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Commentary

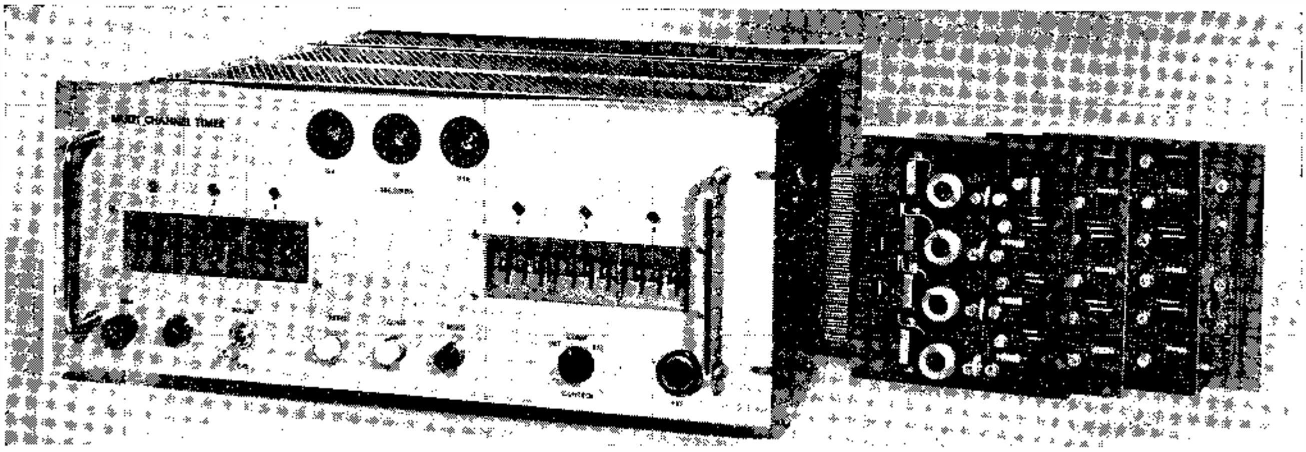
'APPROACHING Automation' is the title of a new film which has been issued by the Ministry of Technology. It had its initial showing at an exposition, with the same title, which was held at the City University (London) from 15 to 19 September. This exposition was one of a series which it is planned to hold in various parts of the country; at Bristol from 18 to 20 October, at Manchester from 24 to 29 October, and at Cardiff, Edinburgh and Paisley at dates to be announced. These expositions, which are aimed at the medium and smaller sized firms, are being organized by the Regional Officers of the Ministry of Technology in association with appropriate University departments and take the form of film shows, short talks by experts in various phases of control and automation, together with discussion sessions and small exhibitions of automatic control devices provided by commercial firms, research associations and Government departments. The object of these events is to make the smaller firms increasingly aware of the increases in productivity and greater control of quality, which can be achieved by the wider use of existing, low-cost control devices and to provide contacts through which firms can investigate the possibilities which are open to them.

This step-by-step approach to automatic control which the Ministry is promoting for the smaller firms complements the Ministry's activities in the field of more advanced automation; it is already actively supporting the installation of a number of advanced systems through the National Research Development Corporation which is the source of public support for the direct application of known automation techniques to major schemes in industry.

The word 'automation' tends to conjure up, even to those who should know better, a picture of vast automatically operated process plants, on-line computers and push-button control of complicated operations. This is not necessarily a wrong picture, but rather it is the ultimate one, and, in fact, automation may amount to no more than the addition of an hydraulic actuator to a previously hand-operated fly-press. Nor is automation necessarily a costly affair: simple schemes may be achieved with very modest capital outlay and still show considerable gains in productivity, better and more consistent quality, and in greater operator safety. Indeed, automation may best be regarded not so much as a system which must be brought into full operation at a given moment, but rather as an evolutionary process that can be achieved in stages. This applies particularly to the smaller firm which is often best able to introduce automatic control in successive steps, learning as it does so and progressively modifying its procedures to take full advantage of the new techniques. Thus, if for a start only one or two individual operations are automated, the result will often make clear the possibilities and advantages of further automatic control. The first steps need be neither drastic nor costly and, for example, could be restricted to the feeding of a single machine and perhaps be extended progressively to include the machining of a complete workpiece and its ejection. The next step could be to automate two or more machines in this way and then link them with automatic transfer devices to pass work from one to the other. The automation of a single operation may in itself solve a particular production problem but, if a succession of linked processes are so controlled, the overall benefits will generally far exceed the sum of those to be expected from each individual step.

As has been stated many times, in this journal and elsewhere, there is nothing new about automation but it is surprising how many firms, especially the smaller ones, fail to appreciate how much can be achieved by the use of simple automatic control schemes, and the value of these expositions in bringing such matters to light could be enormous. The size of the dividend to be gained from increasing productivity among these firms can be appreciated from the fact that nearly three-quarters of the firms in Great Britain employ fifty workers or fewer.

One tends to couple automation and electronics together but, of course, a great deal of automatic control can be carried out without any aid at all from electronics and, it must be admitted, there are occasions where automatic control can be achieved more simply and efficiently by non-electronic means. On the other hand, there are many instances where electronic devices can provide a simple solution to a complex problem and many, many more where electronics is the only medium which can provide a solution at all. In fact, automation may be said to affect the electronics industry in two ways; both internally and externally, so to speak. From an internal point of view, there are many processes in component and equipment manufacture which can successfully be automated and, of course, a great deal has already been done in this direction. Nevertheless, it is surprising to find how many firms there are in such a modern and forward-looking industry who do not seem fully to realize the great advantages which can accrue from automatic methods. Externally, the electronic industry has a great deal to offer and a great deal to accomplish. Electronics is the most versatile and flexible aid to automation, and the main job now, as it has been in the past, is to make the needs of other industries known and understandable to the electronic engineer and, conversely, to make the possibilities of electronics known to the remainder of industry.



A Versatile Multi-channel Timer

By A. I. Nizan*, B.Sc., M.Sc.

An electronic timer is described which possesses six independent timing channels. The range of each channel is 0·1 to 99·8sec. Timing within this range may be selected in intervals of 0·1sec. Accuracy is practically equal to that of the mains frequency, but may be raised if an accurate external clock oscillator is used. This oscillator may at the same time affect the time scale and thereby extend the range of the instrument. In addition, any channel out of the six can function as a triple coincidence gate. The timer may be operated in a single or cyclic mode. This instrument has been found very useful in analogue and hybrid computation.

(Voir page 694 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 701)

A FACILITY frequently needed for large analogue computers is a timer with several independent outputs, each of which may be preset within the range of 0·1 to 100sec. The various needs encountered in analogue computer work are best met if these outputs take the form of fast-acting switches. Accuracies of 0·1sec or better are required. It is also necessary that the timer could be slaved to the computer control. Other desirable features are visual display, performance of some elementary logic functions, etc.

Conventional mechanical timers are generally motor-driven devices with a shaft-rotated cam actuating several microswitches in turn. Accuracies achieved are of the order of 0·1 to 0·5sec, yet it is quite difficult to set the exact time. Moreover, switches suffer from the bounce and wear typical for most mechanical moving parts. Even more sophisticated types of mechanical clocks[1] seldom exceed accuracies of 0·5sec when time ranges are as large as those designated above.

Electronic timers on the other hand—may be sub-divided into two categories: Those in which time is measured 'analogically', and others based on clock-pulse counting.

'Analogue' timers use some kind or other of linear charging (capacitors, magnetic cores, etc.)[2,3] and are generally equipped with output facilities—e.g. relays that enable them to control other instruments. Many industrial timers are of this type. Linearity is the main problem[2], and elaborate means of compensation are needed to attain acceptable accuracies over wide ranges of time. Visual display of elapsed time is generally unavailable.

Timers utilizing counting techniques can be made to measure time very accurately. In fact, these are counters coupled to electronic clock pulse generators. A large variety of these is commercially available. However, they lack those output control facilities which for the present purpose are indispensible, as well as the provision of a number of parallel independent output channels.

The electronic timer to be described (shown above) combines the advantages of the counter-type timers with the output capabilities of the analogue type: The accuracy is actually equal to that of the mains frequency—the internal clock being mains-synchronized. Should an increase in accuracy be desired, a more accurate external clock may be added and the task of the internal clock transferred to it. There are six independent timing channels with a time range of 0·1sec to 99·8sec each. Operating time may be selected over the whole range in steps of 0·1sec. Every channel output consists of three contacts of a quick-acting, mercury-wetted-contact relay. The timer may be controlled by the computer (automatic), manually, or externally.

