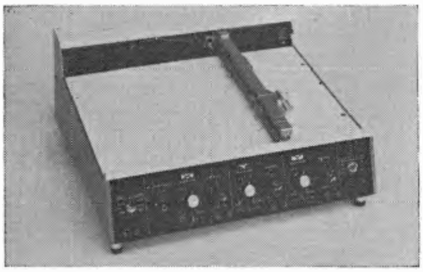
EE 86 759 *for further details*

TWO-PEN X-Y RECORDER

Hewlett Packard, Dallas Road, Bedford

(Illustrated above right)

A new bench type two-pen $X - Y_1 - Y_2$ recorder has been introduced by the Moseley Division of Hewlett-Packard. The new model 2FA is a companion to the previously offered 2FRA which is for rack mounting installations only. This is a precision 11in × 17in recorder which has two pens independently controlled in the vertical axis and integrally controlled in the horizontal axis. Thus three



channels are established, allowing simultaneous plotting of two dependent variables against a third independent variable. Horizontal separation of the pens is 0·1in. Available as an option is a third fixed event pen to produce identification marks at significant points in a recording.

Each axis of the 2FA has 11 calibrated d.c. input ranges with sensitivities from 0·5mV/in to 50V/in. Continuously variable scale factors in each range may be established by operating a transfer switch which permits arbitrary full scale range adjustments. In addition, the four most sensitive ranges may be converted to potentiometric input. Input resistance on all calibrated ranges is 1MΩ at null.

Included in the 2FA as a standard feature is a 5-range time-base which operates on the X-axis only. This provides sweeps from 0·5sec/div to 50sec/div corresponding to 7·5sec to 750sec for full scale pen traversal. The time-base supplies a built-in independent variable for plotting two curves simultaneously.

Other features are the 'Autogrip' electrostatic paper hold-down, continuous electronic reference, and 0·2 per cent accuracy on all ranges.

EE 86 760 *for further details*

WIDE RANGE OSCILLATOR

General Radio Co. (U.K.) Ltd, Bourne End, Buckinghamshire

(Illustrated below)

General Radio Company's new type 1310-A oscillator is a transistorized, general-purpose, constant-output signal source covering the frequency range from 2c/s to 8Mc/s in six decade bands. Very low distortion (under 0·25 per cent from 50c/s to 50kc/s) that is independent of load, high output (20V open-circuit), and provision for phase locking to a frequency standard are other features which make the new oscillator a useful signal source for the laboratory.

The instrument uses a capacitance-tuned, RC Wien-bridge oscillator driving

a low-distortion, shortable output amplifier. Output power is 160mW into 600Ω.

EE 86 761 *for further details*

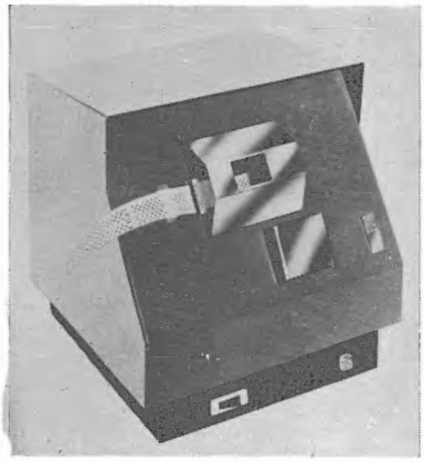
PAPER TAPE PUNCH

Automatic Punched Tape Ltd, Commerce House, Commercial Road, Paddock Wood, Kent

(Illustrated below)

The punch type P30 will operate at speeds of up to 30 characters/sec on 5, 6, 7, or 8-unit tape.

Facility is provided for automatically checking parity by contacts on the punch head. The contacts also give an indication of when the punch has actually passed through the paper. The P30 is fitted with a reverse feed facility and it is possible to advance the tape manually. The take-up reel, powered by an a.c. motor, and the tape supply reel each have a capacity of 1000ft of tape.



The punch can be supplied to operate at 24 or 48V d.c. A metal silencing cover is provided with the standard model: this is easily removed to allow a new supply of tape to be loaded.

The punch mechanism only for panel mounting can be supplied on request. Optional extras include an Ink Marker for verifier systems and a 6-unit teletypesetter punch-block.

EE 86 762 *for further details*

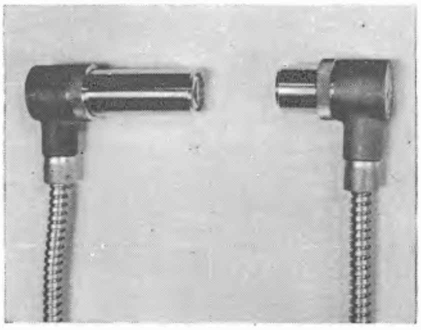
PHOTOCELL AND PROJECTOR HEADS

Panelectronics Ltd, Eastern Way, Bury St. Edmunds, Suffolk

(Illustrated above right)

This new additional series of 'Moniteye' photo-head, and companion light projectors is intended to provide for all general purpose transmitted light photoelectric requirements, and is suitable for operation over distances ranging from 3in to 12ft, higher powered projectors are available for longer distances if required. The units are of sturdy construction, completely sealed, and substantially oil and waterproof. Sealed connexions 3ft long are brought out at 90° to the axis of the light beam, and are protected by miniature flexible steel conduit.

A useful feature is the adjustable



mounting clamp to fix the tube rigidly in position. As the body detaches readily from the lens tube, correct sighting is not disturbed when the lamp is changed.

A choice of parallel or converging collimated light beams is available to provide for both long and close range optical requirements.

Four alternative types of cell can be accommodated by the receivers, these are transistor, cadmium selenide, and cadmium sulphide photo-conductive cells, and photo-emissive cells.

EE 86 763 *for further details*

LEVEL INDICATOR/CONTROLLER

Thomas Industrial Automation Ltd, Electronic Centre, Deansgate Lane, Altrincham, Cheshire

(Illustrated below)

This versatile instrument, the 'Levelrator', provides continuous indication of level/contents, together with high and/or low level control, on free-flowing solids and liquids. Visual and/or audible alarm can be provided at both the high and low level points, which can be independently set to operate over the full length of the electrode for most applications.

Only one electrode is required to provide continuous indication and high and/or low level control. The ingenious design of the electrode system can ensure good accuracies, and near-linear scales.

The Levelrator is suitable for applications on all sizes and shapes of container, and can be used in conjunction with both rigid and flexible electrodes.

Remote indicators can be supplied to customers' requirements, and these can be sited anywhere on the plant. The

indicating meter is completely independent of the high and/or low level control points.

Further economies can be achieved by the specially designed 'multi-point' system, which enables centralization of equipment where more than one container is involved.

EE 86 764 *for further details*

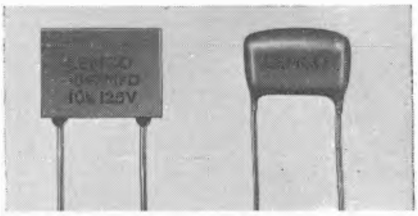
METALLIZED POLYESTER CAPACITORS

London Electrical Manufacturing Co. Ltd, Bridges Place, Parsons Green Lane, London, S.W.6

(Illustrated below)

LEMCO is now producing two new ranges of metallized polyester capacitors for use primarily with printed circuits, and a range of values based on the logarithmic scale is immediately available.

The two ranges differ in finish and size, offering a choice between fully moulded and resin insulated units, meeting the requirements of H5 DEF 5011 and H3 respectively. Solder coated wires are used for terminations to facilitate



easy soldering during production. The resin moulded form is designed to meet stringent climatic and vibration conditions, and for this purpose it incorporates an anti-moisture trap plinth which gives a very firm mounting.

A range of capacitances is available from 0·01μF to 0·22μF with standard tolerances of ±5 per cent ±10 per cent ±20 per cent at a working voltage of 160V d.c. At 20°C the insulation resistance is greater than 30kMΩ and power factor less than 0·0075 when measured at 1kc/s. The operating temperature range is −55°C to +85°C.

The material used in construction is pretested for faults and then wound on totally enclosed machines. Finally, before encapsulation, they are stabilized by a controlled heat treatment which results in a very reliable component. Special attention has been given to strengthening the end terminations with the added benefits of low inductance and power factor.

EE 86 765 *for further details*

VIBRATION EXCITERS

Derritron Electronic Vibrators Ltd, 24 Upper Brook Street, London, W.1

(Illustrated on page 696)

Derritron Electronic Vibrators Ltd is now producing two small electromagnetic vibration exciters. These vibration exciters utilize permanent magnets for the field and are designed for high frequency testing of sub-assemblies and components. As with all Derritron vibrators, except the 2½ lb. thrust model,

large diameter magnesium alloy table ; used and 5 specimen fixing points are rovided in the table. The vibrators have een designated types VP4 and VP4B. lowever, they are identical except that ne VP4B is supplied with a blower to rovide additional cooling of the drive oil and can therefore be driven by a nore powerful amplifier and will deliver nore thrust. The frequency over which ull thrust can be delivered is 5c/s to 900c/s and the vibrator is useful to 0kc/s and over, the first major table esonance is at 8900c/s and the thrust btainable from the VP4 and 100WLF mplifier is 24 lb. and from the VP4B nd 250WLF amplifier is 34 lb. The able weight is 0·99 lb. These vibration xciters may be supplied either with or vithout trunnion fixing. When supplied vith the trunnion fixing type T4 the ibrator may be operated in either the ertical of horizontal direction or any osition in between.

EE 86 766 *for further details*

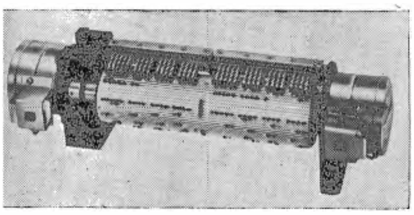
PROGRAMMING SWITCH

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustrated below)

A new Actan programming switch with i-directional, solenoid selected, timing notor drives is now available from ealectro Ltd. This new unit features lrives of 1rev/min and 1rev/h lthough other speeds can be supplied o drive the switch in the forward direction while the 1rev/min drive provides rapid 'return-to-zero' function without having to go through the full 360° lrum rotation. This feature is paricularly useful in testing applications vhen the test must be terminated as soon s a failure is indicated, requiring the est to be restarted.

Programme drums with up to 120 ositions and 51 contact stations are



offered with this driving arrangement and are completely adjustable without tools. Standard contacts are supplied as form C and are rated at 2A. Special U.L. approved contacts rated at 10 or 20A, 115V a.c. are also available.

EE 86 767 *for further details*

TACHO-RATIOMETERS

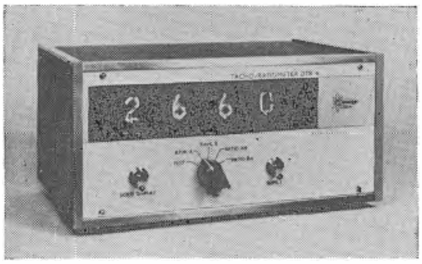
Advance Controls Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustrated below)

Advance Controls Ltd has released details of a new and extensive range of digital tachometers and tachometer/ratiometers introducing a large clear numerical display as an alternative to Dekatron display.

Sixteen standard models are available for early delivery, the variations in specification including either four or five decades and the use of either an internal crystal oscillator or the supply frequency as a time reference.

A wide range of applications include the accurate measurement of frequency,



shaft or machine speed or the ratio between two speeds to three places of decimals. This ratiometer technique enables percentage stretch or percentage draw measurements, to 0·1 per cent to be made in textile, plastic, paper metal or any other continuous process industries.

Input signals are usually obtained from magnetic transducers but other devices can be used, including tachogenerators, proximity detectors or photoelectric devices.

EE 86 768 *for further details*

LOW NOISE REED SWITCH

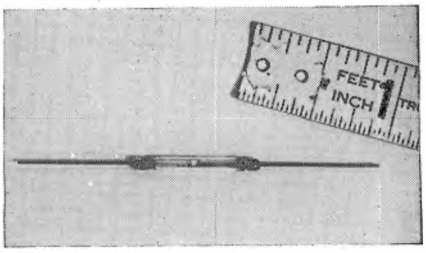
Flight Refuelling Ltd, Wimborne, Dorset

(Illustrated above right)

A miniature reed switch with thermal e.m.f. properties approximately 10 per cent of conventional reed switches, the MSMF-2 also incorporates a contact material specifically designed for low level circuits.

Exteremely low and stable contact resistances, combined with a minimum gap, considerably reduce mechanically generated noise. These characteristics, plus the low thermal e.m.f. property, make the MSMF-2 very suitable for low level or dry circuit applications.

Overall length: 2·250in
Glass length: 0·800in maximum
Diameter: 0·090in maximum
Thermal e.m.f.: Approximately 5μV/°C
Life expectancy: Over 10×10^6 operations at full rating. Almost infinite



at dry circuit loads.
Maximum contact rating: 28V d.c., 10mA.

EE 86 769 *for further details*

DUAL DIVERSITY RECEIVING TERMINAL

Racal Communications Ltd, Western Road, Bracknell, Berkshire

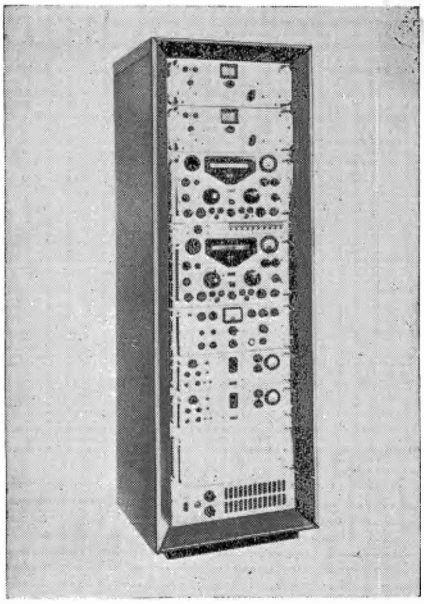
(Illustrated below)

The Racal RA.133 dual diversity receiving terminal is now in production. It is intended for use by broadcasting authorities throughout the world for reception of speech and music programmes in the international h.f. broadcast bands, which will then be rebroadcast by local transmitters or wire relay networks. A number of these terminals is being supplied to the BBC.

This receiving terminal covers a frequency range of 1 to 30Mc/s, or 2 to 24Mc/s with added r.f. pre-selection. The pre-selection circuits provide enhanced selectivity so that an interfering signal 5 per cent off-tune is attenuated by approximately 100dB.

Facilities are provided for dual-diversity reception of double or single sideband signals. Either the upper or lower sideband may be selected; in d.s.b. operation, this permits the use of the sideband which has least interference.

Two RA.117C receivers are incorporated in the RA.133 terminal, providing stability and setting accuracy. These RA.117 receivers may be separately-tuned or operated in a master-slave relationship, with the second v.f.o. signal



from one of them controlling the 'kilocycles' setting of both. This method of operation simplifies tuning and accurate frequency setting of both receivers is thereby assured.

An input attenuator with five switched settings, is fitted to each receiving path, and five switched i.f. passbands between 1·2kc/s and 13kc/s are incorporated. Diversity combining of the upper, lower, or double sideband signals from the two receiver paths is accomplished in passive networks, and comprehensive visual and aural monitoring facilities are provided.

EE 86 770 *for further details*

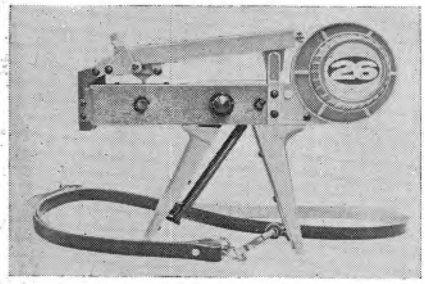
CABLE BINDING TOOL

Hellermann Electric Ltd, Gatwick Road, Crawley, Sussex

(Illustrated below)

Hellermann Electric Ltd has introduced a new cable binding tool—the Tyton—that is claimed to save almost 50 per cent in time and cost over similar methods of permanent binding.

The Tyton is cartridge loaded with 20 fasteners to enable 20 non-stop bindings to be made and because of the strength of each binding less cable bindings are required. The greater strength is obtained by the means of



fastening which is by deformation—the nylon strapping is not pierced and therefore it is not weakened.

The nylon strapping which is supplied on reels is automatically cut off when completing the binding. This prevents wastage of strapping and provides a tool that is ready for use at any moment. This also means that any size loom can be bound at any time without interruption or delay.

A halter strap, worn round the neck, is supplied with the tool to enable the operator to have both hands free when not binding cables.