The capabilities of the unit are extended by the provision that any channel may be transferred manually to a triple-coincidence circuit for logical operation.

Principles of Operation

The block diagram of the unit is given in Fig. 1. First its function as a six-channel timer will be described, and next it will be shown how the timing channels of the unit can be utilized in performing coincidence functions.

When the unit acts as a timer, a gated train of pulses is fed into a chain of four 10:1 dividers. Gating of the shaped clock pulses is performed by a signal supplied from the control unit. The absence of synchronization between the gating signal and the clock pulses gives rise

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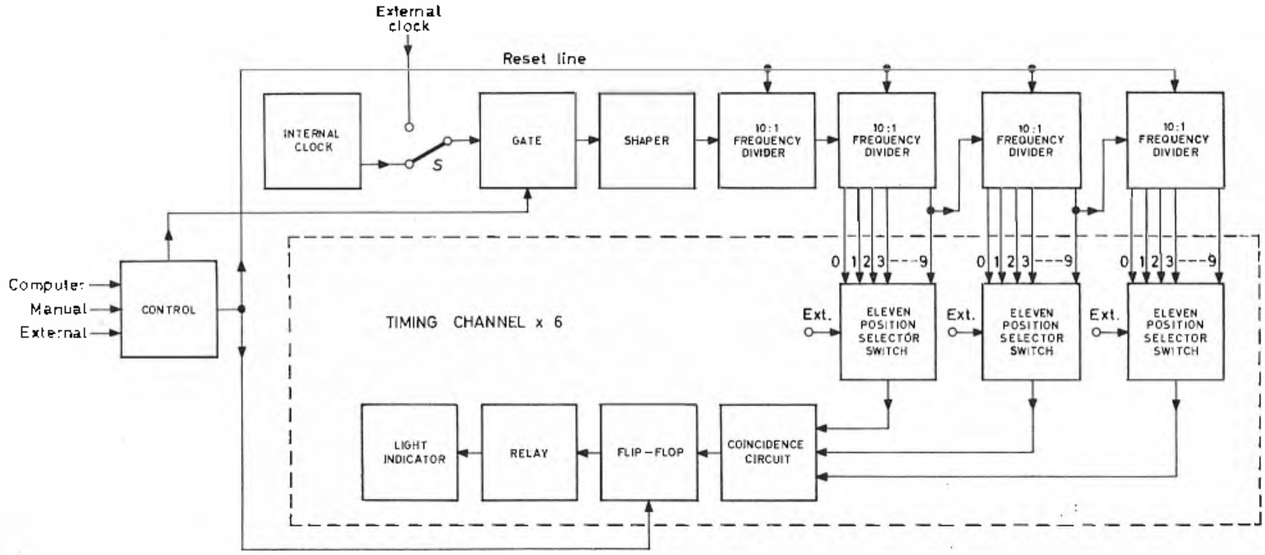


Fig. 1. Arrangement of multi-channel timer

to a maximum time jitter of one clock pulse. This time jitter is reduced to a permissible value by raising the clock frequency one decade higher than called for by the desired 0·1sec resolution. The clock frequency will thus be 100p/s while the desired 10p/s repetition rate is attained at the output of the first divider in the chain. The last three dividers perform the function of time-interval multiplication. Each of these possesses 10 outputs arranged in decimal order. A timing range of 99·9sec is thus attained with a resolution of 0·1sec.

The blocks described so far are common to all channels. The timing channels are each composed of three 11-position selector switches, a triple-coincidence circuit, flip-flop, relay and light indicator. Every selector connects a chosen output from one of the last three dividers with one input of the triple-coincidence circuit. When all three inputs carry a signal simultaneously, the coincidence circuit triggers the flip-flop, as a result of which the channel relay becomes energized. The make (or break) action of the relay constitutes the timing operation of the channel. Indication of its operation is given by a panel-mounted light indicator.

The time range may be altered by substituting an external clock in place of the internal one, which will feed the unit through the external clock input. Such a clock may also serve to raise the accuracy of the unit, in case the accuracy of the internal clock—which is mains-frequency synchronized—is insufficient.

To put the unit into action, a 'Compute' command is given, upon which the gate opens and pulses enter the frequency divider stages—i.e. timing action of the unit is started. This action may be interrupted by closing of the gate. The corresponding command is termed 'Hold'. From this state, the unit may be commanded to proceed with the pulse-counting timing operation — again by a 'Compute' command—or else it may be reset to the position in which it will be ready to start anew; the 'Reset' command resets the frequency dividers and channel flip-flops (the latter causing the de-energization of the channel relays) and causes the gate to close. The 'Reset' command may, of course, follow the 'Compute' command.

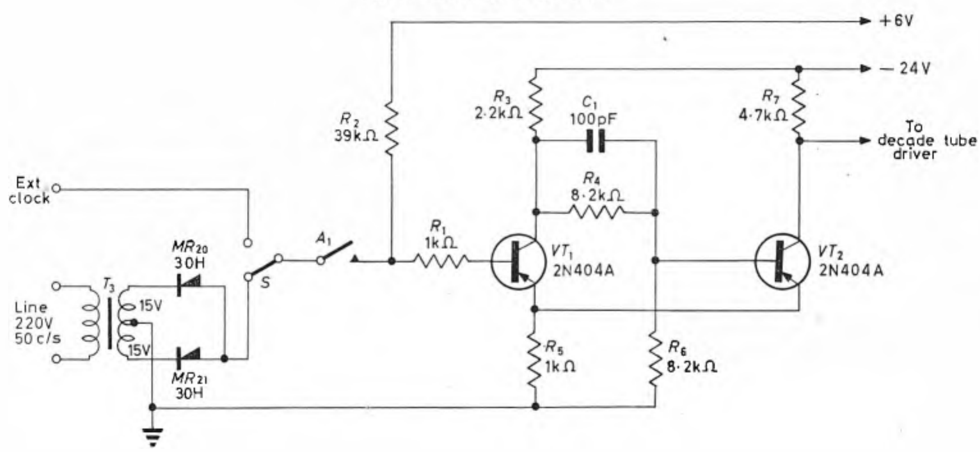
These three commands—which for the sake of convenience are named after the corresponding analogue-computer commands—make up the unit's control modes. As already mentioned they may be effected manually, automatically (by the computer), or externally.

Each of the three inputs of a coincidence circuit may be fed by an external signal (which does not come from the internal dividers). This is accomplished by setting the corresponding selector to position '10', whereby the coincidence circuit input is connected to the external signal source. Thus any timing channel may be used to perform coincidence functions—the output taking the form of relay-contact action.

Circuit Description

The schematic circuit diagram of the internal clock, gate and shaper is given in Fig. 2. The clock is composed of a line transformer T_3 and a full-wave rectifier (MR_{20}; MR_{21}) acting as a frequency doubler. Contacts A_1 (of relay A) form the gate. When they are closed, the

Fig. 2. Clock pulse generator



rectified output waveform is fed into the Schmitt circuit (VT_1; VT_2) which constitutes the shaper. When driven from an external clock, the connexion to gate and shaper is made by switch *S*.

A frequency divider composed of a gas-filled counting tube and its driver circuit is shown in Fig. 3. The advance trigger pulse which is supplied to the decade tube through transformer T_1 is generated by the following current-switching action: Transistor VT_3—acting as a parallel through the primary of T_1 during the 'on' state of VT_3. Full compensation is attained when the ratio of this current and the pulse current equals the duty ratio of the latter.

The circuit described possesses the following advantages:

(a) An advance trigger pulse is attained with a non-regenerative circuit. The decade tube counting reliability is thus increased by the reduction of its sensitivity to noise pulses.