EE 86 771 *for further details*

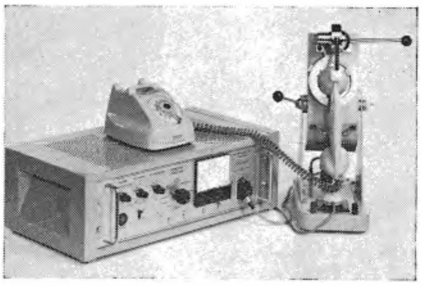
TELEPHONE SET TEST EQUIPMENT

Telefon Fabrik Automatic A/S, 7 Amaliecade, Copenhagen, Denmark

(Illustrated above right)

This telephone set test equipment is designed for electrical testing of complete telephone sets, handsets, telephone transmitters, receivers, and bell sets. The test comprises measurements of receiving, sending and side-tone efficiency, microphone noise, impedance, ringing, and direct voltage and current.

The test equipment is primarily intended for production and acceptance tests, but is of an accuracy sufficient to



justify its use in the laboratory, and operation of the equipment is sufficiently simple, to be carried out by personnel with no technical skill, after a minimum of instruction.

The test equipment comprises two sections. The electronic section, EJ5-01, consists of electronic circuits, controls, and a meter. The acoustic section, the artificial head, EJ5-02, comprises an artificial ear, an artificial mouth, and a universal test jig into which practically any handset can be quickly inserted and secured in the proper position. The two sections are interconnected by a cable.

During test the telephone set is connected to the equipment via a d.c. feeding circuit built into EJ5-01. The feeding circuit consists of a built-in stabilized power source of 24V or 48V, a feeding coil and a variable series resistor so that normal occurring line current can be obtained. If required, the telephone set can be fed via an artificial transmission line, for example, the TFA type EG 4-.

Transmission properties are measured using a weighted white noise instead of speech. It has been demonstrated that this method gives results equivalent to those obtained by means of voice to ear loudness tests. The weighting of the white noise consists of an attenuation increasing 6dB per octave with frequency. Thus, the frequency spectrum of the acoustic noise is roughly the same as the spectrum of an average human voice.

EE 86 772 *for further details*

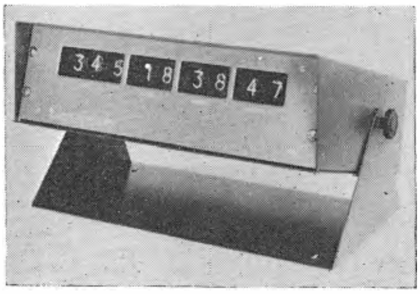
TIMING UNITS

System Engineering Services Ltd, Radley House, 35-39 South Ealing Road, London, W.5

(Illustrated below)

Timing units that display days, hours, minutes and seconds are now available from Systems Engineering Services Ltd.

The EECO 869 is available in two models; the 869-6 displays hours, minutes and seconds. The 869-9 displays days in addition to the information displayed by the 869-6 model. Up to three displays



can be driven from one EECO time code generator.

The display window features a polarized light screen that makes the lighted Nixie display clear and sharp. The 1·35in diameter Nixie tubes are rated at 200 000 hours operation.

The 869 may be rack mounted or a special fixture is available for wall, desk or ceiling mounting. Both models measure 3½in in height, 19in in width and 6¼in in depth, including mating connector.

Temperature range of the unit is −20°C to +55°C (−4°F to +131°F).

Other EECO time display units, the parallel-input model, EECO 865 and serial-input model, EECO 866 or 867, are used primarily to provide a remote display of time generated by a time code generator.

EE 86 773 *for further details*

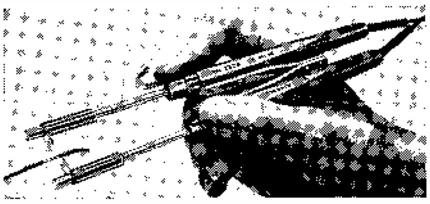
WIRE STRIPPERS

Light Soldering Developments Ltd, 28 Sydenham Road Croydon, Surrey

(Illustrated below)

The 'Adamin' model 2B6 is a new miniature thermal wire stripper for use with p.v.c. polythene, nylon and similar low melting-point insulating materials.

The Adamin stripper allows one-handed operation, using a simple tweezer action. It strips coverings up to about



5/32in diameter with no risk of damaging the conductors.

The model 2B6 is available for either 12 or 24V. Power consumption is approximately 14W, which provides just sufficient heat to melt the covering. Greater heat produces excessive melting and can cause liquid covering to be smeared along the stripped portion of the wire.

Any a.c. or d.c. source of 12 or 24V (depending upon the rating of the stripper) giving 1A or 0·5A respectively. Alternatively the 'Litesold' isolating transformers, type LT12 (12V 25W), LT24 (24V 25W) or MT24 (24V 55W) will enable the strippers to be used from any mains power point, with the additional advantage that the three-heat output tappings of these transformers provide some adjustment of temperature.

The wire to be stripped must be gripped firmly between the blades for a second or so to allow them to melt through the covering. A straight pull will then complete the stripping operation.

A similar tool, but operating at a higher temperature and known as model 2B24, is available for use with p.t.f.e. insulation. This does not melt the p.t.f.e. but will soften it sufficiently for the blades to shear it.

EE 86 774 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Electronic Division

Date: 18 October.
Lecture: Infra-red Devices—Propagation, Generation and Reception.
By: G. Phillips.
Date: 6 October.
Discussion: Filling Gaps in VHF/UHF Service Areas.
Date: 11 October.
Colloquim: **Telegraph and Data Transmission.**
(All wishing to attend must register. Forms available from the I.E.E. Secretary).
Date: 13 October.
Discussion: What is Group Delay?
Opened by: P. S. Brandon.
Date: 14 October.
Lecture: Long Wavelength Laser Generation.
By: L. E. S. Mathias.
Date: 18 October.
Lecture: The Industrial Training Act.
By: A. Lindley.
Date: 21 October.
Discussion: Instrument Scale Graduation (Based on the draft BSI specification).
Opened by: L. B. S. Golds.
Date: 25 October.
Lecture: Optoelectronics.
By: W. J. Poppelbaum.
(Jointly with the I.E.R.E. Computers Group).
Date: 28 October. Time: 6 p.m.
Lecture: The P.A.L. Colour Television System.
By: W. Bruch (Germany).
(Jointly with the I.E.R.E. and the Television Society).

Control and Automation Division

Date: 19 October.
Discussion: Does Industry Need Automatic Control

Power Division

Date: 26 October.
Colloquium: British and Continental Practice in the Design of Domestic Electric Appliances and Related Electricity Supply and Installation Factors.
All wishing to attend must register. Forms available from the I.E.E. Secretary).

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Radar Group

Held at: I.E.R.E. Lecture Room, 9 Bedford Square, W.C.1.
Date: 6 October. Time: 6 p.m.
Paper on: "SECAR"—A Modern SSR Ground Interrogator and Decoding Equipment.
By: H. W. Cole.
Held at: I.E.R.E. Lecture Room, 9 Bedford Square, W.C.1.
Date: 13 October. Time: 6 p.m.
Paper on: Signal Processing Using Optical Techniques.
By: D. C. Copper.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Date: 20 October. Time: 6 p.m.
Discussion: Computer/Instrument Interfaces.

Joint I.E.E./I.E.R.E Computer Groups

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 25 October. Time: 5.30 p.m.
Paper on: Opto-Electronics.
By: W. J. Poppelbaum.

Joint I.E.E./I.E.R.E. Television Society

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 28 October. Time: 6 p.m.
Lecture: P. A. L. Television System.
By: W. Bruch.

South-Western Section

Held at: University of Bristol Engineering Laboratories, University Walk, Clifton, Bristol 8.
Date: 20 October. Time: 7 p.m.
Paper on: Laser Technology and Applications.
By: J. MacDowal.
Held at: Bath Technical College, James Street West, Bath.
Date: 28 October. Time: 7 p.m.
Paper on: Infra-red Technology and Applications.
By: V. Roberts.

West Midland Section

Held at: North Staffs College of Technology, College Road, Stoke-on-Trent.
Date: 12 October. Time: 7 p.m.
Paper on: Computers—Present and Future.
By: R. Woolf.
Held at: Department of Electronic and Electrical Engineering, University of Birmingham.
Date: 28 October. Time: 6.30 p.m.
By: G. R. M. Garratt.

South Wales Section

Held at: Welsh College of Advanced Technology, Cardiff.
Date: 13 October. Time: 6.30 p.m.
Paper on: Phase Shift Networks Using Field Effect Transistors.
By: W. Gosling.

East Anglian Section

Held at: Hornchurch College of Further Education, 42 Ardleigh Green Road, Hornchurch.
Date: 12 October. Time: 7 p.m.
Paper on: Satellite Communications.
By: W. J. Bray.
Held at: North East Essex Technical College and School of Art, Sheepen Road, Colchester.
Date: 26 October. Time: 7.30 p.m.
Paper on: Some Examples of Simulation and Computer Techniques in Electronic Engineering.
By: N. A. Huttly.

Scottish Section

Held at: Department of Natural Philosophy, The University, Drummond Street, Edinburgh.
Date: 6 October. Time: 7 p.m.
Paper on: Input Spectrum and the Ability to Optimize.
By: I. Cochrane.
Held at: The Institute of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Date: 7 October. Time: 7 p.m.
Paper on: Input Spectrum and the Ability to Optimize.
By: I. Cochrane.
Held at: Carlton Hotel, North Bridge, Edinburgh 1.
Date: 28 October. Time: 6 p.m.
Paper on: Electronics in Medicine—Frustrations and Fulfilment.
(Joint I.E.R.E./I.E.E. Medical Group meeting).
By: J. M. A. Lenihan.

Southern Section

Held at: Farnborough Technical College.
Date: 14 October. Time: 7 p.m.
Paper on: Fuel Cells, a Customer's Viewpoint.
By: W. R. S. Davidson.
Held at: Lanchester Theatre, University of Southampton.
Date: 26 October. Time: 6.30 p.m.
Paper on: Loudness of Sonic Rooms and Similar Sounds.
(Joint meeting with the Royal Aeronautical Society).
By: E. E. Zepler.

Merseyside Section

Held at: Walker Art Gallery, Liverpool.
Date: 20 October. Time: 6.30 p.m.
Paper on: Colour Television.
By: G. N. Patchnett.

Thames Valley Section

Held at: J. J. Thomson Physical Laboratory, University of Reading.
Date: 19 October. Time: 7.15 p.m.
Paper on: Research in Electronic Instrumentation at the University of Reading.
By: E. A. Faulkner.

RADAR AND ELECTRONICS ASSOCIATION

Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Date: 12 October. Time: 7 p.m.
Lecture: Transmitting Aerial Systems for Television Broadcasting on U.H.F.
By: G. C. Platts

THE TELEVISION SOCIETY

Meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.
Date: 15 October.
Lecture: Problems Connected with the use of Colour Film for Colour Television.
(The Rank Organization, Film Processing Division).
By: F. P. Gloyns.
Held at: I.E.E. Savoy Place, London, W.C.2.
Date: 28 October. Time: 6 p.m.
Lecture: The P.A.L. Colour Television System.
By: W. Bruch.

PUBLICATIONS RECEIVED

VACTRIC CONTROL EQUIPMENT LTD is a new 24-page illustrated colour brochure which gives the latest information on the Company's activities in the servo component and other precision electromechanical fields, and sections also describe its associates Vactric Precision Tools Ltd, precision tool-makers, A. P. Besson & Partner Ltd, manufacturers of precision electronic components, and Rotron-Vactric Europa, manufacturers of air-moving devices.

Copies are available from Vactric Control Equipment Ltd, Garth Road, Morden, Surrey.

METAL CASED TUBULAR CAPACITORS is one of the latest technical bulletins from TCC. The wire-ended paper tubular capacitors described in this bulletin form a complete range of capacitors suitable for use in both Service and professional electronic equipment and offer a greater resistance to humidity than that provided by wax-dipped or plastic-moulded types. The range of capacitance extends from 200pF up to 1·0μF, and most capacitors are given a d.c. rating at 70°C and 100°C as well as an a.c. rating at 70°C. The d.c. voltage ratings at 70°C range from 150V to 1 000V, and there is a special range with a.c. voltage ratings up to 800V r.m.s.

Copies are available from The Telegraph Condenser Co. Ltd, North Acton, London, W.3.

TELESHIFT, the new G.E.C. Electronics frequency shift signalling system for transmitting control and supervisory signals over land lines, telephone lines, microwave and radio links etc., is described in a new publication IC/TT/2.

The four-page folder covers the Teleshift principle and the basic system as well as showing how, as a result of the modular design approach used, Teleshift can be readily combined with other compatible equipments in G.E.C. Electronics' range to build complex, sophisticated telemetry systems.

Copies of this leaflet are obtainable from G.E.C. Electronics Ltd, East Lane, Wembley, Middlesex.

HEAT SINKS FOR SEMICONDUCTOR POWER RECTIFIERS AND THYRISTORS booklet lists and illustrates ranges of cast aluminium heat sinks ready drilled to take standard semiconductor studs. Included in the publication are power dissipation curves for the various heat sinks showing conditions for convection cooling and forced air cooling.

Copies of this publication (MF/187X) are available from Standard Telephones & Cables Ltd, Semiconductor Division, Edinburgh Way, Harlow, Essex.

MULLARD SEMICONDUCTORS DESIGNERS GUIDE is now available in a new, revised edition.

The Guide contains Quick Find Charts for Transistors which list the devices under the main headings of collector voltage, total dissipation and cut-off frequency.

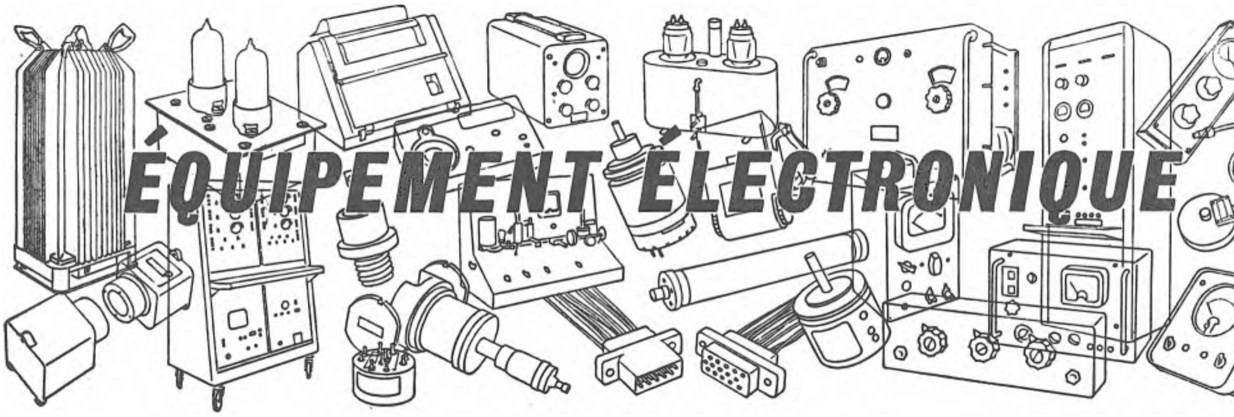
The basic information is augmented in subsequent pages under device headings which also include the full range of diodes, rectifier diodes and thyristors.

Full dimensional drawings and details of the international encapsulation outlines to which the devices comply are given at the back of the booklet.

Request for copies should be made on company headed notepaper to, Technical Office, Industrial Markets Division, Mullard Ltd, Mullard House, Torrington Place, London, W.C.1.

A SELECT BIBLIOGRAPHY ON SEMICONDUCTOR RELIABILITY was compiled by B. J. S. Williams as a contribution to the Symposium on Engineering for Reliability in the Design of Semiconductor Components held at Hatfield College of Technology in May 1965. The arrangement is by broad subject heads to provide a general scanning facility; the list of contents is supplemented by subject and author indexes. Copies are available, price 4s., from the Hertfordshire County Council Technical Library and Information Service, Hatfield College of Technology, Hatfield, Hertfordshire.

ÉQUIPEMENT ÉLECTRONIQUE



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 692 à 697

MINUTERIES DE PROCESSUS

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustration à la page 692)

Une nouvelle gamme de trois minuteries de processus actionnées par moteur synchrone et spécialement conçues pour les applications industrielles a été réalisée par la société Magnetic Devices Ltd. Toutes trois comportent une nouvelle caractéristique car elles offrent la possibilité d'un retour à la position initiale en cas de panne de courant ou du maintien à la position interrompue jusqu'au retour du courant qui complètera le temps non-parcouru.

Les appareils de la série 934 peuvent effecteur la plupart des fonctions exigées par le contrôle des processus et l'automatisation. Il comporte trois gammes, dont l'une peut être choisie en plaçant un sélecteur à la position voulue. Le retard peut être réglé indéfiniment à l'intérieur de la gamme de temps choisie.

Le contrôle s'effectue à l'aide d'un interrupteur d'impulsions extérieur ou d'un interrupteur arrêt-marche. Il est réglable pour trois gammes de retard de temps de 12, 0 à 60 et 0 à 120 sec; 0 à 24, 0 à 120 et 0 à 240 sec; 0 à 6, 0 à 30 et 0 à 60 min; 0 à 3, 0 à 15 et 0 à 30 h. Quatre microrupteurs sont utilisés, deux fonctionnant instantanément et deux après la période de retard de temps.

Les appareils de la série 931 et 932 sont à contacts de charge unique ou de double charge respectivement, avec microrupteurs de permutation utilisés comme contacts de charge. Le retard est à réglage infini sur l'échelle de la gamme de temps choisie. Le contrôle s'effectue au moyen d'un interrupteur arrêt-marche extérieur. Les gammes de retard de temps sont: 0 à 12, 0 à 24, 0 à 60, 0 à 120 et 0 à 240 sec; 0 à 6, 0 à 30, 0 à 60 min; 0 à 3, 0 à 15 et 0 à 30 h.

Les divers modèles comprennent ceux à montage encastré et à la base avec connexions de bornes à vis et connexions à fiches.

Le fonctionnement s'effectue par moteur synchrone et embrayage électromagnétique qu'on amorce simultanément en fermant un interrupteur de commande externe. Le temps restant à parcourir est indiqué à n'importe quel moment par la position d'une aiguille noire.

La precision de calage est d'environ 1,5 % du cycle complet, avec précision de répétition supérieure à 1 %. La tension de fonctionnement est de 240 V, 110 V ou 24 V c.a. et la puissance de fonctionnement est d'environ 13 VA. La tension de l'essai-éclair est de 2500 V.

Toutes les pièces mobiles ou sous tension sont protégées par une housse transparente à l'épreuve de la poussière avec bornes de connexion en retrait. La gamme de températures ambiantes va de −40° C à +60° C. Les composants de la série 931 et 932 pèsent 1,65 kg et ceux de la série 934 1,85 kg.

EE 86 751 *pour plus amples renseignements*

CONDENSATEUR ENCAPSULÉ EN CAOUTCHOUC AU SILICIUM

The Plessey Co. Ltd. Components Group, Kembrey Street, Swindon, Wiltshire

Une nouvelle gamme de condensateurs encapsulés en caoutchouc au silicium qui seraient les premiers à être réalisés en Grande-Bretagne peut être obtenue maintenant du groupe des composants de The Plessey Co. Ltd.

Les condensateurs encapsulés en caoutchouc au silicium allient la facilité de la fabrication à une résistance exceptionnelle à la température tout en évitant les difficultés d'isolement et de conditionnement qui se produisent dans le cas des enveloppes métaliques. Ils présentent également des avantages particuliers dans de nombreuse applications.

Dans le cas des grands condensateurs conçus pour porter de forts courants et fonctionner à des températures ambiantes élevées, par exemple, il est conseillable de diviser le courant entre plusieurs composants, ce qui donne souvent lieu à des formes difficiles à encapsuler. Il n'y a pas de problème, cependant, dans le cas du caoutchouc au silicium car les éléments peuvent être groupés sur un même bloc et suspendus dans un coffret moulé de forme appropriée. Le moulage est ensuite rempli d'un silastomère catalysé qui devient après traitement un encapsulage durable et solide pouvant fonctionner sans risque de détérioration aucun à des températures allant jusqu'à 250° C.

Un condensateur typique de 30 μF, conçu pour recevoir 115 V c.a. avec des surtensions allant jusqu'à 2000 V efficaces à 400 Hz fonctionnera à des températures ambiantes allant jusqu'à 100° C. Ses dimensions sont de 12 cm × 5 cm × 7,6 cm.

L'encapsulage au caoutchouc de silicium trouve également une application importante dans les condensateurs au mica étudiés pour l'utilisation à des températures maxima de 150° C. Si l'on préfère l'encapsulage à la construction ouverte, le caoutchouc au silicium est pratiquement le seul matériau donnant satisfaction. Un condensateur au mica mesurant 5 cm × 4,4 cm × 3,8 cm et conçu pour l'utilisation à une tension continue de 350 V est parfaitement sûr en service à des températures pouvant atteindre jusqu'à 250° C.

Lorsqu'on emploie l'encapsulage au caoutchouc au silicium, les bornes classiques sont normalement remplacées par des câbles volants ou des fils en cuivre étamé. Des fils en argent pur sont employés pour des températures dépassant 150° C. Bien que le processus ne puisse assurer le même degré de protection contre l'humidité qu'un métal,

INTERKAMA

ELECTRONIC ENGINEERING

occupera le Stand 4254 dans le Hall D2 à INTERKAMA (Congrès et Exposition Internationaux d'Instrumentation et d'Automatisation) qui se tiendra à Düsseldorf, Allemagne Occidentale, du 13 au 19 octobre 1965. Nos visiteurs y seront les bienvenus.

en raison de la difficulté de pouvoir sceller les câbles, ce détail est rarement d'importance lorsque l'encapsulage au caoutchouc est souhaitable.

EE 86 752 *pour plus amples renseignements*

OSCILLATEUR ET INSTRUMENT DE MESURE DU NIVEAU

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(*Illustration à la page 692*)

La division d'appareils de contrôle de la société Standard Telephones and Cables Ltd vient de mettre au point une paire supplémentaire de matériels portatifs fonctionnant sur batterie pour produire des signaux et des niveaux de mesure dans la gamme de 300 Hz à 300 kHz. Ces appareils ont été conçus pour les systèmes acoustiques de diffusion et de téléphone à plusieurs circuits.

L'oscillateur 74254-A a un niveau de sortie variable en plots de 1 dB de 0 à —49 dBm dans des circuits de 75, 125, 140 et 600 Ω. L'atténuateur comprend deux commandes de 0 à 40 dB en plots de 10 dB et de 0 à 9 dB en plots de 1 dB. Une commande de niveau réglé est incorporée à l'appareil de même qu'un instrument de mesure étalonné à ±0,25 dBm. La precision du niveau de sortie est de ±0,1 dB à une sortie de 0 dB (20°C) sur toute la gamme de fréquence, bien qu'à d'autres calages d'atténuation une erreur supplémentaire de 0,2 dB puisse se produire. Entre 10°C et 40°C, les changements de température ambiante peuvent causer des erreurs ne dépassant pas ±0,2 dB. La teneur en harmoniques de la sortie ne dépasse pas 1 %. Le cadran de fréquence à diamètre de 12,5 cm comporte une échelle gravée ne lisant que de "300" à "3 000", et un interrupteur de gamme à facteurs de ×1, ×10, et ×100 produisant des gammes de fréquence de 300 à 3 000 Hz, de 3 à 30 kHz et de 30 à 300 kHz. La précision de fréquence est de ±1 %. La consommation électrique de l'oscillateur est d'environ 50 mA à partir d'une alimentation de 12 V composée de huit piles seches de 1,5 V. Il pèse 5,4 kg avec ses batteries et mesure 311 × 203 × 181 mm.

L'appareil de mesure de niveau 74255-A a été conçu pour être utilisé en liaison avec l'oscillateur 74254-A et il a la même gamme de fréquence. La gamme de mesure de l'instrument est de +21 dBm à —51 dBm et il peut être utilisé sur circuit équilibré ou déséquilibré de 75, 125, 140 et 600 Ω. Les niveau continus et à terminaison peuvent être mesurés et une précision de ±0,25 dB peut être obtenue à la température d'étalonnage. L'emploi de la contre-réaction assure non seulement une caractéristique de fréquence à réponse linéaire mais fournit également une alimentation à haute impédance à l'instrument de mesure de sorte qu'une forme presque linéaire peut être obtenue sur l'échelle de 11,25 cm. L'instrument de mesure est étalonné de +1 dB à —11 dB par plots de 0,2 dB et fonctionne en liaison avec une commande à commutation qui permet de couvrir une gamme de niveau d'entrée de 72 dB.

L'instrument de mesure du niveau pèse 5,7 kg et mesure 311 × 203 × 281 mm comme l'oscillateur; il emploie huit piles sèches de 1,5 V dont il consomme environ 40 mA.

EE 86 753 *pour plus amples renseignements*

PLAQUETTES DE CIRCUIT CIMPRIMÉ

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(*Illustration à la page 693*)

Les plaquettes Veroboard à pas de 0,05 pouce répondent à la demande d'un certain nombre de fabricants de composants électroniques, désirant obtenir une plaquette "pilote" pour circuits intégrés qui n'entraîne pas le frais extrêmement élevés et les retards qui se produisent dans la fabrication des plaquettes de circuit imprimé classiques selon les tolérances très rigoureuses exigées par l'emploi de composants microminiature. Cette nouvelle gamme de plaquettes Veroboard est fabriquée avec des bandes de papier imprégné de résine synthétique de 0,031 pouce d'epaisseur avec bandes de cuivre ou de cuivre-nickel de 0,0015 pouce d'épaisseur et d'une largeur de 0,035 pouces sur une des côtés. Ces bandes sont percées de trous de 0,025 pouce de diamètre sur une matrice de 0,050 pouce × 0,050 pouce. Elles peuvent être obtenues sous forme de panneaux de 6 pouces carrés ou sous formes de cartes à fiches s'adaptant à des connecteurs à pas de 0,050 pouce fabriqués par Ultra Electronics (Components) Ltd.

EE 86 754 *pour plus amples renseignements*

INTERRUPTEURS SÉLECTEURS ROTATIFS

I.D.M. Electronics Ltd, 1 Arkwright Road, Reading, Berkshire

(*Illustration à la page 693*)

Cette gamme d'interrupteurs sélecteurs rotatifs à fonctionnement manuel a été réalisée pour répondre à la demande d'instruments de haute qualité qui sont nécessaires aux fabriquants ayant besoin d'interrupteurs sûrs et d'un fonctionnement uniforme pendant toute la durée de vie de l'instrument.

Les galettes de l'interrupteur sont en moulages Alkyd remplis de verre pour assurer une bonne résistance électrique et mécanique et des qualités de commande de position et de résistance à l'humidité. Un frotteur de conception inédite, accouplé à des contacts en argent solide, assure une résistance de contact stable et basse et la protection contre la corrosion. Des types à court-circuit ou non court-circuitables sont livrables et jusqu'à 12 positions par galette peuvent être prévues. Les contacts sont enfermés, ce qui les protège contre la poussière ou toute autre forme de contamination.

EE 86 755 *pour plus amples renseignements*

COMMUTATEUR À LEVIER

Shawford Control Gear Co. Ltd, 715 Tudor Estate, Abbey Road, Park Royal, London, N.W.10

(*Illustration à la page 693*)

La société Shawford Control Gear fournit une nouvelle version de selecteur à levier à deux et quatre directions. Le nouveau modèle comprend un manchon de verrouillage, qui doit être soulevé à la main avant de pouvoir déplacer le levier sélecteur, afin d'empêcher le fonctionnement accidentel.

Le manchon est polarisé dans la position verrouillée, et le déverrouillage s'effectue en sous levant un disque concentrique par rapport au levier, ce qui permet ensuite d'effecteur n'importe quelle selection voulue. Le contact fermé ainsi obtenu peut être temporaire ou maintenu suivant les besoins.

Sur ce modèle, ainsi que sur tous les autres modèles ultérieurs de sélecteur à levier, le fonctionnement par came des plongeurs de bloc de contact a été remplacé par des pièces se déplaçant verticalement, et éliminant ainsi les charges latérales et par conséquent le risque de collage.

Le commutateur est muni de blocs de contact normaux bi-polaires et, si nécessaire, on peut ajouter une paire supplémentaire de manière à doubler le nombre des contacts. La puissance nominale est de 4 A à 220 V c.a. avec une tension maxima de 550 V c.a.

EE 86 756 *pour plus amples renseignements*

SONDE DE CHANGEMENT D'IMPÉDANCE

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

(*Illustration à la page 694*)

Une sonde de changement d'impédance indépendante, permettant d'utiliser sans erreur de charge des instruments de contrôle à faible impédance, tels que les voltmètres ou les oscilloscopes, est fournie par la société Kynmore Engineering Co. Ltd. Construite par la filiale de Swindon de la société Adams-Norken Ltd, et portant la référence EPO-1, cette sonde se compose d'un amplificateur à gain unitaire logé, avec son interrupteur arrêt-marche et deux batteries remplaçables, dans un coffret scellé comportant deux câbles à double blindage.

Cette sonde se caractérise par son impédance d'entrée élevée qui est de 8 mΩ en parallèle avec 1 pF seulement. L'impédance de sortie est de 200 Ω en série avec 50 μF, donnant un rapport de 40 000 : 1. La gamme dynamique est de 4 V de crête à crête et la largeur de bande est de 10 Hz à 3 MHz. La sonde constitue donc un accessoire pratique à l'usage général des laboratoires et convenant à n'importe quelle application exigeant une impédance d'entrée élevée et une impédance de sortie réduite. La linéarité est supérieure à 1,0 pour cent et le bruit est de 20 V de crête à crête lorsqu'il est mesuré avec une impédance de 100 kΩ. La distorsion harmonique mesurée à 1 V, 10 kHz, est inférieure à 0,25%. La sonde a été soumise a des essais climatériques suivant la spécification DEF 113.

Le corps de la sonde mesure 7,62 cm × 4,44 cm × 2,22 cm, les deux batteries remplaçables du type Mallory TM175 étant reliées en série et donnant une durée de vie de 80 heures de fonctionnement continu. La câble de la sonde est normalement muni d'une fiche BNC à double blindage mais il existe des appareils modifiés avec des câbles différents ou des spécifications différentes.

Pour le bobinage sur un équipement ou pour les applications à fiches, il existe une version normale encapsulée (EPO-2), logée dans un boîtier en aluminium mesurant 1,9 cm de diamètre × 3,81 cm de long, sans batterie ou condensateurs de couplage. Elle est fournie avec broche de soudage, broche de circuit imprimé ou broche de base de lampe.

EE 86 757 *pour plus amples renseignements*

CONNECTEUR DE CIRCUIT INTÉGRÉ

Ferranti Ltd, Kings Cross Road, Dundee, Ecosse

(Illustration à la page 694)

La société Ferranti Ltd a réalisé un nouveau connecteur S.8 qui a été spécialement mis au point pour recevoir des circuits intégrés enfermés dans des encapsulages à 8 conducteurs T0-5.

Le connecteur S.8 a été principalement prévu pour être utilisé dans des plaquettes de circuit imprimé mais il peut être également de grande valeur dans des matériels de prototype, des projets de développement et des processus de contrôle.

Deux des caractéristiques du connecteur S.8 sont, en premier lieu, son poids réduit qui n'est que de 1 g, et l'emploi de contacts à ressort à faible puissance qui permettent de contrôler avec précision la force de contact. Ces contacts sont en cuivre de béryllium doré et leur épaisseur est de 5 microns.

Pour lui assurer la stabilité dimensionnelle et d'excellentes propriétés électriques, le corps du connecteur S.8 est moulé en nylon de verre et pour faciliter l'introduction correcte du T0-5 la couronne du corps est en saillie ce qui permet de fixer la langue du T0-5.

Les détails de performance caractéristiques des connecteurs S.8 sont comme suit:

Résistance d'isolement: 1×10^7 MΩ

Résistance de contact: 40 MΩ (auquels les conducteurs du circuit intégrés contribuent environ 10 MΩ)

Auto-inductance de chaque contact: 0,001 μH

Capacitance entre les contacts adjacents: 0,2 pF

Force d'introduction: 600 g

Gamme de température: -55°C à $+120$°C.

EE 86 758 *pour plus amples renseignements*

SYSTÈME DE MARQUAGE DE CÂBLES

Label Processes Ltd, 17-19 Crescent, Salford 5

(Illustration à la page 694)

Un nouveau système rapide économique et d'un emploi facile a été réalisé par la société Label Processes Ltd pour l'identification des câbles et fils métalliques. Ce système se compose d'un tube en matière plastique sous forme de rouleau imprimé sur la moitié de sa longueur et coupé sur l'autre moitié. Cette méthode de construction permet de garder les sections imprimées sous forme de rouleau jusqu'à ce qu'on en fasse usage tout en permettant de detacher sans difficulté chacune des sections du tube imprimé.

On fait passer chacune des sections détachées autour du câble ou du fil pour pouvoir indiquer clairement et rapidement le type de câble ou de fil dont il s'agit. Des chiffres, des symboles ou n'importe quel genre de renseignement peuvent être appliqués sur le tube et, en utilisant différentes couleurs de tube plastique, on obtient facilement un système de codage combiné d'informations/couleur. Le tube est imprimé et taillé sur la machine LP70 (fabriquée comme d'autres machines de ce genre par Label Processes Ltd.) qui coupe et imprime automatiquement 3 600 sections par heure. La méthode d'estampage à chaud est utilisée car elle assure une impression entièrement sèche, sans bavures, et résistante au frottement. Cet impression est donc utilisable immédiatement.