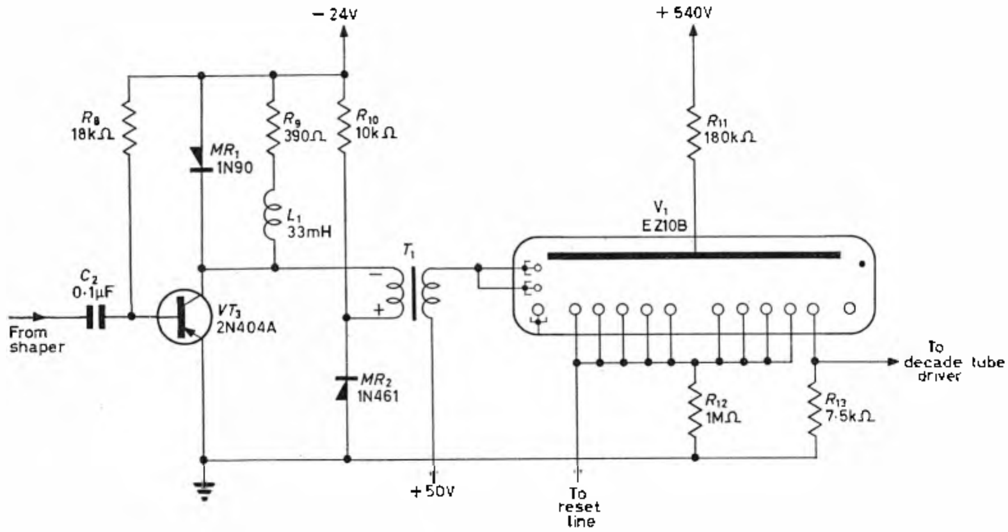


Fig. 3. First 10:1 divider

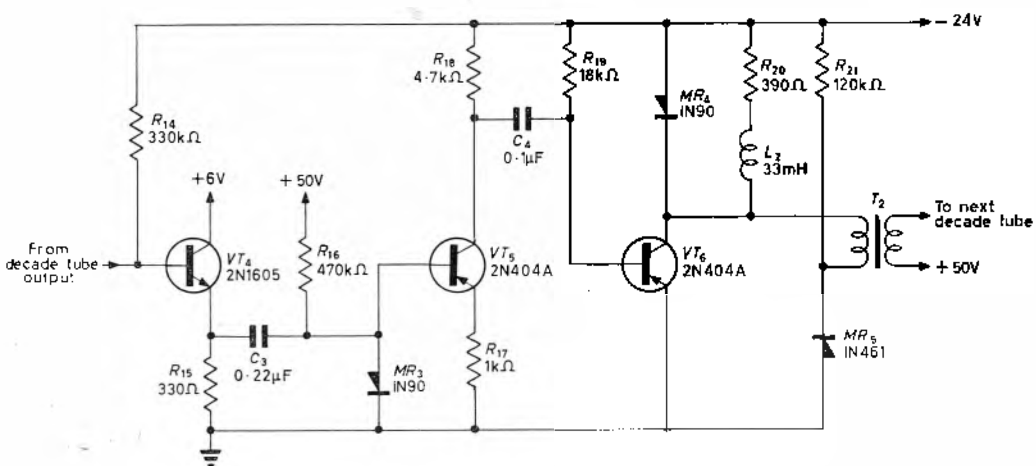


Fig. 4. Decade tube driver

switch—is normally saturated and passes a current through inductance L_1 and resistor R_9. A positive output pulse from the shaper—differentiated by C_2—cuts off VT_3, routing the current through diode MR_1 and the primary of T_1. Consequently, a fast-rising negative voltage pulse—the amplitude of which equals the supply voltage—develops on the primary of T_1, resulting in a 120V secondary voltage pulse of 20μsec width, which suffices to drive the decade tube[4].

The unipolarity of the input pulses to T_1 causes a shift of the quiescent point on the *BH* curve of the transformer core. The effect of such a shift is a reduction of the transformer's output pulse width which depends on the primary inductance. Resistor R_{10} and diode MR_2 compensate this effect by driving a small opposing current

(b) Despite the small dimensions of the transformer, a fairly wide trigger pulse is achieved through an optimal exploitation of the transformer core.

When driven from another counting tube, a small modification in the driver circuit of the following counting tube has to be introduced, namely a buffer and invertor amplifier must be added (Fig. 4). The emitter-follower VT_4 reduces the loading on the 9th dynode of the counting tube, while VT_5 amplifies the signal pulse which thereafter attains a form similar to the output pulse of the Schmitt circuit. The advance pulse is taken from the (negative) trailing edge of the counting tube output waveform.

The schematic circuit of a timing channel is given in Fig. 5. Here again the decade tube outputs—having in

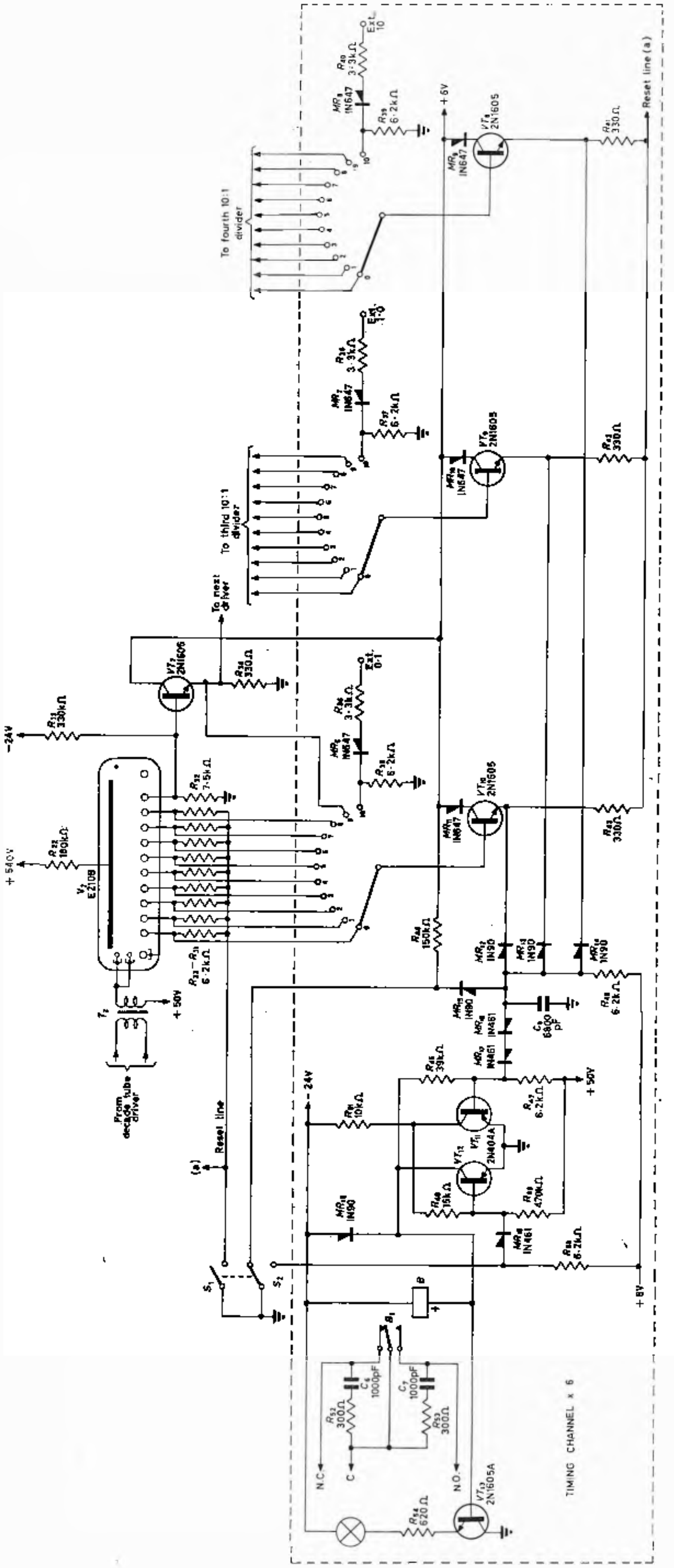


Fig. 5. The timing channel

this case a fan-out of six—are isolated from the triple-coincidence circuit (MR_{12} to MR_{14}) by emitter-followers VT_8 to VT_{10}. Position 10 on each selector switch allows the inputs to the coincidence circuit to be fed by external signals.

If all three inputs of a coincidence circuit carry signals simultaneously, it will produce an output which will trip the flip-flop—VT_{11}: VT_{12}—thus energizing the channel relay.

The latter is of the mercury-wetted-contact type and its contact making (or breaking) constitutes the timing action of the channel.

Diodes MR_{16} and MR_{17} introduce a voltage shift which removes singles and doubles. Capacitor C_5 produces a small delay to compensate for the propagation time of the clock pulses through all divider stages. The resultant time delay in channel operation is practically negligible.

Timing accuracy of the unit is equal to that of the mains frequency ±10msec. When a timing channel is triggered from an external source (by a +6V 20μsec pulse) the channel relay will operate after a maximum delay of 1·1msec.