Le système LP70 peut être employé par des ouvriers non qualifiés et imprimer des chiffres consécutifs ou des renseignements variables sur des surfaces de 7,62 cm × 5,08 cm.

EE 86 759 *pour plus amples renseignements*

ENREGISTREUR X-Y À DEUX PLUMES

Hewlett Packard, Dallas Road, Bedford

(Illustration à la page 694)

Un nouvel enregistreur $X-Y_1-Y_2$ à deux plumes du type à banc d'essai a été crée par la division Moseley de la société Hewlett-Packard. Le nouveau modèle 2FA constitue le pendant du modèle 2FRA qui n'est prévu que pour montage sur bâti. Il s'agit en effet d'un enregistreur de précision mesurant 27,94 cm × 43,18 cm qui comporte deux plumes à contrôle indépendant dans l'axe vertical et à contrôle intégral dans l'axe horizontal. Trois canaux sont ainsi établis, permettant le tracé simultané de deux variables dépendants par rapport à un troisième variable indépendant. La séparation horizontale des plumes est de 0,25 cm. On peut obtenir sur demande une troisième plume fixe qui produit des marques d'identification à certains points importants de l'enregistrement.

Chaque axe de l'enregistreur 2FA comprend 11 gammes d'entrée de courant continu étalonnées dont les sensibilités vont de 0,5 mV/pouce à 50 V/pouce. Des facteurs d'échelle à variation continue dans chaque gamme peuvent être établis en actionnant un commutateur de transfert qui permet les réglages arbitraires de gamme sur la totalité de l'échelle. En outre, les quatre gammes les plus sensibles peuvent être converties en entrées potentiomètriques. La résistance d'entrée sur toutes les gammes étalonnées est de 1 MΩ à zéro.

L'enregistreur 2FA est muni de manière courante d'une base de temps à 5 gammes qui ne fonctionne que sur l'axe X. Cette base assure des balayages de 0,5 sec/div à 50 sec/div correspondant à 7,5 sec à 750 sec pour une traversée de la plume sur la totalité de l'échelle. La base de temps fournit un variable indépendant incorporé pour tracer simultanément deux courbes.

L'appareil comprend, enfin, une griffe de papier électrostatique "Autogrip", une référence électronique continue et une précision de 0,2% sur toutes les gammes.

EE 86 760 *pour plus amples renseignements*

OSCILLATEUR À GAMME ÉTENDUE

General Radio Co. (U.K.) Ltd, Bourne End, Buckinghamshire

(Illustration à la page 694)

Le nouvel oscillateur type 1310-A de la general Radio Company est une source de signaux transistorisée universelle et à sortie constante couvrant la gamme de fréquence de 2 Hz à 2 MHz en six bandes de décades. Il se caractérise par une très faible distorsion (moins de 0,25% de 50 Hz à 50 kHz) indépendante de la charge, une sortie élevée (20 V en circuit ouvert) et un verrouillage de phase à un étalon de fréquence. Ces particularités font du nouvel oscillateur une source de signaux utile à l'usage des laboratoires.

L'instrument utilise un oscillateur RC à pont de Wien accordé en capacité qui actionne un amplificateur à faible distorsion et à sortie pouvant être court-circuitée. La puissance de sortie est de 160 mW dans 600 Ω.

EE 86 761 *pour plus amples renseignements*

PERFORATRICE DE BANDE PERFORÉE

Automatic Punched Tape Ltd, Commerce House, Commercial Road, Paddock Wood, Kent

(Illustration à la page 695)

La perforatrice type P30 fonctionne à des vitesses allant jusqu'à 30 caractères/sec sur bande à 5, 6, 7, ou 8 unités.

La parité peut être contrôlée automatiquement au moyen des contacts sur la tête de la perforatrice. Les contacts indiquent également le moment où la perforatrice a effectivement traversé le papier. Elle est munie d'un dispositif d'avancement inversé qui permet de faire avancer la bande à la main. La bobine enrouleuse, entraînée par un moteur à courant alternatif, et le rouleau d'avancement de la bande ont tous deux une capacité de 305 m de bande.

La perforatrice peut être prévue pour l'utilisation à 24 ou 48 V c.c. Un couvercle métallique d'assourdissement est founi avec le modèle standard; ce couvercle peut être enlevé facilement pour permettre de charger un nouveau rouleau de papier.

Le mécanisme de perforation pour montage sur panneau seulement peut être fourni sur demande. Les accessoires facultatifs comprennent un marqueur à l'encre pour les systèmes de vérification et un bloc perforateur à télétype à six éléments.

EE 86 762 *pour plus amples renseignements*

CELLULES PHOTOÉLECTRIQUES ET TÊTES DE PROJECTION

Panelectronics Ltd, Eastern Way, Bury St. Edmunds, Suffolk

(Illustration à la page 695)

Cette nouvelle série de têtes photoélectriques "Moniteye" et de projecteurs de lumière a été prévue pour tous les usages photoélectriques de transmission de lumière et elle peut être utilisée sur des distances allant de 7,62 cm à 3,65 m. Des projecteurs de grande puissance peuvent être fournis sur demande pour de plus longues distances. Les éléments sont de construction robuste, entièrement scellés et à l'épreuve de l'eau et de l'huile. Des connexions scellées de 91 cm de long sont placées à 90° de l'axe du faisceau luminex et protégées par une conduite miniature en acier souple.

Il existe en outre une pince de montage réglable qui permet de fixer le tube de manière parfaitement rigide. Etant donné que le corps se détache facilement du tube de la lentille, l'exactitude de la mire n'est pas dérangée lorsqu'on change la lampe.

Un choix de faisceaux lumineux parallèles ou collimatés convergents est prévu pour les applications optiques à longue et à courte distance.

Les récepteurs peuvent recevoir quatre types différents de cellules, à savoir des cellules à transistors, au sélénium de cadmium, des cellules photoconductrices au sulfure de cadium et des cellules photo-émissives.

EE 86 763 *pour plus amples renseignements*

CONTRÔLEUR-INDICATEUR DE NIVEAU

Thomas Industrial Automation Ltd, Electronic Centre, Deansgate Lane, Altrincham, Cheshire

(Illustration à la page 695)

Le "Levelrator" est un instrument d'une grande souplesse d'emploi qui fournit une indication continue de niveau et de contenu, et assure le contrôle à niveau élevé ou bas de solides et liquides circulant librement. Un dispositif d'alarme visuel ou audible peut être fixé tant sur les points de niveau élevé que sur ceux de niveau réduit. Ce dispositif peut être réglé indépendamment pour fonctionner sur la longueur totale de l'électrode pour la plupart des applications.

Une seule électrode est nécessaire pour fournir une indication continue et assurer le contrôle de niveau élevé ou de niveau réduit. La conception ingénieuse du système à électrode garantit une grande précision et des échelles quasi-linéaires.

Le Levelrator peut être utilisé sur toutes les formes et toutes les dimensions de conteneurs. Il peut en outre être utilisé en liaison avec des électrodes souples ou rigides.

Des indicateurs à distance peuvent être fournis selon les spécifications du client et ces indicateurs peuvent être fixés à n'importe quel point de l'installation. L'indicateur est entièrement indépendant des points de contrôle de niveau élevé ou de niveau réduit.

On peut réaliser une économie en utilisant le système "à plusieurs points" de conception spéciale qui permet de centraliser le matériel lorsqu'on emploie plusieurs conteneurs.

EE 86 764 *pour plus amples renseignements*

CONDENSATEURS AU POLYESTER MÉTALLISÉ

London Electrical Manufacturing Co. Ltd, Bridges Place, Parsons Green Lane, London, S.W.6

(Illustration à la page 695)

La société LEMCO produit maintenant deux nouvelles séries de condensateurs au polyester métallisé, prévus principalement pour les circuits imprimés, et une gamme de valeurs basée sur l'échelle logarithmique peut être fournie immédiatement.

Les deux gammes diffèrent par le fini et les dimensions, offrant un choix entre des éléments entièrement moulés et des éléments isolés à la résine, répondant ainsi aux spécifications H5DEF 5011 et H3 respectivement. Des fils revêtus de soudure sont utilisés comme terminaisons pour faciliter le soudage durant la production. Le type à moulage de résine a été étudié pour répondre à des conditions climatériques et vibratoires sévères. Il comprend donc un dispositif de protection contre l'humidité qui assure un montage très ferme.

Une gamme de capacitance est prévue de 0,01 μF à 0,22 μF avec des tolérances normales de $\pm 5\%$ $\pm 10\%$ $\pm 20\%$ à une tension de régime de 160 V c.c. A 20°C, la résistance d'isolement est supérieure à 30 k$M\Omega$ et le facteur de puissance est inférieur à 0,0075 lorsqu'il est mesuré à 1 kHz. La gamme de température de fonctionnement s'étend de -55°C à $+85$°C.

Les matériaux utilisés dans la construction de ces condensateurs ont été précontrôlés puis bobinés sur des machines entièrement enfermées. Enfin, avant l'encapsulage, ces matériaux sont stabilisés par traitement thermique contrôlé ce qui garantit des composants d'une grande fiabilité. Les terminaisons d'extrémité ont été renforcées avec un soin spécial et elles ont donc l'avantage d'un facteur réduit d'inductance et de puissance.

EE 86 765 *pour plus amples renseignements*

EXCITATRICES DE VIBRATIONS

Derritron Electronic Vibrators Ltd, 24 Upper Brook Street, London, W.1

(Illustration à la page 696)

La société Derritron Electronic Vibrators Ltd produit maintenant deux petites excitatrices électromagnétiques de vibrations. Ces excitatrices de vibrations utilisent des aimants permanents et ont été conçues pour le contrôle à haute fréquence de sous-assemblages et de composants. Comme sur tous les vibrateurs Derritron, à l'exception du modèle à poussée de 1,13 cm, la nouvelle excitatrice comprend une table en alliage de magnésium à grand diamètre, munie de cinq points de fixation des pièces soumises au contrôle. Les vibrateurs portent les références type VP4 et type VP4B. Ils sont identiques, sauf que le modèle VP4B est fourni avec un ventilateur qui assure le refroidissement supplémentaire de la bobine d'entraînement et il peut donc être entraîné par un amplificateur plus puissant et fournir une plus grande poussée. La fréquence à laquelle la poussée totale peut être fournie va de 5 Hz à 8900 Hz et le vibrateur peut être utilisé avec profit jusqu'à 10 kHz et au-dessus. La première table principale de résonance est à 8900 Hz et la poussée pouvant être obtenue du VP4 et de l'amplificateur 100WLF est de 10,88 kg et celle du VP4B et de l'amplificateur 250WLF est de 15,41 kg. Le poids de la table est de 0,41 kg. Ces excitatrices de vibrations peuvent être fournies avec ou sans porte-tourillons. Lorsque le vibrateur est fourni avec le porte-tourillons type T4 il peut être actionné dans la direction verticale ou dans la direction horizontale ou dans n'importe quelle position intermédiaire.

EE 86 766 *pour plus amples renseignements*

COMMUTATEUR DE PROGRAMMATION

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustration à la page 696)

La société Sealectro Ltd fournit maintenant un nouveau commutateur de programmation, du nom de Actan, comprenant des commandes de moteur bidirectionnelles à solénoïdes et à minutage. Ces commandes peuvent atteindre des vitesses de 1 tour/min et 1 tour/heure bien que d'autres vitesses puissent être prévues sur demande. La commande de 1 tour/heure est utilisée pour entraîner le commutateur dans la direction avant tandis que la commande de 1 tour/min assure une fonction rapide de "retour à zéro" sans devoir passer par la rotation complète de tambour de 360°. Cette caractéristique est particulièrement utile pour les opérations de contrôle lorsque l'essai doit être terminé aussitôt qu'une panne est indiquée car il faut à ce moment là recommencer l'essai.

Des tambours de programme comprenant jusqu'à 120 positions et 51 stations de contact sont offerts avec ce dispositif d'entraînement et ils sont entièrement réglables sans outils. Des contacts de type standard sont fournis sous la forme C et leur puissance nominale est de 2 A. Il existe également des contacts spéciaux d'une puissance de 10 ou de 20 A, 115 V c.a.

EE 86 767 *pour plus amples renseignements*

QUOTIENTMÈTRE TACHYMÉTRIQUE

Advance Controls Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustration à la page 696)

La société Advance Controls Ltd vient de publier des détails d'une nouvelle et importante gamme de tachymètres numériques et de quotientmètres tachymétriques qui offrent un affichage numérique clair et de grandes dimensions en variante à l'affichage au Dékatron.

Seize modèles standard seront livrés prochainement, les variations dans la spécification comprenant soit 4 soit 5 décades et l'emploi d'un oscillateur piézoélectrique interne ou l'emploi de la fréquence d'alimentation comme référence de temps.

La gamme étendue d'applications de ces instruments comprend la mesure précise de la fréquence, la mesure de la vitesse de machine ou d'arbre ou la mesure du rapport entre deux vitesses à trois décimales. La technique de ce quotientmètre permet de mesurer le pourcentage d'étirage ou le pourcentage de traction, à 0,1 % de précision, dans les industries textiles, plastiques, des métaux à base de papier ou d'autres industries à processus continu.

Les signaux d'entrée s'obtiennent habituellement des transducteurs magnétiques mais d'autres dispositifs peuvent être utilisés y compris les générateurs tachymétriques, les détecteurs de proximité ou les appareils photoélectriques.

EE 86 768 *pour plus amples renseignements*

COMMUTATEUR À LAME VIBRANTE À FAIBLE BRUIT

Flight Refuelling Ltd, Wimborne, Dorset

(Illustration à la page 696)

Le commutateur miniature à lame vibrante, MSMF-2, a des propriétés thermiques de force électromagnétique d'environ 10% de celle des commutateurs à lame vibrante classiques. Il comporte un matériau de contact spécifiquement conçu pour les circuits à niveau réduit.

Les résistances de contact extrêmement vastes et stables, alliées à un entrefer minimum, réduisent considérablement le bruit produit mécaniquement. Ces caractéristiques, ajoutées à la propriété de force électromagnétique thermique réduite, font du MSMF-2 un instrument particulièrement indiqué pour les applications à circuit sec ou à faible niveau.

Longueur hors-tout: 2,25 pouces
Longueur du verre: 0,80 pouces max.
Diamètre: 0,09 pouces max.
Force électromagnétique thermique: Environ 5 μV/°C
Durée de vie: Dépassant 10×10^6 opérations sur la puissance nominale totale. Durée infinite sous charges de circuit sec.
Indice nominal de contact maximum: 28 V c.c., 10 mA.

EE 86 769 *pour plus amples renseignements*

BORNE DE RÉCEPTION À DOUBLE DIVERSITÉ

Racal Communications Ltd, Western Road, Bracknell, Berkshire

(Illustration à la page 696)

La borne de réception à double diversité, Racal RA.133, est maintenant en cours de production. Cet instrument a été étudié à l'intention des services de radiodiffusion en n'importe quelle partie du monde, pour la réception de programmes d'émissions musicales ou de parole dans les bandes d'émission internationale haute fréquence, qui peuvent être ensuite retransmises par transmetteurs locaux ou réseaux de relais de distribution. Un certain nombre de ces bornes a été fourni à la BBC.

La nouvelle borne de réception courve une gamme de fréquence de 1 à 30 MHz ou de 2 à 24 MHz, avec présélection haute fréquence supplémentaire. Les circuits de présélection assurent une plus grande sélectivité de sorte qu'un signal d'interférence hors-accord de 5 % est atténué dans une mesure d'environ 100 dB.

La borne comprend également la réception à double diversité de signaux latéraux uniques ou doubles. On peut choisir soit la bande latérale supérieure soit la bande inférieure; dans le fonctionnement à bande latérale double, cette caractéristique permet d'utiliser la bande latérale ayant le moins d'interférences.

Deux récepteurs RA.117C sont incorporés à la borne RA.133 garantissant la stabilité et la précision du réglage. Les récepteurs RA.117 peuvent être accordés séparément ou utilisés dans un rapport de récepteur principal à récepteur asservi, le second signal de l'oscillateur à fréquence variable de l'un des récepteurs contrôlant le réglage en kilocycles des deux récepteurs. Cette méthode de fonctionnement simplifie l'accord et le réglage de fréquence précis des deux récepteurs est donc assuré.

Chaque parcours de réception est muni d'un atténuateur d'entrée à cinq réglages de commutation, et cinq passe-bandes de fréquence intermédiaire entre 1,2 kHz et 13 kHz sont incorporés. La combinaison en diversité des signaux supérieurs, inférieurs ou de double bande latérale des deux parcours de récepteurs s'effectue dans des réseaux passifs et des dispositifs de contrôle visuel et oral complets sont prévus.

EE 86 770 *pour plus amples renseignements*

OUTIL DE SERRAGE DE CÂBLES

Hellerman Electric Ltd, Gatwick Road, Crawley, Sussex

(Illustration à la page 697)

La société Hellerman Electric Ltd a réalisé un nouvel outil pour la ligature des câbles, le Tyton, qui ferait économiser environ 50 % de temps et de frais par rapport à d'autres méthodes de ligature permanente.

Le Tyton est un appareil à cartouches avec 20 agrafes pour pouvoir effectuer 20 ligatures sans arrêt et, en raison de la force de chacune des ligature, un nombre inférieur de ligatures est nécessaire. La force supérieure des ligatures s'obtient par la fixation, c'est à dire par déformation, l'attache en nylon n'étant pas percée et n'étant donc pas affaiblie.

L'attache en nylon fournie sur rouleau est automatiquement coupée lorsque la ligature est achevée. On évite ainsi de gaspiller les attaches et on dispose d'un outil prêt à l'emploi à n'importe quel moment. Cet avantage permet de lier n'importe quelle grandeur de câble à n'importe quel moment sans interruption ou retard.

Une courroie de support, portée autour du cou, est fournie avec l'outil pour permettre à l'opérateur de garder les deux mains libres lorsqu'il ne lie pas des câbles.

EE 86 771 *pour plus amples renseignements*

EQUIPEMENT DE CONTRÔLE TÉLÉPHONIQUE

Telefon Fabrik Automatic A/S, 7 Amaliegade, Copenhague, Danemark

(Illustration à la page 697)

Ce matériel de contrôle téléphonique a été conçu pour les essais électriques d'appareils téléphoniques complets, d'appareils manuels, d'émetteurs téléphoniques, de récepteurs et d'appareils de sonnerie. Les essais pouvant être effectués comprennent la mesure de l'efficacité de réception, d'émission et de tonalité latérale, de bruit microphonique, d'impédance, d'appel, et de tension et de courant continu.

Ce matériel a été conçu essentiellement en vue des essais de production et de réception, mais sa précision est suffisante pour justifier son emploi en laboratoire, et son fonctionnement est suffisament simple pour pouvoir être exécuté par un personnel non qualifié après un minimum de formation.

L'équipement de contrôle comprend deux sections. La section électronique, EJ5-01, comprend les circuits électroniques, les commandes et un instrument de mesure. La section acoustique, EJ5102 comprend la tête artificielle, une oreille artificielle, une bouche artificielle et un calibre de contrôle dans lequel pratiquement n'importe quel écouteur téléphonique peut être introduit et fixé dans la position voulue. Les deux sections sont reliées entre elles par un câble.

Durant l'essai, l'appareil téléphonique est relié au matériel au moyen d'un circuit d'alimentation en tension continue incorporé au EJ5-01. Le circuit d'alimentation se compose d'une source de courant stabilisé incorporée de 24 V ou 24 V, d'une bobine d'alimentation et d'une résistance série variable pour pouvoir obtenir le courant de ligne normal. Le cas échéant, l'appareil téléphonique peut être alimenté par ligne de transmission artificielle comme, par exemple, la ligne TFA type EG4.

Les propriétés de transmission sont mesurées en utilisant le bruit blanc pondéré au lieu de la parole. Il a été démontré que cette méthode donne des résultats équivalents à ceux qu'on obtient au moyen des essais d'intensité sonore voix/oreille. La pondération du bruit blanc consiste en une atténuation augmentant de 6 dB par octave en fonction de la fréquence. Ainsi le spectre de fréquence du bruit acoustique est à peu près le même que celui d'une voix humaine moyenne.

EE 86 772 *pour plus amples renseignements*

INSTRUMENTS DE MINUTAGE

System Engineering Services Ltd, Radley House, 35-39 South Ealing Road, London, W.5

(Illustration à la page 697)

La société System Engineering Services Ltd fournit maintenant des instruments de minutage qui indiquent les jours, les heures, les minutes, et les secondes.

L'indicateur EECO 869 est livrable en deux modèles; le modèle 869-6 indique les heures, les minutes et les secondes. Le modèle 869-9 indique les jours, en plus de l'information affichée par le modèle 869-6. Jusqu'à trois indicateurs peuvent être actionnés par un seul générateur de code de temps EECO.

La fenêtre d'affichage comprend un écran lumineux polarisé qui rend l'indication éclairée Nixie claire et distincte. La durée de vie nominale des tubes Nixie est de 200 000 heures de fonctionnement.

L'instrument de minutage 869 peut être monté sur bâti ou sur support spécial pour montage sur mur, pupitre ou plafond. Les deux modèles mesurent 8,9 cm de haut, 48,26 cm de large et 15,5 de profondeur, connecteur de couplage compris.

La gamme de température de l'instrument va de −20°C à +55°C.

Les autres indicateurs de temps EECO, à savoir le modèle à entrée parallèle EECO 865 et le modèle à entrée en série EECO 866 ou 867, sont utilisés principalement pour fournir une indication à distance du temps produit par un générateur de code de temps.

EE 86 773 *pour plus amples renseignements*

APPAREILS À DÉNUDER LES FILS MÉTALLIQUES

Light Soldering Developments Ltd, 28 Sydenham Road, Croydon, Surrey

(Illustration à la page 697)

L'appareil "Adamin", modèle 2B6, est un nouveau dispositif miniature thermique pour dénuder les fils métalliques. Il peut être utilisé avec le polythène au chlorure de polyvynile, le nylon et d'autres matériaux d'isolement analogues à point de fusion bas.

L'appareil à dénuder Adamin peut être utilisé avec une seule main, par l'emploi de pinces. Il peut dénuder des revêtements d'un diamètre d'environ 0,39 cm sans risque d'endommager les conducteurs.

Le modèle 2B6 peut être fourni pour 12 ou 24 V. La consommation électrique est d'environ 14 W, ce qui assure une chaleur suffisante pour faire fondre le revêtement. Une chaleur supérieure produit une fusion excessive et peu causer des bavures du revêtement liquide sur la partie dénudée du fil métallique.

N'importe quelle source de courant continu ou de courant alternatif de 12 ou de 24 V (suivant la puissance nominale de l'appareil à dénuder) fournit un courant de 1 A ou de 0,5 A respectivement. En variante, les transformateurs d'isolement "Litesold", type LT12 (12 V, 25 W), LT24 (24 V, 25 W) ou MT24 (24 V, 55 W), permettent d'utiliser l'appareil à partir de n'importe quelle prise de courant secteur, avec l'avantage supplémentaire que les trois prises de sortie thermique de ces transformateurs permettent de régler la température.

Le fil à dénuder doit être fixé fermement entre les lames pendant une ou deux secondes pour leur permettre de fondre à travers le revêtement. L'opération de dénudage est alors complétée en tirant légèrement sur le fil.

Il existe un outil semblable, le modèle 2B24, mais qui fonctionne à une température plus élevée et qui est prévu pour les matières plastiques isolantes thermorésistantes. Cet outil ne fait pas fondre la matière plastique isolante thermorésistante mais l'amollit suffisament pour que les lames puissent la couper.

EE 86 774 *pour plus amples renseignements*

Résumés des Principaux Articles

Un appareil de mesure automatique des co-ordonnées par R. L. Farrel et G. E. Moore

Résumé de l'article aux pages 644 à 651

On obtient des renseignements précis sur les parcours de vol de fusées et autres engins du même genre en mesurant les positions de leurs images par rapport à celle des étoiles sur plaque photographique. A l'aide du comparateur Zeiss, qui est un appareil de mesure optique des co-ordonnées, on peut effectuer des lectures de ces plaques d'une précision supérieure à 1 micron.

L'appareil en question enregistre les mouvements du support de la plaque dans le comparateur et fournit une image visuelle décimale ainsi qu'une sortie sur carte perforée.

Les mesures de deux co-ordonnées s'obtiennent à un degré de précision de 1% en comptant les franges Moire. Les dessins de frange sont lus par des cellules photovoltaïques et amplifiées. Des circuits spéciaux déterminent la direction du mouvement. Les compteurs décimaux utilisés pour accumuler le comptage des franges sont capables de fonctionner à grande vitesse et d'inverser leur mouvement rapidement; ils contiennent certains principes inédits de réalisation. Un système d'alarme spécial, basé sur le concept de décélération critique, détecte les erreurs dues aux pannes de l'appareil électronique ou optique.

L'étude de quelques circuits de commutation à diodes tunnel par R. V. L'Archeveque et B. McA. Sayers

Résumé de l'article aux pages 652 à 655

Les auteurs analysent les circuits à diodes tunnel pour action monostable et bistable (20 mcps) à vitesse moyenne. Ils fournissent des équations de conception pour ces circuits basées sur des calculs de minutage d'un oscillateur de relaxation. Ils décrivent, enfin, un discriminateur à niveau réduit stable.

Production de haute tension à partir de matériaux piézoélectriques en céramique par C. Pearcy

Résumé de l'article aux pages 656 à 659

Les perfectionnements apportés depuis peu aux matériaux piézoélectriques ont permis de produire des sources d'impulsions à haute tension basées sur la transformation directe de l'énergie mécanique en énergie électrique. Ce principe a été exploité dans la production de générateurs haute fréquense à étincelles comportant des cylindres en céramique PZT (titanate zirconate de plomb). Des dispositift de ce genre ont été utilisés avec succés pour l'allumage de divers appareils à gaz, de fusibles d'impac de missiles et de blocs d'alimentation haute tension à faible courant tels que ceux employés dans le matériel à rayons & portatif.

Cet article donne un aperçu des diverses méthodes de réalisation nécessaires à la fabrication de ces dispositifs. Il donne des listes d'équations ainsi que des indications sur les caractéristiques physiques des matériaux en céramique PZT.

Calculatrice analogique de prévision et de contrôle de stocks par R. Calvert et J. Mildwater

Résumé de l'article aux pages 660 à 666

Les progrès réalisés récemment dans la recherche opérationnelle ont donné lieu à des méthodes perfectionnées de prédiction à court terme. La calculatrice dont il est question ici utilise les méthodes de transformation de rapport pour résoudre une série d'équations donnant des renseignement statistiques pour les prévisions et le contrôle de stocks.

La description générale des circuits d'addition, de soustraction, de multiplication et de division est suivie par des détails de l'application du système à un groupe particulier d'équations.

L'antenne sinueuse par T. S. M. Maclean

Résumé de l'article aux pages 667 à 669

L'auteur décrit un mouveau type d'antenne à rayonnement longitudinal et à ondes progressives ayant la forme d'une ligne sinueuse. Cette antenne se compose d'un conducteur formé d'éléments dont la longueur est d'un peu plus d'une demi longueur d'onde et de coudes à angle droit. Le rayonnement longitudinal s'effectue sur une largeur de bande d'environ 12% ayant un niveau de lobe secondaire supérieur à 6dB.

Stimulateur à deux transistors pour utilisations physiologiques par J. Opalinski

Résumé de l'article aux pages 670 à 673

Cet article décrit un stimulateur physiologique compact à deux transistors pouvant être utilisé pour l'enseignement avec couplage électromagnétique par diode Zéner. Un générateur de minutage séparé avec transistor à unijonction fait également partie de l'appareil.

Un réseau en pont T pour le contrôle magnétique du courant alternatif par S. N. Bhattacharyya et J. K. Choudhury

Résumé de l'article aux pages 674 à 677

Le réseau en pont T a certains avantages précis par rapport au pont conventionnel pour le contrôle magnétique à courant alternatif. Ces avantages sont dûs en particulier au fait qu'il est relativement insensible aux effets de couplage capacitif parasites qui se manifestent de façon particulièrement effective aux fréquences acoustiques et élevées. Dans la recherche qui fait l'objet de cet article on a utilisé une configuration du réseau en pont T qui constitue essentiellement un circuit compensé à résonance en série en équilibre. Les résultats des essais sont indiqués afin de montrer l'application du réseau au contrôle magnétique dans différentes conditions d'excitation du spécimen. Des mesures on été effectuées sur des fréquences allant de 50Hz à 40kHz.

Réduction directe d'un circuit à mailles multiples à un réseau à quatre bornes par D. W. W. Rogers

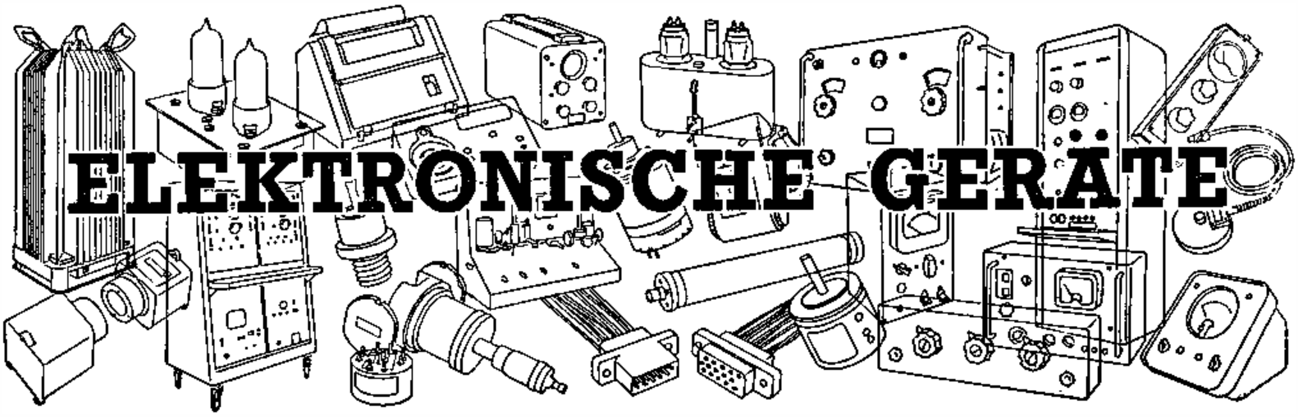
Résumé de l'article aux pages 678 à 680

Il s'agit ici d'une méthode pour obtenir les équations à quatre bornes d'un réseau à mailles multiples. Cette méthode est basée sur une technique suivant laquelle les paramètres d'impédance (ou d'admittance) de n'importe quel réseau peuvent être inscrits directement sous forme d'impédance réfléchie. Une méthode analogue basée sur l'analyse du flux de signaux a été employée par Robichaud.

Un détecteur à diode sensible aux phases par B. S. Leela et R. Shankar

Résumé de l'article aux pages 681 à 683

Un circuit détecteur à diode sensible aux phases est analysé en détail du point de vue de la non-linéarité. Les solutions d'équations obtenues par calculatrice on donné lieu à de nouveaux critères pour le choix de valeurs de composantes. Il est démontré, en particulier, que des chiffres de non-linéarité considérablement inférieurs peuvent être obtenus par des restrictions relativement mineures sur le niveau de signal. Les résultats sont présentés sous forme de tableau à lecture facile.



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben

Übersetzung der Seiten 692 bis 697

Zeitsteuergeräte

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Abbildung Seite 692)

Drei neue, speziell für industrielle Anwendung konstruierte synchronmotorgetriebene Zeitsteuergeräte wurden von Magnetic Devices angekündigt. Alle drei haben das neue Konstruktionsmerkmal, das gestattet, den Zeitschalter bei Speisespannungsausfall entweder in die Ursprungsposition zurückzustellen oder bis nach Wiedereinschalten in der unterbrochenen Position zu halten, wenn die noch nicht verstrichene Zeit abläuft.

Die Baureihe 934 kann die meisten der in der Verfahrenssteuerung und Automatisierung erforderlichen Funktionen durchführen und hat drei Bereiche, die durch Positionierung eines Wahlschalters eingestellt werden. Innerhalb der Skala des eingestellten Bereiches kann die Verzögerung stufenlos geregelt werden.

Die Steuerung erfolgt über einen externen Impulsschalter oder Ein-Ausschalter und ist für die drei Zeitbereiche 0...12, 0...60 und 0...120 s; 0...24, 0...120 und 0...240 s; 0...6, 0...30 und 0...60 min; 0...3, 0...15 und 0...30 h einstellbar. Es werden vier Mikroschalter verwendet, von denen zwei sofort und zwei nach Ablauf der Zeitverzögerungsperiode betätigt werden.

Baureihen 931 und 932 haben einen bzw. zwei Verbraucherkontakte, wobei Mikroumschalter als Verbraucherkontakte Verwendung finden. Im Rahmen der Skala des eingestellten Zeitbereiches ist die Verzögerung stufenlos einstellbar. Steuerung erfolgt mittels eines externen Ein-Ausschalters; die Zeitverzögerungsbereiche sind 0...12, 0...24, 0...60, 0...120, 0...240 s; 0...6, 0...30, 0...60 min; 0...3, 0...15, 0...30 h.