To reset the counting tubes, conduction in each of them has to be transferred to dynode 9. All other dynodes are then disconnected from ground through opening of contacts S_1. For the same reason, the emitter-followers VT_8 to VT_{10} are also disconnected from ground. The eventual path through the base-collector diodes of these transistors to supply and ground is prevented by silicon diodes MR_9 to MR_{11}.

'Reset' is effected by de-energizing relays A and S. The former has one pair of contacts which constitute the gate. They open on 'Reset' and 'Hold', and close on 'Compute'. Relay S has one normally-open contact pair S_1 which, as described, disconnects the reset line from ground and another changeover contact, S_2; the normally-closed contact of S_2 shorts MR_{15} to ground while the potentials on other coincidence input lines are undefined during the reset period. During this period the channel flip-flop is reset through R_{50} and MR_{18}. In the 'Compute' and 'Hold' states, relay S is energized and consequently MR_{15} and MR_{11} are back-biased and inoperative.

Indication of channel operation is given by a small panel-mounted pilot lamp controlled by VT_{13}.

Power supplies are conventional. Their regulation—which is somewhat better than 2 per cent—is not critical.

Conclusions

A timer has been described which is

suitable for a large variety of applications. It possesses six independent timing channels in which quick-acting mercury-wetted-contact relays are utilized. These contacts are free from bounce and wear and can, if necessary, handle appreciable power. The switching action can be speeded up by replacing the relay by a suitable electronic switch (triggered by the channel flip-flop). Moreover, an output pulse may be obtained to trigger successive circuits. The unit may be turned into an events counter by another minor change—i.e. by attaching the fourth decade tube to the front panel along with the other three, which at 'Compute' will give a full display of the counted pulses fed through the external clock input. Cyclic operation of the timing channels is another optional feature of the instrument: a switch whose contacts bridge contacts A_1 (which form the gate) will put the timer into continuous operation. If one of the channels is used for 'Reset' (through the external control), cyclic operation of the unit will result. This mode of operation can be utilized in low frequency synthesizing or any other periodic function. Only slight design alternations are needed to increase the number of timing channels.

Acknowledgments

This work was carried out under the auspices of the Scientific Department, Israel Ministry of Defence. The author is much indebted to Mr. S. Rozen for helpful discussions and suggestions as well as for his guiding comments on this article. Thanks are also due to Mr. G. Bitansky who carried out the practical work and to whose suggestions and initiative the many versatile features of the instrument are due.

Finally, the author wishes to express sincere appreciation to Mr. E. Gross for help in preparing this article.

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All-Weather Aircraft Landing

A new British designed and manufactured instrument landing system has been introduced by Standard Telephones and Cables Ltd that will enable suitably equipped aircraft to land safely in conditions of zero visibility.

Already, orders worth £425 000 have been placed for the new equipment and associated aerials, known as STAN. 37/38/39, by the Civil Aviation Department of the Board of Trade in the UK (£225 000), and the Department of Civil Aviation of Australia (£200 000). Equipments will be installed in the first place at Birmingham and Stansted and shortly afterwards at London (Heathrow). In Australia the equipment will be used at Sydney Airport and Melbourne's new Tullamarine Airport, which is now under construction.

To achieve the extremely high degree of reliability necessary for fully automatic landing, valves have been replaced by solid-state devices, the simplest possible design principles have been adopted, vital items of equipment have been duplicated, the need for routine servicing has been reduced to a minimum and highly sophisticated monitoring has been embodied to detect the slightest deviation from normal operation.

In accordance with existing ICAO classification all-weather landing is being introduced in three major stages or categories. Category I covers approach minima of 200ft cloudbase and 2 600ft visibility. Category II reduces these minima to 100ft and 1 300ft and Category III, which has not yet been exactly defined, will provide for landing in weather conditions inferior to Category II and will ultimately include landing without external visual reference.

STAN. 37 (localizer), STAN. 38 (glide path), and STAN. 39 (marker) fully comply with the recommendations of ICAO's All-Weather Operations Panel and meet the requirements of the Civil Aviation Department of the Board of Trade for all-weather landing. System flexibility provides for a Category II facility to be subsequently upgraded to Category III if required.

A number of significant features are included in this new equipment.

The individual localizer and glide path ground equipments are designed to contribute a landing failure risk of less than 1 in 10^9 in order to establish an overall guidance system failure risk of 1 in 10^7.

The equipment uses solid-state techniques throughout, and only high integrity components are used.

Protection against course bends, caused by reflections from buildings and high ground in the vicinity of the airport, is obtained by a unique technique known as 'quadrature clearance' the key to meeting Category III requirements without increase in system complexity. Separate 'course' and 'clearance' signals are radiated on the same carrier frequency and from the same aerial, but the 'course' and 'clearance' tones are in phase quadrature, so that unwanted reflections due to the clearance lobes do not interfere with the course signals.

A single transmitter and modulator produces the transmissions, with a significant gain in equipment simplicity and thus system reliability.

The quadrature clearance technique (patented by STC) provides even greater rejection of unwanted reflections than the two-frequency course/clearance method of earlier designs. If on a given site a beam-bend of 0·4 per cent d.d.m. occurs with the two-frequency system, quadrature clearance would reduce it to 0·14 per cent d.d.m. (difference in depth of modulation).

The normal type of frequent routine maintenance has been virtually "designed out" of the equipment itself, thus removing the risk of introducing other faults due to human error, and reducing maintenance costs. Such maintenance has been replaced by regular inspection and test of the monitor alarm system.

Constant impedance mechanical modulators are used (an STC patent) with a tone-frequency stability better than ±0·1 per cent.

Localizer, glide path and marker transmitters are duplicated and have a predicted m.t.b.f. of 10 000 hours minimum.

Localizer and glide path transmissions are monitored with duplicated field probes near to the aerials, and these are cross-connected to internal equipment monitors. Automatic changeover to the standby transmitter is initiated on a "two-out-of-three" monitor-agreement basis. Each standby transmitter runs continuously and is also monitored internally. The changeover (from operating to standby) takes place in less than half a second.

Solid-state design has enabled localizer and glide path equipments to be operated from 48V/24V high grade Planté batteries in conjunction with constant voltage chargers.

In the event of mains failure the equipment will operate from the batteries for two hours without loss of performance.

The low-mass low impact strength aerial is frangible and in the event of any section being struck by an aircraft, that section would disintegrate.

High Q stripline matrices are used for the r.f. structures that feed the aerials, i.e. power hybrids, phasers and matching units. Hence, manufacture to repeatable and extremely close tolerance is possible.

The dual marker beacon transmitters with monitor and facilities for remote control are housed in a pole-mounted weatherproof container, no building or major site works being required. The power requirements are extremely low and can be supplied over the remote control telephone cable, thus eliminating the need for expensive power distribution to remote sites. In the event of mains or cable failure a local battery supply ensures continuous operation of the equipment.

Impedance Convertors Using Wide-Band Feedback Amplifiers

By F. Butler, O.B.E., B.Sc., M.I.E.E., M.I.E.R.E.

Discussion and extension of recent work, briefly reported by R. L. Ford and F. E. J. Girling, concerning the use of linear amplifiers associated with passive feedback elements to simulate low-loss inductances. Some related circuits for capacitance simulation are also described and a first-order treatment of the theory of both types is given.

(Voir page 694 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 701)

In a brief communication[1], R. L. Ford and F. E. J. Girling described an interesting circuit using resistance-capacitance networks in conjunction with linear amplifiers to simulate a low-loss inductance. They quoted, without proof because of space limitations, expressions for the value of the simulated inductance. While working on a first-order treatment of the theory it occurred to the author that there is at least one possible rearrangement of their circuit which also gives inductance simulation and several which give equivalent results for capacitance.

The theory of all these circuits will be given, followed by some practical information about suitable amplifiers, a topic which was not discussed by the authors quoted.

Inductance Simulation

The procedure followed by Ford and Girling is to take a wide-band phase-reversing amplifier and, by the use of associated *RC* elements, make it simulate a lossy inductance. A second amplifier is then used to simulate a negative resistance with a value sufficient to cancel the shunt loss-resistance of the inductor. Over-compensation of the loss is possible, in which case it is possible to generate continuous oscillations by connecting a suitable capacitor in parallel with the inductor. Their circuits are shown schematically in Figs. 1, 2 and 3. The simple theory of Fig. 1 rests on the fact that an amplifier with an intrinsic gain, (without feedback), approaching infinity, requires only an infinitesimal input to produce a finite output. When feedback components are added, (in this case R and C), the input current is simply $i_1 = E/R$. Assuming that the amplifier has an infinite input impedance, the current in the capacitor C is the same as that in the resistance and is very nearly the same whatever the input impedance may be, provided only that the amplifier has sufficient gain. The gain of the amplifier, with feedback, is then simply the ratio of the feedback impedances $1/j\omega C$ and R, i.e. $m = 1/j\omega CR$.