Modelle sind für Grundflächenmontage und bündigen Einbau mit Schraubklemmen- oder Steckeranschluss lieferbar.

In Betrieb werden ein Synchronmotor und eine elektromagnetische Kupplung durch Schliessen eines externen Steuerschalters erregt. Zu jeder Zeit wird die noch nicht verflossene Zeit durch die Position eines schwarzen Zeigers angegeben.

Die Einstellgenauigkeit ist etwa 1,5 Prozent des kompletten Zyklus mit einer Wiederholbarkeit von besser als 1 Prozent. Betriebsspannungen sind 240 V~, 110 V~ oder 24 V~ und die Betriebsleistung etwa 13 VA. Die Überschlagsprüfspannung ist 2500 V.

Alle stromführenden und bewegten Teile werden durch eine transparente staubdichte Haube geschützt, und die Anschlussklemmen sind eingelassen. Der Umgebungstemperaturbereich ist −40° ... +60° C. Geräte der Baureihen 931 und 932 wiegen 850 g, die der Serie 934 wiegen 907 g.

EE 86 751 *für weitere Einzelheiten*

Mit Silikonkautschuk umhüllte Kondensatoren

The Plessey Co. Ltd, Components Group, Kembrey Street, Swindon, Wiltshire

Aus einem neuen Fertigungsprogramm kann der Geschäftsbereich Bauelemente der Plessey Co Ltd mit Silikonkautschuk umhüllte Kondensatoren liefern; man nimmt an, dass es sich dabei um die ersten in Grossbritannien hergestellten Kondensatoren dieser Art handelt.

Mit Silikonkautschuk umhüllte Bauelemente kombinieren einfache Herstellung mit ungewöhnlich hoher Temperaturfestigkeit und vermeiden die bei Metallgehäusen auftretenden Pack- und Isolationsschwierigkeiten. Sie bieten ausserdem bei vielen Anwendungsmöglichkeiten spezielle Vorteile.

Bei grossen Kondensatoren, die für hohe Ströme bemessen sind und bei hohen Umgebungstemperaturen betrieben werden, ist es z.B. wünschenswert, den Strom über mehrere Wickel zu verteilen, was oft zu komplizierten Formen für die Einkapselung führt. Bei Verwendung von Silikonkautschuk besteht hier jedoch kein Problem, da die Wickel nur mit Band zusammengehalten in einem gepressten Kasten geeigneter Form aufgehängt werden. Diese Form wird dann mit katalysiertem Silastomer gefüllt, was nach Aushärten eine dauerhafte Einkapselung ergibt, die ohne Degradierung Temperaturen bis zu 250° C widerstehen kann.

Ein typischer 30-μF-Kondensator, der für 115 V~ mit Stossspannungen bis zu 200 V_{eff} bei 400 Hz bemessen ist, kann in Umgebungstemperaturen bis zu 100°C arbeiten und hat 121 × 51 × 76 mm Aussenabmessungen.

Die Silikonkautschukumhüllung von Glimmerkondensatoren für Betriebstemperaturen von 150° C und darüber ist ein weiteres bedeutendes Anwendungsgebiet. Silikonkautschuk ist praktisch der einzige zufriedenstellende Werkstoff, wenn statt der offenen Bauart Einkapselung erforderlich ist. Glimmerkondensatoren von 1 μF haben Abmessungen von 51 × 44 × 38 mm und arbeiten bei Betriebsspannungen von 350 V— ohne Schädigung in Temperaturen bis zu 250° C.

Die herkömmlichen Anschlüsse werden bei Silikonkautschuk-Einkapselung normalerweise durch freie Zuleitungen oder verzinnte Kupferdrähte ersetzt. Für Temperaturen über 150°C finden Feinsilberdrähte Verwendung. Trotzdem dieses Verfahren nicht denselben Feuchtigkeitsschutz wie Metallgehäuse gibt, da man die Durchführung der Zuleitungen nicht abdichten kann, spielt das selten dort eine Rolle, wo sonst die Kautschukumhüllung wünschenswert ist.

EE 86 752 *für weitere Einzelheiten*

Oszillator und Pegelmessgerät

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(*Abbildung Seite 692*)

Der Geschäftsbereich Prüfgeräte der Standard Telephones & Cables Ltd gibt die Einführung eines komplementären Paares tragbarer, batteriebetriebener Geräte für die Erzeugung von Signalen und das Messen von Pegeln im Frequenzbereich 300 Hz...300 kHz bekannt. Sie sind hauptsächlich für Anwendung in Tonfrequenz-, Rundfunk- und Mehrkreisfernsprechsystemen bestimmt.

Der Ausgangspegel des Oszillators 74254-A ist in 1-dB-Stufen von 0... −49 dBm an 75, 125, 140 und 600-Ω-Schaltungen regelbar. Der Abschwächer hat zwei Bedienelemente: von 0...40 dB in 10-dB-Stufen und von 0...9 dB in 1-dB-Stufen. Ein Pegeljustierregler und ein zugehöriges, in ±0,25 dBm geeichtes Messgerät sind vorhanden. Die Genauigkeit des Ausgangspegels ist bei 0 dB Leistung (20° C) über den gesamten Frequenzbereich ±0,1 dB, jedoch kann bei anderen Einstellungen des Abschwächers ein zusätzlicher Fehler von 0,2 dB auftreten. Änderungen der Umgebungstemperatur zwischen 10 und 40° werden Fehler von nicht mehr als ±0,2 dB hervorrufen. Der Klirrfaktor des Ausgangs wird 1 Prozent nicht überschreiten. Die Frequenzskala hat 127 mm Durchmesser und ist nur einmal für Werte von "300" bis zu "3000" graviert; der Bereichsschalter hat Faktoren von ×1, ×10 und ×100, wodurch Frequenzbereiche von 300...3000 Hz, 3...30 kHz und 30...300 kHz entstehen. Die Frequenzunsicherheit ist ±1 Prozent. Der Oszillator entnimmt der aus acht 1,5-V-Trockenbatterien bestehenden 12-V-Stromversorgung 50 mA. Er wiegt mit Batterien 5,4 kg und hat Abmessungen von 311 × 203 × 181 mm.

Das Pegelmessgerät 74255-A wurde zusammen mit dem Oszillator 74254-A und für denselben Frequenzbereich entwickelt. Der Messumfang des Gerätes ist +21 dBm...−51 dBm; es kann mit symmetrischen oder unsymmetrischen Schaltungen von 75 Ω, 125 Ω, 140 Ω und 600 Ω eingesetzt werden. Man kann sowohl durchgehende wie abgeschlossene Pegel messen, und bei der Temperatur, bei der die Eichung vorgenommen wurde, kann eine Messgenauigkeit von ±0,25 dB erwartet werden. Durch Anwendung von Gegenkopplung wird nicht nur ein linearer Frequenzgang erreicht, sondern auch eine hochohmige Speisung des Messgerätes, so dass die 115 mm lange Skala eine fast lineare Form hat. Das Messinstrument ist in Stufen von 0,2 dB von +1 dB...−11 dB geeicht und arbeitet mit einem umschaltbaren Regler zusammen, wodurch Überdeckung des gesamten Eingangspegelbereiches von 72 dB ermöglicht wird.

Das Pegelmessgerät wiegt bei 311 × 203 ×181 mm Aussenabmessungen 5,7 kg. Es wird wie der Oszillator aus acht Trockenbatterien von 1,5 V gespeist, denen es 40 mA entnimmt.

EE 86 753 *für weitere Einzelheiten*

Leiterplatten

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(*Abbildung Seite 693*)

Auf Veranlassung einiger Hersteller elektronischer Geräte wird "Veroboard" jetzt mit 1,27 mm (0,05") Rastergrundmass produziert. Es eignet sich besonders als "Mutterplatte" für integrierte Schaltungen, ohne dass man zu dem äusserst hohen Aufwand und den Verzögerungen, die mit der herkömmlichen Erzeugung von Druckschaltungsplatten zu den für Mikrominiatur-Bausteinen erforderlichen engen Toleranzen gezwungen wird. Dieses neue Veroboard wird aus 0,78 mm starkem Hartpapier mit 0,04 mm dicken und 0,9 mm breiten Kupfer- oder Kupfer-Nickelstreifen auf einer Seite hergestellt, in die Löcher von 0,64 mm Durchmesser in einer Matrize von 1,27 × 1,27 mm gestanzt werden. Veroboard gibt es in quadratischen Platten von 152 × 152 mm (6 × 6") oder als Steckkarten für Federleisten mit 1,27 mm Grundmass, die von Ultra Electronic (Components) Ltd hergestellt werden.

EE 86 754 *für weitere Einzelheiten*

Drehschalter

I.D.M. Electronics Ltd, 1 Arkwright Road, Reading, Berkshire

(*Abbildung Seite 693*)

Eine Auswahl von Hand betätigter Drehschalter wurde entwickelt, um dem Bedarf der Instrumenthersteller an Schaltern zu genügen, die während der ganzen Lebensdauer des Instrumentes zuverlässig und in ihrer Leistung gleichförmig bleiben.

Die Schalterebenen sind auf glasgefüllten Alkydpresslingen aufgebaut, die gute elektrische und mechanische Festigkeit, Masshaltigkeit und feuchtigkeitshemmende Eingenschaften haben. Eine Kontaktarm-Spezialkonstruktion und die Hartversilberung der Kontakte und des Kontaktarms ergeben zusammen einen niedrigen und konstanten Kontaktwiderstand sowie Rostschutz. Schleppkontakt-Typen und nichtschleppende Typen sind lieferbar; jede Ebene kann bis zu 12 Schaltstellungen haben. Die Kontakte sind zum Schutz gegen Staub und Verschmutzung umschlossen.

EE 86 755 *für weitere Einzelheiten*

Hebelschalter

Shawford Control Gear Co. Ltd, 715 Tudor Estate, Abbey Road, Park Royal, London, N.W.10

(*Abbildung Seite 693*)

Shawford Control Gear können nunmehr Hebel-(Steuerknüppel-) Wahlschalter in neuer Zwei- und Vierwegeausführung liefern. Das neue Modell hat eine Sicherungshülse, die man von Hand anheben muss, ehe man den Schalter betätigen kann, wodurch zufällige Betätigung verhindert wird.

Die Hülse ist für die Sperrposition vorgespannt; der Schalter wird durch Anheben einer mit dem Hebel konzentrischen Scheibe freigegeben, wonach jede gewünschte Einstellung vorgenommen werden kann. Der auf diese Weise hergestellte Kontakt kann entweder vorübergehend oder auf Wunsch dauernd bestehen.

In diesem Modell—sowie in allen weiteren Modellen bestehender Hebelwahlschalter—wurde die Nockenbetätigung der Kontaktblockkolben durch vertikal bewegliche Betätigungsorgane ersetzt, wodurch Seitenbelastung beseitigt und die sich daraus ergebende Gefahr des Klebens verhindert wird.

Der Schalter ist mit zweipoligen Kontaktblöcken in Standardausführung ausgerüstet; man kann jedoch noch ein weiteres Paar montieren und damit die Kontaktzahl verdoppeln. Die Nennbelastung ist 4 A bei 220 V~ mit 550 V~ Höchstspannung.

EE 86 756 *für weitere Einzelheiten*

Impedanzanpassungssonde

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

(*Abbildung Seite 694*)

Kynmore Engineering Co. Ltd kann jetzt eine in sich geschlossene Impedanzanpassungssonde liefern, die gestattet, niederohmige Prüfgeräte wie z.B. Voltmeter oder Oszillografen ohne durch Belastung verursachte Fehler zu verwenden. Das von der Firma Adams-Norken Ltd in Swindon hergestellte Gerät Modell EPO-1 besteht aus einem Verstärker mit 1:1 Verstärkung, der zusammen mit seinem Ein-Ausschalter und zwei ersetzbaren Batterien dicht in einem Spritzgusskörper verschlossen und mit doppelt abgeschirmten Zuleitungen versehen ist.

Ein Merkmal dieses Tasters ist sein hochohmiger Eingang von 8 MΩ parallel mit nur 1 pF. Die Ausgangsimpedanz ist 200 Ω in Serie mit 50 μF, was ein Verhältnis von 40 000:1 ergibt. Der Dynamikbereich ist 4 V_{SS} und die Bandbreite 10 Hz...3 MHz. Die Sonde stellt daher eine bequeme Anpassungseinheit für allgemeine Laborarbeiten dar und ist für jede Anwendung geeignet, die hohe Eingangs- und niedrige Ausgangsimpedanz erfordert. Die Linearität ist besser als 1,0 Prozent und das mit einer Impedanz von 100 kΩ gemessene Rauschen 20 V_{SS}. Der bei 1 V, 10 kHz gemessene Klirrfaktor ist niedriger als 0,25 Prozent. Die Einheit wurde Klimatests nach DEF 133 unterworfen.

Der Sondenkörper hat 76,2 × 19 × 22,2 mm Abmessungen; die beiden in Serie geschalteten Mallory-Batterien geben bei Dauerbetrieb 80 Stunden Lebensdauer. Üblicherweise wird die Sondenzuleitung mit einem doppelt abgeschirmten BNC-Stecker geliefert, doch können modifizierte Sonden mit unterschiedlichen Zuleitungen oder technischen Daten hergestellt werden.

Zum Verdrahten in Geräte oder für Anwendung als Steckvorsatz ist eine verkapselte Standardausführung EPO-2 lieferbar, die ohne Batterien oder Kopplungskondensatoren in einem Aluminiumgehäuse von 19 mm Durchmesser und 38,1 mm Länge untergebracht ist. Sie kann mit Lötstiften, Druckschaltungsstiften oder Röhrensockelstiften geliefert werden.

EE 86 757 *für weitere Einzelheiten*

Steckverbindung für integrierte Schaltungen

Ferranti Ltd, Kings Cross Road, Dundee, Schottland

(Abbildung Seite 694)

Ferranti Ltd hat eine neue Steckverbindung S.8 eingeführt, die besonders für Aufnahme der in Gehäusen TO-5 mit acht Zuleitungen verkapselten integrierten Schaltungen entwickelt wurde.

Die Steckverbindung S.8 ist in erster Linie für gedruckte Schaltungen gedacht und wird in Prototypgeräten, Entwicklungsprojekten und Testeinrichtungen besonders zweckdienlich sein.

Zwei besondere Merkmale der Steckverbindung S.8 sind das geringe Gewicht von nur 1 g und die Verwendung von schwachen Federn für die Kontakte, wodurch die Kontaktkraft genau kontrolliert werden kann. Die Kontakte sind aus Beryllium-Kupfer hergestellt und haben eine 5 Mikron starke Hartgoldplattierung.

Zur Erzielung von Massbeständigkeit und ausgezeichneten elektrischen Eigenschaften wird der Körper der Steckverbindung S.8 aus glasgefülltem Nylon gepresst; die korrekte Einführung des TO-5 wird durch eine Nase am Körperrand gewährleistet, die mit der Zunge des TO-5 ausgerichtet wird.

Typische Daten für die Steckverbindung S.8 sind:

Isolationswiderstand
  1×10^7 MΩ
Kontaktwiderstand
  40 mΩ (zu denen die Zuleitungen der integrierten Schaltung ca. 10 mΩ beitragen)
Selbstinduktivität jedes Kontaktes
  0,001 μH
Kapazität zwischen benachbarten Kontakten
  0,2 pF
Einsteckkraft
  600 p
Temperaturbereich
  −55° C...+120° C

EE 86 758 *für weitere Einzelheiten*

Kabelmarkierungssystem

Label Processes Ltd, 17-19 Crescent, Salford 5

(Abbildung Seite 694)

Label Processes Ltd hat ein schnelles, einfaches und wirtschaftliches Verfahren zur Kennzeichnung von Kabeln und Drähten eingeführt. Das System besteht aus einem Kunststoffschlauch in Rollenform, der bedruckt und halb durchgeschnitten ist. Da der Schlauch nur halb durchgeschnitten wird, bleiben die bedruckten Abschnitte in Rollenform, bis sie benötigt werden; jeder bedruckte Schlauchabschnitt lässt sich jedoch leicht abtrennen.

Der abgetrennte Abschnitt wird einfach auf das Kabel oder den Draht geschoben und kennzeichnet damit schnell und klar jedes Kabel oder jeden Draht. Man kann Nummern, Symbole und Information jeder Art auf den Schlauch aufdrücken; durch Verwendung verschiedenfarbiger Kunststoffschläuche entsteht ein kombiniertes Informations- und Farbkennzeichnungssystem. Der Schlauch wird auf der Maschine LP70 (eins der verschiedenen von Label Processes Ltd hergestellten Modelle) automatisch bedruckt und geschnitten, und zwar 3000 Abschnitte pro Stunde. Das Drucken erfolgt nach dem Wärmeeindruckverfahren und ist daher völlig trocken, schmierfrei und abriebfest, so dass der bedruckte Schlauch sofort Verwendung finden kann.