If Z_i is the input impedance presented to the generator E, the circuit equations are:

$$i_1 = E/R; \quad i_2 = (1 + m)\,E/R_1;$$

$$\frac{i_1 + i_2}{E} = 1/Z_i = (1/R) + (1/R_1) + (m/R_1)$$

But $m = 1/j\omega CR$ so that:

$$1/Z_i = (1/R) + (1/R_1) + \frac{1}{j\omega CRR_1} \quad \ldots\ldots\ldots (1)$$

Thus Z_i corresponds to the parallel combination of R, R_1 and $j\omega CRR_1$. The last term represents an inductance $L = CRR_1$, so that Z_i is an inductance L shunted by a resistance $r = RR_1/(R + R_1)$, as shown in Fig. 1. This loss-resistance is objectionable, but it can be cancelled by negative resistance as produced by the circuit of Fig. 2. In this, the voltage across the resistance R_1 is $E - 2E = -E$. The current is $i = -E/R_1$, so that the resistance actually presented to the signal source E is $-R_1$.

Turning now to Fig. 3 and applying similar reasoning it will be found that:

$$i_1 = -(E/R_1) \;;\; i_2 = (1 + 2m)\,E/R_1 \;;\; m = \frac{1}{j\omega CR}\,;$$

$$\frac{i_1 + i_2}{E} = 1/Z_i = -(1/R_1) + (1/R_1) + (2m/R_1).$$

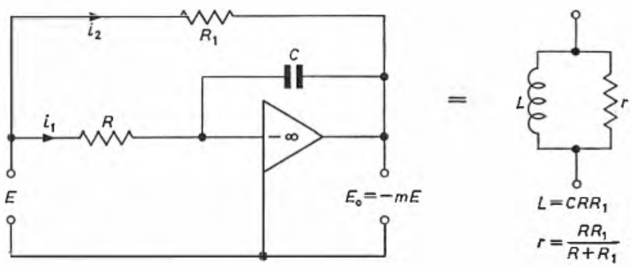


Fig. 1. Low-Q inductance simulator

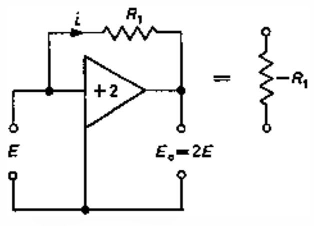
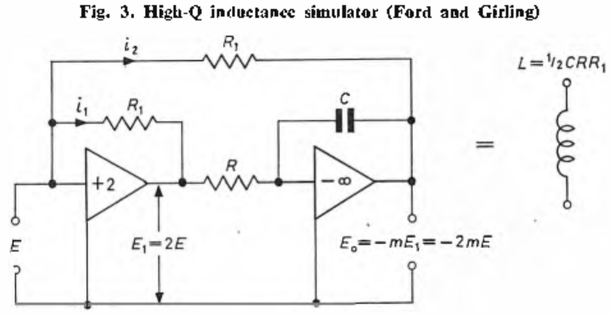


Fig. 2. Negative resistance circuit

Fig. 3. High-Q inductance simulator (Ford and Girling)



Hence:

$$Z_1 = R_1/2m = \tfrac{1}{2}\,j\omega CRR_1 \quad \ldots\ldots\ldots\ldots \quad (2)$$

This corresponds to a pure inductance $L = \frac{1}{2}CRR_1$. The loss resistance has been cancelled. Note also that the value of the simulated inductance has been halved as compared with the figure in equation (1). The circuit of Fig. 3 makes use of tandem-connected amplifiers. Another possible arrangement is shown in Fig. 4 in which the two amplifiers are in parallel.

In this case:

$$i = -(E/R_2)\,;\quad i_1 = E/R\,;$$

$$i_2 = (1+m)\,E/R_1\,;\quad m = \frac{1}{j\omega CR}$$

Hence:

$$1/Z_1 = \frac{i + i_1 + i_2}{E} = -(1/R_2)+(1/R)+(1/R_1) + \frac{1}{j\omega CRR_1}$$

In this case Z_1 becomes a pure reactance if $1/R_2 = (1/R) + (1/R_1)$, i.e. if the negative resistance R_2 is such as to cancel the parallel combination of R and R_1. The simulated inductance is then $L = CRR_1$, (twice the value previously quoted for the case of Fig. 3).

If $C = 0{\cdot}1\mu$F, $R = R_1 = 100$kΩ, the simulated inductance is 1000H with a Q-factor which may be made almost infinite. An actual inductor with such properties is physically unrealizable. Moreover, if such a component could be made, its inductance would be frequency-dependent because of stray capacitance and voltage-dependent because of the non-linear magnetic properties of the core of Mumetal or similar material which would be required.

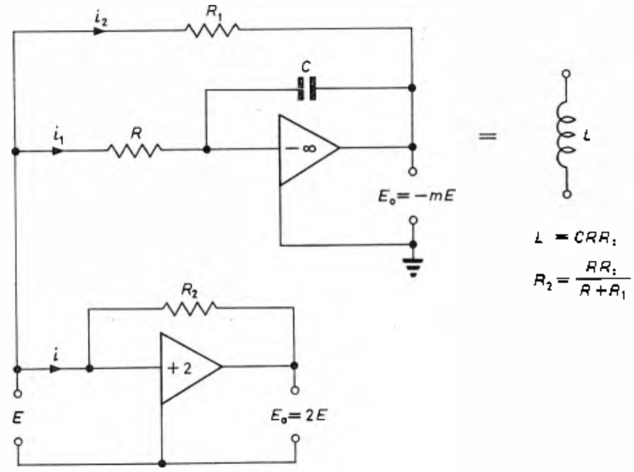


Fig. 4. Variant of Ford and Girling's inductance simulator

Capacitance Simulation

There is not quite the same inducement to develop a capacitance simulator as there is for the inductance counterpart since capacitors of large size and low loss are readily available. Nevertheless there are many applications for a continuously variable unit of high capacitance and wide range. It might seem unreasonable to ask for a low-loss capacitor covering the range 100 to 1000μF, but this can be synthesized without difficulty and its value may be varied by a single potentiometer or alternatively by a ganged pair.

In a recent article[2] the author discussed one way of achieving capacitance multiplication. Some alternative methods will now be considered. All of them owe something to the work of Ford and Girling. Schematic circuits are shown in Figs. 5 and 6. Fig. 5(a) is derived from Fig. 3 merely by interchanging the positions of R_1 and C and adjusting the analysis to suit the change. The results show an obvious correspondence with the inductance case and there is no point in repeating the simple algebraic manipulations.

In Fig. 5(b) the integrating RC amplifier is replaced by a differentiator CR_1. In Fig. 6 the two amplifiers are simply connected in parallel and not in tandem as in Fig. 5. One point which emerges is that the capacitance multiplying factor is no longer the simple ratio of two resistances; instead the factor is $1 + (R_1/R)$. If desired, another C may be shunted across R in the $+2$

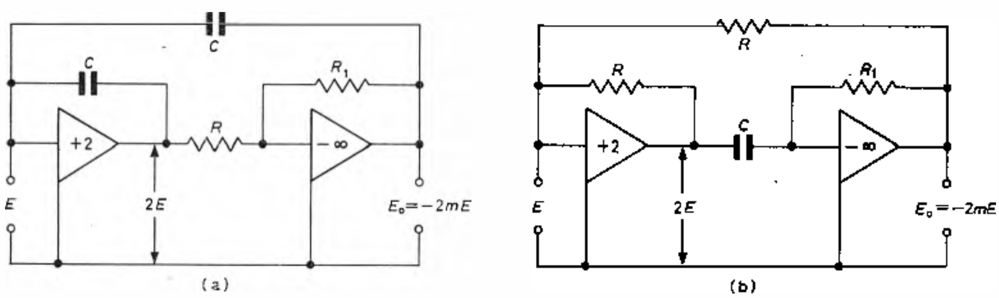
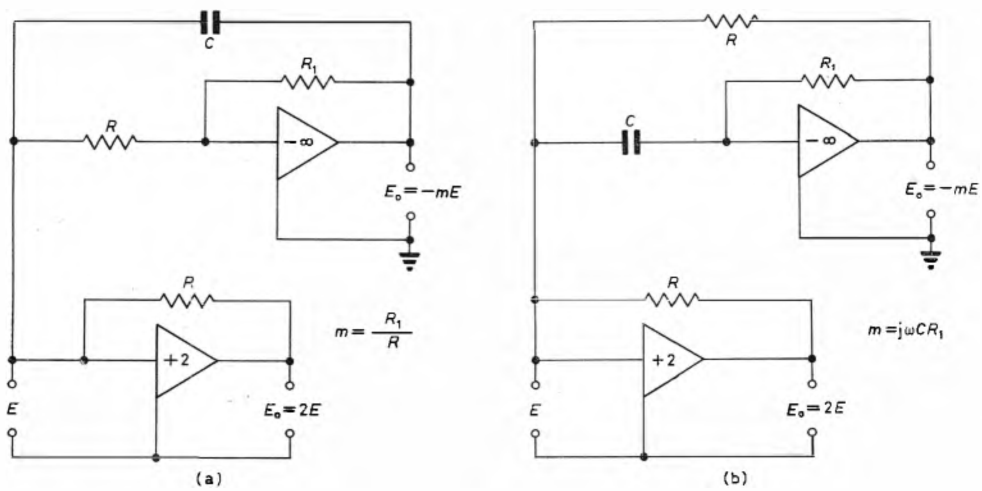


Fig. 5. Capacitance multipliers

(a) Simulated capacitance $C_1 = (2R_1/R)C$, $m = R_1/R$ (b) Capacitance $C_1 = (2R_1/R)C$, $m = j\omega CR_1$

Fig. 6. Alternative capacitance simulators

Simulated capacitance $C_1 = (1 + R_1/R)C$



amplifier. This will produce a negative input capacitance sufficient to restore the multiplying factor to the value of R_1/R.

Some Practical Amplifier Circuits

Ideally the amplifiers for use in these applications should be wide-band devices, d.c. coupled throughout and with the input and output terminals at the same direct potential so that, for example, a resistance could be connected between input and output without requiring a blocking capacitor and without upsetting the amplifier bias arrangements. To meet all these requirements calls for a complex circuit using a dozen or more transistors and requiring dual power supplies. A variety of commercially-available integrated-circuit operational amplifiers could be adapted for use in reactance convertors but much simpler circuits can be designed to give a satisfactory performance, provided only that the input impedance is moderately high and the output impedance very low. Intelligent design will also ensure that a large output voltage swing is available, which in favourable cases can have a peak-to-peak value not far short of the supply voltage.

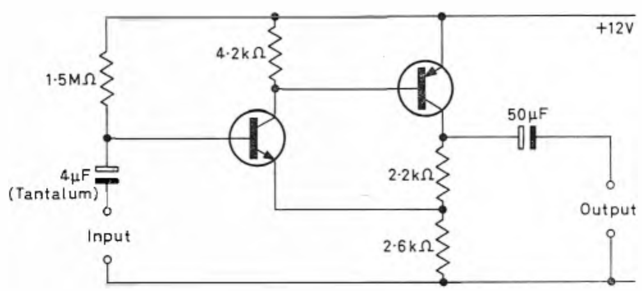


Fig. 7. Non phase-reversing amplifier with gain = +2

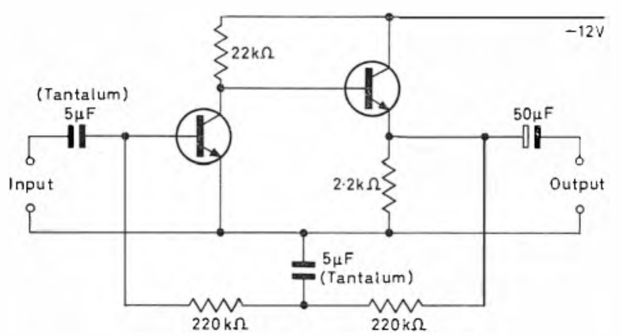


Fig. 8. High-gain phase-reversing amplifier

A suitable circuit for the +2 amplifier is shown in Fig. 7. It employs a complementary pair of silicon planar epitaxial transistors and achieves near-perfect linearity by the use of strong negative feedback. Measured resistor values are shown but quite minor adjustments would allow for the use of close-tolerance preferred-value components. The base bias resistor should in any event be set by trial to give symmetrical waveform clipping at the overload point.

Fig. 8 shows one possible version of the high-gain phase-reversing amplifier. The base-bias circuit shown gives strong d.c. feedback which stabilizes the operating conditions. For use down to low frequencies, a large decoupling capacitor is necessary, (5μF in this particular case). This, in conjunction with the 220kΩ resistors, has a very long time-constant which is sometimes objectionable. As an alternative, a single bias resistor could be used, connected between the positive supply line and the base of the first transistor. Unfortunately, this modification gives relatively poor temperature stability.

Individually, the two amplifiers are quite stable under any normal conditions of use, even when using high-gain v.h.f. transistors. It may be found that, when they are connected in series or parallel in a reactance convertor, instability is encountered. This may take the form of h.f. oscillations, most readily provoked by connecting a capacitive load, such as an oscilloscope input cable, across the output of the phase-reversing amplifier. A good earth connexion, combined with earthing arrangements which avoid ground loops, is generally enough to restore stability. In most of the circuits, a still better arrangement is to take the desired output signal from the end of the +2 amplifier, which seems inherently more stable than the other.

Measurements made on inductance simulators using these two amplifiers gave close agreement with the theory. Another test was made by shunting a known capacitance across the input terminals of the circuits of Figs. 3 and 4 and adjusting the negative resistance circuit to produce continuous oscillations. The calculated frequency agreed with the observed value.

Refinements of the Theory

In an appendix to the author's earlier article[2] an expression was derived for the gain of a feedback amplifier of finite intrinsic gain and finite impedance. This theory can be invoked to derive more exact expressions for the characteristics of reactance convertors. Some tedious algebra is involved and the analytical effort required is scarcely justified by the resulting small improvement in accuracy. A fractional change in the setting of the negative resistance circuit is sufficient to offset the imperfections of real as distinct from ideal amplifiers and the value of the simulated reactance is not a strong function of the open-loop amplifier gain.

Conclusion

There is a wide range of useful applications for reactance convertors of the type described. The idea of an active circuit is particularly attractive now that integrated-circuit operational amplifiers are becoming available at reasonable prices. A great deal of fundamental work remains to be done and there is scope for the development of practical circuits which are at once simple and predictable in performance.

Most circuits operate well up to a frequency of a few hundred kilocycles per second but there are problems at frequencies in the h.f. and v.h.f. bands. Very few of the operational amplifiers available today have bandwidths in excess of 12Mc/s. Fortunately, many of the most useful applications of reactance convertors are at the low-frequency end of the spectrum and it is here that the inductance simulator is particularly useful.

Finally, it is perhaps worth emphasizing that one of the most valuable features of an active impedance convertor is the possibility of applying local or remote control of the simulated reactance by purely electronic means. Tunable oscillators, filters and self-balancing a.c. bridges are items in which this feature is of interest and concern.

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2. Butler, F. Active Impedance Converters. *Wireless World.* 71, 600. (1965).

Emitter-Coupled Monostable Multivibrator Design

By R. Kitai*

This class of multivibrator is widely used in a particular mode in which the output pulse duration is nearly linearly related to the voltage at the base of one of the transistors. However, this linear-mode region has upper and lower limits. It is shown that a stage can be designed, using a universal family of curves, provided that a resistor ratio is put equal to a d.c. voltage ratio. Little sacrifice if any is incurred by using this restriction. The most linear region of operation is established, and account is taken of transistor terminal voltages. Theory is confirmed by experiment.

(Voir page 694 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 701)

THE emitter-coupled monostable multivibrator of Fig. 1 is well known to provide an output pulse, the duration of which is very nearly linearly proportional to the voltage v of the base of transistor VT_1 under certain operating conditions. The circuit is often used to provide a linear-variable delay, and it is also widely applied in low-frequency pulse duration modulation systems and in pulse generators. Of its four possible modes of operation, determined by whether either or both transistors saturate or not, the linear modulation characteristic obtains only when transistor VT_2 is saturated in the stable state with VT_1 unsaturated during the astable state. Now the allowable range of v for linear variation of output pulse duration t is restricted. As v is lowered towards E_2, t falls and it vanishes at some stage, while at an upper limiting value of v, either the circuit becomes astable and oscillates freely, or transistor VT_1 reaches saturation and the linearity of the t/v characteristic ceases abruptly. Furthermore, the range of values of v for linear-mode operation is quite small compared with the supply voltage $E_1 + E_2$, calling for care in design. Unfortunately there are so many parameters which determine the operation that design procedures are often haphazard. Also, it is not easy to show, theoretically, why the modulation characteristic is nearly linear[1,2].