Die LP70 kann auch aufeinanderfolgende Nummern oder veränderliche Informationen auf Flächen bis zu 76 × 51 mm drucken und kann von ungelernten Kräften bedient werden.

EE 86 759 *für weitere Einzelheiten*

Zweischreibwerk-X-Y-Schreiber

Hewlett Packard, Dallas Road, Bedford

(Abbildung Seite 694)

Die Abteilung Moseley von Hewlett-Packard hat einen neuen X-Y_1-Y_2-Tischschreiber eingeführt. Das neue Modell 2FA ist ein Seitenstück des bereits früher angebotenen Modells 2FRA, das es nur für Gestelleinbau gibt. Dieser Präzisionsschreiber mit Abmessungen von etwa 28 × 43 cm hat zwei in der Vertikalachse unabhängig voneinander und in der Horizontalachse integralgesteuerte Schreibwerke. Es ergeben sich also drei Kanäle, die gleichzeitiges Aufzeichnen von zwei abhängigen Variablen gegen eine unabhängige Variable gestatten. Die horizontale Trennung der beiden Schreibwerke ist 2,54 mm. Gegen Zuschlag kann man ein drittes feststehendes Schreibwerk anbauen lassen, das Markieren von kennzeichnenden Vorgängen während des Registrierens gestattet.

Jede Achse des 2FA hat 11 geeichte Gleichspannungs-Eingangsbereiche mit Empfindlichkeiten von 0,5 mV/Zoll bis zu 50 V/Zoll (ca. 0,2 mV/cm bis zu 20 V/cm). Durch Betätigung eines Schalters kann man für jeden Bereich kontinuierlich regelbare Skalenfaktoren einstellen, da in dieser Position der Vollausschlag des Gerätes willkürlich einstellbar ist. Ausserdem können die vier empfindlichsten Bereiche auf potentiometrischen Eingang umgestellt werden. Der Eingangswiderstand ist in allen geeichten Bereichen bei Null 1 MΩ.

Modell 2FA hat eine eingebaute Zeitbasis mit fünf Teilbereichen, die jedoch nur für die X-Achse Anwendung finden kann. Sie gibt Ablenkungen von 0,5 s/Teilung bis zu 50 s/Teilung, was einem Vollausschlag des Schreibwerkes von 7,5 ... 750 s entspricht. Die Zeitbasis gibt eine unabhängige Variable für das gleichzeitige Auftragen von zwei Kurven.

Weitere Merkmale sind die elektrostatische Einrichtung "Autogrip", die das Papier gegen die Auflage hält, sowie die kontinuierliche elektronische Bezugsangabe und 92 Prozent Genauigkeit für alle Bereiche.

EE 86 760 *für weitere Einzelheiten*

Grossbereichoszillator

General Radio Co. (U.K.) Ltd, Bourne End, Buckinghamshire

(Abbildung Seite 694)

Der neue Oszillator 1310-A der General Radio Company ist eine transistorisierte Mehrzweck-Signalquelle mit Konstant-Ausgangssignal, die den Frequenzbereich 2 Hz ... 2 HMz in sechs Dekadenbändern überstreicht. Der sehr niedrige Klirrfaktor (unter 0,25 Prozent zwischen 50 Hz und 50 kHz) ist unabhängig von der Belastung; die hohe Ausgangsspannung (20 V im Leerlauf) und die Möglichkeit, phasenstarr mit einem Frequenznormal zu arbeiten, sind weitere Eigenschaften, die den neuen Oszillator als zweckdienliche Signalquelle für das Labor kennzeichnen.

Das Gerät hat einen Oszillator in RC-Wien-Brückenschaltung mit Kapazitätsabstimmung, der einen überbrückbaren Endverstärker mit niedrigem Klirrfaktor treibt. Die Ausgangsleistung ist 160 mW an 600 Ω.

EE 86 761 *für weitere Einzelheiten*

Lochstreifenstanzer

Automatic Punched Tape Ltd, Commerce House, Commercial Road, Paddock Wood, Kent

(Abbildung Seite 695)

Die Geschwindigkeit des Stanzers P30 für 5-, 6-, 7- oder 8-Kanal-Lochstreifen ist 30 Zeichen je Sekunde.

Die Parität kann durch Kontakte im Stanzkopf automatisch geprüft werden. Die Kontakte zeigen auch an, dass das Papier tatsächlich durchlocht wurde. Modell P30 ist mit Rückwärtsvorschub ausgerüstet, und der Streifen lässt sich auch manuell vorwärtsschieben. Die Aufwickelrolle wird durch einen Wechselstrommotor getrieben, und die Ablaufrollen haben eine Lochstreifenkapazität von je 305 m.

Der Stanzer ist für Betrieb mit 24 oder 48 V— lieferbar. Mit dem Standardmodell wird eine schallschluckende Metallhaube mitgeliefert, die man zum Nachfüllen mit neuem Streifenvorrat einfach entfernen kann.

Auf Wunsch kann die Stanzapparatur allein für Einbauzwecke geliefert werden. Zusätzlich erhältlich gegen Aufschlag sind u.a. ein Tintenmarkierer für das Lochprüfersystem und ein 6-Kanal-Stanzblock für Fernschreiber.

EE 86 762 *für weitere Einzelheiten*

Fotozellen- und Projektorköpfe

Panelectronics Ltd, Eastern Way, Bury St. Edmunds, Suffolk

(*Abbildung Seite 695*)

Diese neuen Fotoköpfe und zugehörigen Lichtprojektoren der neuen "Moniteye"-Zusatzserie sind für alle universellen fotoelektrischen Aufgabenstellungen mit ausgestrahltem Licht bestimmt und über Entfernungen von etwa 90 cm bis zu 3,65 m wirksam; bei Sonderauftrag sind auch leistungsfähigere Projektoren für grössere Entfernungen erhältlich. Die Baugruppen sind robust konstruiert, vollständig abgedichtet und ausreichend öl- und wasserfest. Die 90 cm langen, abgedichteten Zuleitungen sind rechtwinklig zur Achse des Lichtstrahls herausgeführt und durch flexiblen Metallschlauch geschützt.

Sehr zweckdienlich ist die einstellbare Befestigungsklemme, mit Hilfe deren der Linsentubus starr positioniert wird. Da der Körper leicht vom Linsentubus abmontiert werden kann, wird die korrekte Ausrichtung beim Auswechseln der Lampe nicht gestört.

Für optische Anforderungen über lange wie auch kurze Entfernungen stehen parallele oder konvergente Lichtstrahlen zur Verfügung.

In den Empfängern können vier verschiedene Zellentypen Verwendung finden, und zwar Fototransistoren, Kadmiumselenid- und Kadmiumsulfit-Fotowiderstände, sowie Emissionsfotozellen.

EE 86 763 *für weitere Einzelheiten*

Füllstandanzeiger und -wächter

Thomas Industrial Automation Ltd, Electronic Centre, Deansgate Lane, Altrincham, Cheshire

(*Abbildung Seite 695*)

Der sehr vielseitige "Levelrator" gibt eine laufende Anzeige des Füllstandes oder Inhaltes von fliessfähigen Festsubstanzen und Flüssigkeiten und regelt ausserdem das hohe und/oder niedrige Grenzniveau. Visueller und/oder hörbarer Alarm kann für den hohen wie auch den niedrigen Füllstandpunkt vorgesehen werden; beide Punkte lassen sich für die meisten Anwendungen unabhängig voneinander über die volle Elektrodenlänge einstellen.

Man benötigt nur eine Elektrode für die laufende Anzeige und die Regelung des hohen und/oder niedrigen Füllstandes. Die sinnreiche Konstruktion des Elektrodensystems kann hohe Genauigkeiten und nahezu lineare Skalen gewährleisten.

Der "Levelrator" ist für Anwendung in Behältern aller Grössen und Formen geeignet und kann sowohl mit starren wie auch flexiblen Elektroden eigesetzt werden.

Fernanzeigen sind den Anforderungen des Kunden entsprechend lieferbar und in beliebiger Position im Werk zu installieren. Das Anzeigeinstrument ist von den hohen und/oder niedrigen Regelpunkten völlig unabhängig.

Weitere Einsparungen kann man durch Einsatz des speziell entwickelten "Mehrpunkt"-Systems erzielen, das bei mehr als einem Behälter Zentralisierung der Ausrüstung gestattet.

EE 86 764 *für weitere Einzelheiten*

Metallisierte Polyester-Kondensatoren

London Electrical Manufacturing Co. Ltd, Bridges Place, Parsons Green Lane, London, S.W.6

(*Abbildung Seite 695*)

LEMCO stellt nunmehr in erster Linie für gedruckte Schaltungen bestimmte metallisierte Polyesterkondensatoren in zwei neuen Baumustern her. Ein auf einer logarithmischen Wertskala beruhendes Sortiment ist sofort lieferbar.

Die beiden Baumuster unterscheiden sich in bezug auf Oberflächenbehandlung und Grösse und gestatten die Wahl zwischen völlig umpressten und kunststoffisolierten Bauelementen, die den Anforderungen der Klasse H5 bzw. H3 der britischen Vorschrift DEF 5011 genügen. Als Anschlüsse verwendete verzinnte Drähte erleichtern das Löten in der Fertigung. Die kunststoffumpressten Ausführungen widerstehen den strengen Klima- und Schüttelbeanspruchungen und sind für diese Zwecke mit einer als Feuchtigkeitsfangstelle dienenden Leiste ausgerüstet, die auch sehr feste Montage gewährleistet.

Kondensatoren sind in Werten von 0,01 μF bis zu 0,22 μF mit Standardtoleranzen von ± 5 Prozent, ± 10 Prozent und ± 20 Prozent bei einer Betriebsspannung von 160 V— lieferbar. Der Isolationswiderstand ist bei 20°C höher als 30 MΩ und der Leistungsfaktor bei 1 kHz gemessen geringer als 0,0075. Der Betriebstemperaturbereich ist $-55°$ ... $+85$°C.

Die in der Herstellung verwendeten Werkstoffe werden erst auf Fehler geprüft und dann auf völlig umschlossenen Maschinen gewickelt. Abschliessend werden sie von der Einkapselung durch geregelte Wärmebehandlung stabilisiert, woraus sich ein sehr zuverlässiges Bauelement ergibt. Besondere Aufmerksamkeit wurde der Versteifung der Anschlüsse gewidmet, wodurch sich ausserdem durch niedrige Induktivität und niedrigen Leistungsfaktor Vorteile ergeben.

EE 86 765 *für weitere Einzelheiten*

Schwingungserreger

Derritron Electronic Vibrators Ltd, 24 Upper Brook Street, London, W.1

(*Abbildung Seite 696*)

Derritron Electronic Vibrators Ltd stellt jetzt zwei kleine elektromagnetische Schwingungserreger her, in denen Dauermagnete das Magnetfeld bilden. Sie wurden für hochfrequentes Testen von Baugruppen und Bauelementen entwickelt. Wie bei allen Schüttelprüfgeräten von Derritron, mit Ausnahme des Modells für 1,125 kp Schub, wird ein Magnesiumlegierungstisch grossen Durchmessers mit Befestigungspunkten für 5 Prüflinge verwendet. Die Schwingungserreger haben die Typenbezeichnung VP4 und VP4B, sind jedoch identisch mit Ausnahme der Gebläseausrüstung des VP4B, die zusätzliche Kühlung für die Treibspule gibt, so dass sich diese Modelle mit leistungsfähigeren Verstärkern treiben lassen und dadurch eine höhere Schubkraft abgeben können. Der volle Schub kann in dem Frequenzbereich 5 ... 8 900 Hz entwickelt werden, und der Schwingungserreger ist bis zu 10 kHz und darüber nutzbar; die erste grössere Tischresonanz liegt bei 8 900 Hz. Das Modell VP4 kann mit dem Verstärker 100 WLF 10,9 kp Schub abgeben, Modell VP4B mit Verstärker 250 WLF 15,4 kp. Der Tisch wiegt ca. 450 g. Diese Schwingungserreger sind entweder mit oder ohne Zapfenmontage lieferbar. Wenn der Schwingungserreger mit der Zapfenmontage T4 ausgerüstet ist, kann er entweder in vertikaler oder horizontaler Richtung oder in jeder Zwischenlage arbeiten.

EE 86 766 *für weitere Einzelheiten*

Programmierschalter

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(*Abbildung Seite 696*)

Sealectro Ltd kann jetzt einen neuen ACTAN - Programmierschalter liefern, dessen Steuermotorrichtung durch einen Magnetschalter wählbar ist. Diese neue Einheit wird mit 1 UPM und 1 UPH getrieben, auf Wunsch sind jedoch auch andere Geschwindigkeiten lieferbar. Der Schalter wird mit 1 UPH in der Vorwärtsrichtung getrieben; der 1-UPM-Antrieb wird für schnelle Rückstellung auf Null benutzt, so dass der Schalter dafür nicht durch die vollen 360° der Trommelumdrehung zu laufen braucht. Dieses Konstruktionsmerkmal ist hauptsächlich bei Einsatz in Testausrüstungen nützlich, wenn der Test unterbrochen werden muss, sowie sich ein Fehler zeigt, und dann das Programm wieder von vorn beginnen muss.

Programmtrommeln mit bis zu 120 Positionen und 51 Kontaktstationen, die ohne Werkzeuge einstellbar sind, werden mit dieser Antriebsanordnung angeboten. Standardkontakte werden in C-Ausführung und für 2 A bemessen geliefert. Von Underwriters' Laboratory zugelassene Spezialkontakte gibt es auch für 10 oder 20 A, 115 V~.

EE 86 767 *für weitere Einzelheiten*

Tachometer-Quotientenmesser

Advance Controls Ltd, Roebuck Road, Hainault Ilford, Essex

(*Abbildung Seite 696*)

Advance Control Ltd hat Einzelheiten

eines neuen und umfangreichen Programmes für digitale Tachometer und Tachometer-Quotientenmesser bekanntgegeben, das als Alternative zur Dekatron-Darstellung eine grosse, klare Ziffernanzeige einführt.

Es gibt sechzehn kurzfristig lieferbare Standardmodelle, deren technische Daten sich u.a. durch vier oder fünf Dekaden und Verwendung eines internen Quarzoszillators als Bezugsquelle statt der Netzfrequenz unterscheiden.

Unter den zahlreichen Anwendungsmöglichkeiten sind zu erwähnen: genaue Frequenzmessungen und Bestimmen von Wellen- und Maschinengeschwindigkeiten oder dem Verhältnis von zwei Geschwindigkeiten auf drei Dezimalstellen. Mit der Quotientenmessertechnik kann Strecken oder Einziehen von Textilien, Kunststoffen, Papier, Metall oder anderen Materialien in Werken mit kontinuierlicher Fertigung in Prozent gemessen werden.

Die Eingangssignale erhält man üblicherweise von magnetischen Messwertwandlern, es können aber auch andere Geber eingesetzt werden, z.B. Tachogeneratoren, Näherungsmessköpfe oder fotoelektrische Einrichtungen.

EE 86 768 *für weitere Einzelheiten*

Geräuscharmer Zungenschalter

Flight Refuelling Ltd, Wimborne, Dorset

(Abbildung Seite 696)

Das thermische Rauschen des Miniatur-Zungenschalters MSMF-2 ist nur etwa 10 Prozent des in herkömmlichen Zungenschaltern auftretenden; ausserdem ist der Kontaktwerkstoff speziell für Kleinsignalschaltungen ausgewählt.

Äusserst niedrige und konstante Kontaktwiderstände reduzieren zusammen mit einem Minimumspalt das mechanisch erzeugte Geräusch. Mit diesen Eigenschaften und dem sehr niedrigen thermischen Rauschen ist der MSMF-2 besonders für Anwendung in Kleinsignal- und Schutzgasschaltungen geeignet.

- Gesamtlänge: 57,2 mm
- Glaslänge: 20,3 mm max
- Durchmesser: 2,29 mm max
- Thermische EMK: ca. 5 μV/°C
- Lebensdauererwartung: über 10×10^6 Schaltvorgänge bei Vollast; fast unbeschränkt wenn in Schutzgasschaltungen
- Max. Nennlast: 28 V –, 10 mA

EE 86 769 *für weitere Einzelheiten*

Doppel-Diversity-Empfangsstelle

Racal Communications Ltd, Western Road, Bracknell, Berkshire

(Abbildung Seite 696)

Die Fertigung der Racal-Doppel-Diversity-Empfangsstelle RA.133 ist jetzt angelaufen. Das Gerät ist für Einsatz bei Rundfunkgesellschaften in aller Welt bestimmt, die Sprach- und Musikprogramme in den internationalen HF-Bändern empfangen und entweder über Ortssender wieder ausstrahlen oder über Drahtrelaissysteme weiterleiten. Eine Anzahl dieser Stationen wurden bereits an die British Broadcasting Corporation geliefert.