This article shows that if the ratio of supply voltages E_2/E_1 is put equal to the ratio of resistors R_p/R_e, where $R_p = R_cR_b/(R_c+R_b) \simeq R_c$ then the stage can readily be designed using the universal family of curves shown in Fig. 2. The curves relate normalized t to normalized v for the linear range. Furthermore, by resorting to this design procedure, the linearity of the t/v characteristic can readily be demonstrated theoretically, non-linearities can be calculated, and the most linear mode of operation established. The theory also illustrates the sensitivity of the response duration t to variations in v and in component values. The curves of Fig. 2 are based on a theory which assumes ideal transistors: Often these curves are accurate enough to warrant direct use in practice. A more precise analysis is also given, and it is shown that when transistor terminal voltages and finite current gains are taken into account, the curves of Fig. 2 are still valid, but with $y = v/E_1$ replaced by $y \simeq \dfrac{v + 1{\cdot}5\,V_{be}}{E_1}$ where V_{be} is the conducting base-emitter voltage. This correction is small in many cases. Finally, other ratios of R_p/R_e are considered.

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Ideal Theory

In developing the theory it is particularly convenient to take the static-state emitter voltage as zero. Obviously only a single source of supply $(E_1 + E_2)$ need be used in

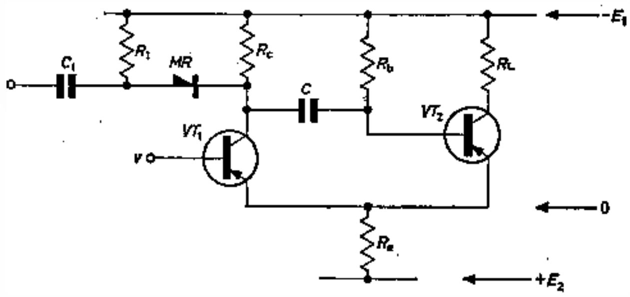


Fig. 1. Basic circuit.
The emitter voltage in the static state is taken as zero

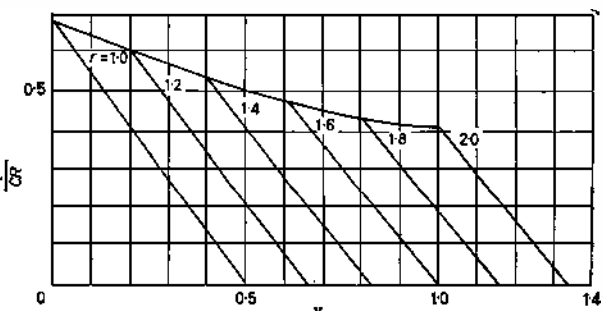


Fig. 2. Curves of t/CR against y for r in the range 1·0 to 2·0.
For ideal transistors $y = v/E_1$
For practical transistors $y \simeq (v + 1{\cdot}5\,V_{be})/E_1$

practice. Throughout the article all voltages E_1, E_2 and v refer to magnitudes, and signs are taken care of so that the results hold for both pnp and npn transistors.

For VT_1 cut off in the static state with VT_2 saturated, v must be positive for pnp transistors, and $h_{FB(min)}$ for VT_2 must exceed R_b/R_L. The ideal theory which follows assumes the following transistor properties:

(1) When conducting, the base-emitter voltage of a transistor is zero.

(2) When saturated, the collector-emitter voltage is zero.

(3) The common-base current gain α for VT_1 is unity.

(4) Base leakage currents and stored charges are negligible.

C_1 and R_1 constitute a differentiating circuit, producing spikes from an input trigger source. Diode *MR* passes positive spikes, via *C*, to the base of VT_2 and the emitters are pulled positive by the same amount. If the trigger voltage spikes reach *v* volts, multivibrator action turns VT_1 on and VT_2 off. The emitters of VT_1 and VT_2 are now held at *v* volts. The switch-over action is usually so fast that the collector of VT_1 goes positive more rapidly than the triggering spike, and diode *MR* becomes reverse-biased, thereby isolating the trigger source—a very desirable characteristic of the circuit. Base and collector waveforms for VT_2 are shown in Fig. 3. In the astable state VT_1 may or may not be saturated, depending on component and voltage values. At $t = 0$ the current step through VT_1 passes through R_c in parallel with R_b. Let:

$$R_p = \frac{R_c R_b}{R_c + R_b}$$

In practice $R_b \gg R_c$, so $R_p \simeq R_c$.

If VT_1 does not saturate, the voltage step at the collector of VT_1 is:

$$e = \frac{E_2 - v}{R_e} R_p$$

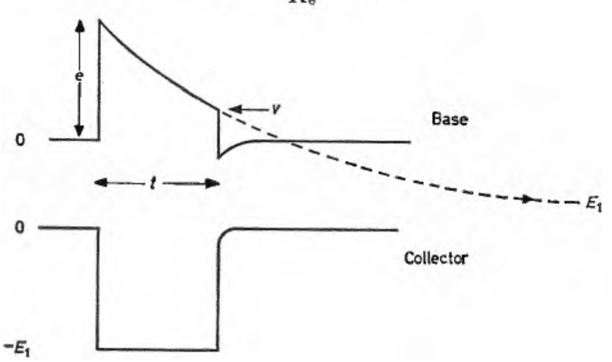


Fig. 3. Base and collector waveforms for transistor VT_2 neglecting transistor base-emitter voltage drop

and if VT_1 saturates:

$$e = E_1 + v$$

With VT_2 cut off, *C* charges through R_b in series with R_c. Let:

$$R = R_b + R_c$$

where $R \simeq R_b$ in practice. Then for VT_1 not saturated:

$$t = CR \ln \left[\frac{(E_2 - v) R_p/R_e + 1}{E_1 + v}\right] \quad \ldots\ldots\ (1)$$

and for VT_1 saturated:

$$t = CR \ln \left[\frac{2E_1 + v}{E_1 + v}\right] \quad \ldots\ldots\ldots\ (2)$$

Now restrict the analysis to the condition:

$$R_p/R_e = E_2/E_1 = r, \text{ say} \quad \ldots\ldots\ldots\ (3)$$

where $E_2/E_1 = R_e/R_L$

Also let:

$$y = v/E_1 \quad \ldots\ldots\ldots\ (4)$$

Then equation (3) becomes:

$$t/CR = \ln \left[\frac{r(r - y) + 1}{1 + y}\right]_{\text{VT}_1 \text{ unsaturated}} \quad \ldots\ (5)$$

and equation (4) becomes:

$$t/CR = \ln \left[\frac{2 + y}{1 + y}\right]_{\text{VT}_1 \text{ saturated}} \quad \ldots\ldots\ (6)$$

Using equations (5) and (6) the following relations are derived:

$$t = 0 \text{ at } y = \frac{r^2}{r + 1} \quad \ldots\ldots\ldots\ (7)$$

The transition to saturation occurs at a value of *y* given by:

$$y_{sat} = r - 1 \quad \ldots\ldots\ldots\ (8)$$

so that:

$$(t/CR)_{sat} = \ln \left(\frac{r + 1}{r}\right) \quad \ldots\ldots\ldots\ (9)$$

The linear range of *y* is $\left(\dfrac{1}{1 + r}\right)$ ……………… (10)

and hence:

$$\frac{\text{Linear range of } v}{\text{Total d.c. supply voltage}} = \frac{1}{(1 + r)^2} \quad \ldots\ldots\ldots\ (11)$$

Equations (5) and (6) are plotted in Fig. 2 for values of *r* between 1 and 2. The steep 'linear' regions are obtained from equation (5). The curves show that the circuit is sensitive to small changes in both *y* and *r*, so that components must have close tolerances for good agreement with theory. On the other hand, the slopes of the linear-region curves are not critically dependent on *r*; nor are the linear ranges of *t* and *v*. In practice, then, it is usual to construct the circuit with a trimmer adjustment in the supply source *v*.