Die Empfangsstelle überstreicht einen Frequenzbereich von 1 ... 30 MHz oder 2 ... 24 MHz mit zusätzlicher HF-Vorselektion. Die Vorselektionskreise erhöhen die Selektivität, so dass ein um 5 Prozent verstimmtes Störsignal ungefähr 100 dB gedämpft wird.

Die Einrichtungen für Doppel-Diversity-Empfang können mit Zwei- oder Einseitenbandsignalen arbeiten. Man kann entweder das obere oder untere Seitenband wählen; im Zweiseitenbandbetrieb kann daher das Seitenband mit den geringsten Störungen eingestellt werden.

Die Empfangsstelle RA.133 ist mit zwei Empfängern RA.117C ausgerüstet, die Stabilität und Einstellgenauigkeit geben. Diese Empfänger RA.117C können entweder getrennt oder in Mutter-Tochter-Beziehung abgestimmt werden. In der letzteren Methode steuert das zweite Signal des regelbaren Schwebungsoszillators in einem Empfänger die 1 Hz-Einstellung von beiden Empfängern. Diese Arbeitsweise vereinfacht die Abstimmung, und die genaue Einstellung beider Empfänger auf die Frequenz ist damit gesichert.

In jeden Empfangsweg ist ein Abschwächer mit fünf wählbaren Einstellungen geschaltet, und es gibt zwischen 1,2 und 13 kHz fünf umschaltbare ZF - Durchlassbandbreiten. Diversity-Kombination der oberen, unteren und Zweiseitenbandsignale der beiden Empfangswege erfolgt in passiven Netzwerken; umfangreiche visuelle und akustische Überwachungseinrichtungen sind vorgesehen.

EE 86 770 *für weitere Einzelheiten*

Kabelbündelwerkzeug

Hellermann Electric Ltd, Gatwick Road, Crawley, Sussex

(Abbildung Seite 697)

Hellermann Electric Ltd hat unter dem Namen Tyton ein neues Werkzeug zum Bündeln von Kabeln eingeführt, das gegenüber anderen Systemen für permanentes Bündeln 50 Prozent Zeit und Kosten einsparen soll.

Das Tyton hat ein Magazin mit 20 Befestigungsstreifen, was 20 Bündelungsoperationen ohne Unterbrechung auszuführen erlaubt, und da jede Bündelung fester ist, braucht man weniger Kabelbindungen. Die grössere Festigkeit wird durch die Befestigungsweise erzielt, da der Nylonstreifen verformt und nicht durch Durchlochen geschwächt wird.

Der Nylonstreifen wird auf Rollen geliefert und jeweils nach Beendigung des Bündelns automatisch abgeschnitten. Dadurch wird Streifenabfall vermieden, und das Werkzeug ist jederzeit einsatzbereit. Weiterhin können auch Kabelbäume jeder Grösse jederzeit ohne Unterbrechung oder Verzögerung gebunden werden.

Ein Halstragband wird mitgeliefert, der Arbeiter hat also beide Hände frei, wenn er nicht gerade Kabel bindet.

EE 86 771 *für weitere Einzelheiten*

Testausrüstung für Fernsprecher

Telefon Fabrik Automatic A/S, 7 Amaliegade, Kopenhagen, Dänemark

(Abbildung Seite 697)

Diese Testausrüstung für Fernsprecher ist für das elektrische Prüfen von kompletten Fernsprechgeräten, Handapparaten, Sprechkapseln, Hörern und Klingeln ausgelegt. Der Test umfasst Messen der Empfangs-, Sende- und Rückhörwirksamkeit, Mikrofongeräusch, Impedanz, Rufen, sowie Gleichspannung und -strom.

Die Testausrüstung ist in erster Linie für Fertigungs- und Abnahmeprüfungen bestimmt, ist jedoch genau genug für Einsatz im Labor; die Arbeitsweise der Ausrüstung ist einfach genug, um Bedienung durch Personal ohne technische Kenntnisse nach kürzester Einweisung zu gestatten.

Die Ausrüstung besteht aus zwei Teilen. Der elektronische Teil EJ5-01 umfasst elektronische Schaltungen, Regelorgane und ein Messgerät. Der akustische Teil—der künstliche Kopf EJ5-02—umfasst ein künstliches Ohr, einen künstlichen Mund, sowie eine Universalprüfeinrichtung, in die praktisch jeder Handapparat eingelegt und richtig positioniert werden kann. Die zwei Teile sind durch Kabel miteinander verbunden.

Während des Testens wird das Fernsprechgerät über eine in den EJ5-01 eingebaute Gleichstromspeiseschaltung angeschlossen. Die Speiseschaltung besteht aus einer eingebauten Konstantstromversorgung für 24 V oder 48 V, einer Speisespule und einem Serienregelwiderstand, so dass man den normalerweise in der Leitung auftretenden Strom einstellen kann. Wenn erforderlich, kann das Fernsprechgerät über eine künstliche Übertragungsleitung wie z.B. den TFA—Typ EG4—gespeist werden.

Übertragungseigenschaften werden statt mit Sprache mit einem bewerteten Weissrauschen gemessen. Es wurde bewiesen, dass diese Methode für Ohrlautheittest den durch Besprechung erzielten Ergebnissen gleichwertig ist. Die Bewertung des Weissrauschens besteht aus einer mit der Frequenz um 6 dB je Oktave steigenden Abschwächung. Auf diese Weise entspricht das Frequenzspektrum des akustischen Rauschens annähernd dem Spektrum der durchschnittlichen menschlichen Sprache.

EE 86 772 *für weitere Einzelheiten*

Zeitgeber

System Engineering Services Ltd, Radley House, 35-39 South Ealing Road, London, W.5

(*Abbildung Seite 697*)

Tage, Stunden, Minuten und Sekunden anzeigende Zeitgeber können jetzt von Systems Engineering Services Ltd bezogen werden.

Modell EECO 869 ist in zwei Ausführungen lieferbar; Typ 869-6 zeigt Stunden Minuten und Sekunden an. Typ 689-9 gibt zusätzlich zu der vom Typ 869-6 angezeigten Information auch noch Tage an. Bis zu drei Anzeigen kann ein Zeitkodegenerator treiben.

Das Anzeigefenster ist mit einem polarisierten Lichtschirm ausgestattet, der die beleuchtete Nixie-Anzeige klar und scharf macht. Die Nixie-Röhren, von denen eine Lebensdauer von 200 000 Betriebsstunden zu erwarten ist, haben 34,3 mm Durchmesser.

Modell 869 ist entweder für Gestelleinbau oder in einer Spezialausführung für Montage an Wand, Tisch oder Decke lieferbar. Beide Modelle sind 89 mm hoch, 483 mm (19″) breit und einschliesslich Steckverbindung 156 mm tief.

Der Temperaturbereich der Einheit ist $-20°$ C ... $+55°$ C.

Andere EECO-Zeitanzeigegeräte—das Modell für Paralleleingabe EECO 865 und die Modelle für Serieneingabe EECO 866 und 867—sind hauptsächlich für Fernanzeige der von einem Zeitkodegenerator erzeugten Zeit gedacht.

EE 86 773 *für weitere Einzelheiten*

Abisolierwerkzeug

Light Soldering Developments Ltd, 28 Sydenham Road, Croydon, Surrey

(*Abbildung Seite 697*)

Das "Adamin"-Modell 2B6 ist ein neues thermisches Miniatur-Abisolierwerkzeug für Einsatz mit Isolierstoffen niedrigen Schmelzpunktes wie PVC, Polyäthylen, Polyamid usw.

Der mit einer Hand betriebene Adamin-Abisolierer arbeitet wie eine einfache Pinzette und kann Umhüllungen bis zu 4 mm Durchmesser ohne Beschädigungsgefahr für den Leiter abisolieren.

Modell 2B6 ist für 12 oder 24 V lieferbar; die Energieaufnahme von ca. 14 W erzeugt gerade genug Wärme, um die Umhüllung zu schmelzen. Höhere Temperaturen würden übermässiges Schmelzen hervorrufen, und die flüssige Umhüllung würde über den abisolierten Teil des Drahtes geschmiert werden.

Je nach Bemessung des Abisolierers kann jede Gleich- oder Wechselstromquelle für 12 oder 24 V, die 1 A bzw. 0,5 A abgibt, für die Speisung benutzt werden. Andererseits kann man die Abisolierer über "Litesold"-Trenntransformatoren Typ LT12 (12 V, 25 W), LT24 (24 V, 25 W) oder MT24 (24 V, 55 W) direkt ans Netz anschliessen, was den zusätzlichen Vorteil hat, dass die drei Abgriffe dieser Umspanner eine gewisse Temperaturregelung erlauben.

Der abzuisolierende Draht muss für ungefähr eine Sekunde fest zwischen den Schneidbacken gehalten werden, um Durchschmelzen der Umhüllung zu gestatten. Durch einen geraden Abzug wird dann der Abisoliervorgang fertiggestellt.

Das bekannte Modell 2B24 ist ein ähnliches Werkzeug, arbeitet jedoch mit höheren Temperaturen und kann für Polytetrafluoräthylenisolierung eingesetzt werden. Diese Isolierung wird nicht durchgeschmolzen; die Backenwärme erweicht sie nur so weit, dass die Backen sie durchschneiden können.

EE 86 774 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Eine automatische Koordinatenmessausrüstung von R. L. Farrel und G. E. Moore

Zusammenfassung des Beitrages auf Seite 664-651

Genaue Information über die Flugbahnen von Raketen usw. erhält man durch Messen der Positionen ihrer Bilder in bezug auf diejenigen von Sternen auf fotografischen Platten. Mit Hilfe eines Zeiss-Komparators, einem optischen Koordinatenmessgerät, können diese Messungen auf Platten mit weniger als 1 Mikron Unsicherheit vorgenommen werden.

Die beschriebene Ausrüstung registriert die Bewegungen des Plattenträgers im Komparator und gibt eine visuelle Dezimalanzeige sowie eine Lochkartenausgabe.

Die Genauigkeit von 1 Mikron beim Messen der Koordinaten wird durch Zählen der Moiré-Streifenbilder erreicht. Die Streifenbilder werden mittels Sperrschichtfotozellen gelesen und verstärkt. Spezialschaltungen bestimmen die Bewegungsrichtung. Die für die Summierung der Streifenzählung verwendeten Dezimalzähler können mit grosser Geschwindigkeit zählen und schnell umgekehrt werden und enthalten einige neuartige Entwurfsprinzipien. Ein auf dem Konzept der kritischen Verlangsamung beruhendes Spezialalarmsystem entdeckt auf Ausfall optischer oder elektronischer Baugruppen zurückzuführende Fehler.

Der Entwurf von Tunneldioden-Schaltkreisen von R. V. L'Archeveque und B. McA. Sayers

Zusammenfassung des Beitrages auf Seite 652-655

Tunneldiodenschaltungen, die monostabil oder bistabil mit mittlerer Geschwindigkeit (20 MHz) schalten, werden rekapituliert. Die Entwurfsgleichungen für diese Schaltungen werden von den Steuerberechnungen für einen Sägezahnoszillator verallgemeinert. Weiterhin wird ein stabiler Vorstufendiskriminator beschrieben.

Hochspannungserzeugung mit piezoelektrischen PZT-Keramikstoffen von C. Pearcy

Zusammenfassung des Beitrages auf Seite 656-659

Jüngste Entwicklungen von piezoelektrischen Werkstoffen haben zur Erzeugung von Hochspannungsimpulsquellen geführt, die auf der direkten Umsetzung mechanischer Energie in elektrische beruhen. Dieses Prinzip wurde in der Herstellung von Funkengeneratoren mit PZT-(Bleizirkonattitanat) Keramikzylindern ausgenutzt. Ausrüstungen dieser Art wurden erfolgreich für das Zünden verschiedener Gasgeräte, Aufschlagzünder für Raketen und für Hochspannungsversorgungen mit niedriger Stromentnahme—wie z.B. für tragbare Röntgengeräte—eingeführt.

Zweck dieses Beitrages ist das Umreissen der verschiedenen zur Herstellung eines solchen Systems erforderlichen Konstruktionsmethoden. Entwurfsgleichungen werden aufgeführt und zusammen mit den einschlägigen mechanischen Eigenschaften der PZT-Keramikstoffe erläutert.

Ein Analogrechner für Prognose und Lagerkontrolle von R. Calvert und J. Mildwater

Zusammenfassung des Beitrages auf Seite 660-666

Aus den jüngsten Fortschritten in der Operationsforschung haben sich bessere Methoden für die kurzfristige Prognose ergeben. Der beschriebene Rechner wendet die Verhältnistransformatorentechnik zur Auflösung einer Reihe von Gleichungen an, die statistische Information für die Vorhersage und Lagerkontrolle geben.

Einer allgemeinen Beschreibung der Schaltungen für Addition, Subtraktion, Multiplikation und Division folgt eine eingehendere Erläuterung der Anwendung des Systems auf ein bestimmtes Gleichungssystem.

Die Mäanderleitungsantenne von T. S. M. Maclean

Zusammenfassung des Beitrages auf Seite 667-669

Eine neuartige Längsrichtantenne mit fortschreitender Welle in Form einer Mäanderleitung wird beschrieben. Sie besteht aus einem nur mit rechtwinkligen Biegungen in Elemente gefalteten Leiter, wobei jedes Element etwas länger als eine halbe Wellenlänge ist. Die Längsrichtstrahlung besteht über eine Bandbreite von etwa 12% mit Nebenkeulenpegeln von besser als 6 dB.

Ein mit zwei Transistoren bestücktes Reizgerät für physiologische Zwecke von J. Opalinski

Zusammenfassung des Beitrages auf Seite 670-673

Der Beitrag beschreibt ein kompaktes, mit zwei Transistoren bestücktes Reizgerät mit elektromagnetischer Kupplung durch eine Zenerdiode für Anwendung im Unterricht. Einbebriffen ist ein mit einer getrennten Doppelbasisdiode bestückter Zeitgeber.

Ein überbrücktes T-Netzwerk für Magnettesten mit Wechselstrom von S. N. Bhattacharyya und J. K. Choudhury

Zusammenfassung des Beitrages auf Seite 674-677

Ein überbrücktes T-Netzwerk für Magnettesten mit Wechselstrom hat gegenüber herkömmlichen Messbrücken gewisse eindeutige Vorteile, hauptsächlich auf Grund seiner verhältnismässigen Unempfindlichkeit gegen die Einwirkung von Streukapazitätskupplungen, die besonders bei Ton- und höheren Frequenzen durch ihre Wirksamkeit auffallen. In der vorliegenden Untersuchung wurde eine Anordnung des überbrückten T-Netzwerkes benutzt, die bei Abgleich im wesentlichen einen kompensierten Serienresonanzkreis darstellt. In dem Beitrag gegebene Versuchsergebnisse illustrieren die Anwendungsmöglichkeit des Netzwerkes für Magnettesten unter verschiedenen Erregungszuständen des Prüflings. Messungen wurden bei Frequenzen zwischen 50 Hz und 40 kHz durchgeführt.

Direkte Verwandlung einer mehrmaschigen Verknüpfung in einen Vierpol von D. W. W. Rogers

Zusammenfassung des Beitrages auf Seite 678-680

Eine Methode zur Erlangung der Vierpolgleichung eines mehrmaschigen Netzwerkes wird beschrieben. Die Methode stützt sich auf eine Technik, nach der die Impedanz- (oder Admittanz-) Parameter jedes Netzwerkes direkt in Ausdrücken reflektierter Impedanz niedergeschrieben werden können. Eine ähnliche Behandlung wurde von Robichaud[1] mit der Signalflussanalyse durchgeführt.

Ein phasenempfindlicher Diodengleichrichter von B. S. Leela und R. Shankar

Zusammenfassung des Beitrages auf Seite 681-683

Eine phasenempfindliche Diodengleichrichterschaltung wird in bezug auf Nichtlinearität eingehend analysiert. Lösung der betreffenden Gleichungen mit Hilfe eines Digitalrechners führt zu neuen Kriterien für die Wahl von Bauelementwerten. Insbesondere wird gezeigt, dass durch verhältnismässig geringe Beschränkungen des Signalpegels wesentlich niedrigere Nichtlinearitätsziffern erreicht werden können. Ergebnisse werden in Form einer schnell übersichtlichen Tabelle dargestellt.