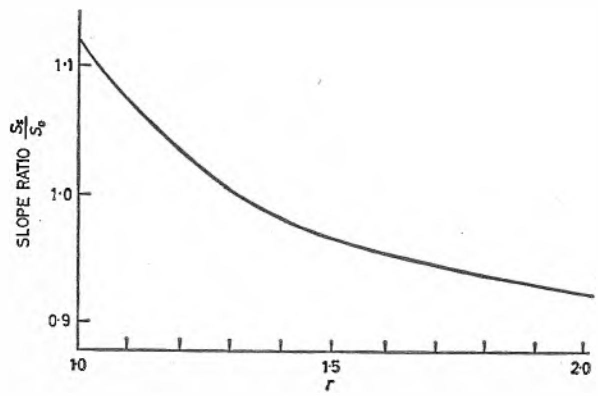


Fig. 4. Ratio of slopes at extremities of linear range against *r*

Response Linearity

The slope of the linear-mode region is obtained by differentiating equation (5), whence:

$$S = d/dy\,(t/CR) = -\frac{r}{(r - y) + 1} - \frac{1}{1 + y} \quad \ldots\ (12)$$

When plotting the linear range of Fig. 2 it is found that the curve for $r = 1$ is bowed slightly downwards, while at $r = 2$, the curve is bowed slightly upward. Thus the most linear mode of operation lies between these values of *r*.

A useful measure of linearity is the ratio of the slopes at the extremities of the 'linear' range. Thus, for a given *r*, let S_o be the slope of the curve of t/CR against *v* at t = 0 and let S_s be the slope at the transition to saturation. Using equations (7), (8) and (12):

$$S_o = -\frac{(r + 1)^2}{r^2 + r + 1}$$

$$S_s = -\frac{r^2 + r + 1}{r(r + 1)}$$

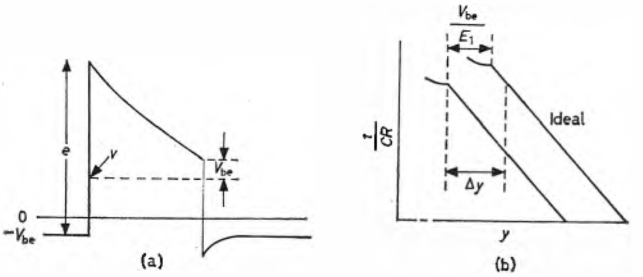


Fig. 5 (a). Waveforms at base of transistor VT_2, taking V_{be} into account. (b). At saturation, the ideal-transistor value of y is reduced by Δy for practical transistors, where $\Delta y > V_{be}/E_1$

Hence:

$$\text{Slope ratio} = S_s/S_o = \frac{(r^2 + r + 1)^2}{r(r+1)^3} \quad \ldots\ldots\ldots (13)$$

Equation (13) is plotted in Fig. 4. The slope ratio is unity at $r^3 = r + 1$, whence:

$$r \simeq 1{\cdot}325$$

Thus linearity is best for r in the region of 1·3.

Correction of Ideal Theory for Practical Transistors

The ideal theory is readily modified to take into account the finite base-emitter voltage V_{be}, the collector-base voltage V_{cb}, and the common-base current gain of the conducting transistor VT_1. Transistor properties not accounted for here are base leakage currents and stored charge. The former are usually negligible at room temperatures in practical circuits, while base stored charge becomes significant only when it is comparable with the charge stored in capacitor C in the static state, viz, CE_1. This condition can usually be avoided by proper selection of transistor types and capacitor values.

The modified theory is given in the Appendix. The assumptions made are that V_{be} for the saturated transistor VT_2 is near the value of V_{be} for transistor VT_1 when not saturated, that V_{be}/E_1 is small and α is near unity, and that regeneration sets in to return the circuit to its stable state when the base-emitter voltage for transistor VT_2 reaches zero. These are all closely realized in practice. Then the waveform at the base of transistor VT_2 is that shown in Fig. 5(a). The results of the theory are as follows:

(1) The slope of the curves of Fig. 2 are altered only slightly. Referring to equation (12), the modified theory has the effect of increasing the first term, and decreasing the second term by small amounts.

(2) For any given r, the value of y at which $t = 0$ is reduced from $y = r^2/(r + 1)$ by an amount which is very near

$$\frac{r^2(1-\alpha)}{(1+r)(1+\alpha r)} + \frac{r+2}{r+1}\,(V_{be}/E_1)$$

Here the first term is usually small compared with the second term. (Perhaps 10 per cent of the magnitude of the second term). Also $\dfrac{r+2}{r+1} = 1{\cdot}5$ at $r = 1$, 1·4 at $r = 1{\cdot}5$ and 1·333 at $r = 2$. So the value of y for $t = 0$ is shifted down by an amount of the order $1{\cdot}5\,V_{be}/E_1$ for cases of practical interest.

(3) For a given r, the value of y at the transition to saturation is reduced below $r - 1$ by an amount which is very near V_{be}/E_1. Also the value of t/CR_{sat} is reduced from $\ln((1+r)/r)$ to $\ln\left[\dfrac{1 + \alpha r - (V_{be}/E_1)}{r}\right]$. This fall in $(t/CR)_{sat}$ is small so that the linear range of t is hardly affected. However, on referring to Fig. 5(b), the value of y in the ideal analysis would need to be reduced by an amount Δy which is in excess of V_{be}/E_1 to maintain the new value of $(t/CR)_{sat}$, and the Appendix shows that the total reduction in y is near

$$\Delta y = \frac{r^2(1-\alpha)}{r^2 + r + 1} + \frac{(r+1)^2}{r^2 + r + 1}\,(V_{be}/E_1)$$

Now $r^2/(r^2 + r + 1)$ varies between 0·333 at $r = 1$ and 0·57 at $r = 2$. Also $(r + 1)^2/(r^2 + r + 1)$ varies between 1·333 at $r = 1$ and 1·29 at $r = 2$. Since α and V_{be} are usually not precisely-known quantities in practice, it suffices to assume that y_{sat} is reduced below the 'ideal' value which gives the new $(t/CR)_{sat}$ by an amount:

$$\left(\frac{1-\alpha}{2}\right) + 1{\cdot}3\,(V_{be}/E_1)$$

The latter two considerations confirm that the slope of the curve of t/CR against y is hardly altered. Comparing the modified values of y at $t = 0$ and at saturation, it suffices in design to shift the curves of Fig. 2 bodily to the left by an amount $1{\cdot}5\,(V_{be}/E_1)$ or alternatively, to replace $y = v/E_1$ by:

$$y = \frac{v + 1{\cdot}5\,V_{be}}{E_1}$$

in Fig. 2.

Experiment and Theory

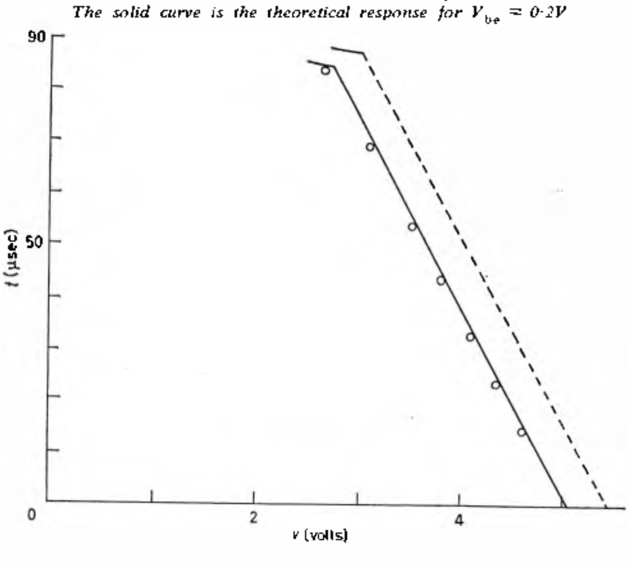
To verify the theoretical results, a circuit was constructed using OC44 germanium pnp transistors which were selected to give $h_{FB} > 100$. Values chosen to give $r = 1{\cdot}5$ were $R_e = 1{\cdot}8\text{k}\Omega$, $R_c = 2{\cdot}8\text{k}\Omega$, $R_b = 72\text{k}\Omega$, $R_L = 1{\cdot}2\text{k}\Omega$, $E_1 = 6\text{V}$, $E_2 = 9\text{V}$ (15V d.c. supply). Hence $R_p \simeq 2{\cdot}7\text{k}\Omega$, $R = 74{\cdot}8\text{k}\Omega$.

Fig. 6(a) shows t against v given by the 'ideal' expression of equation (5) (dashed line) as well as the modified curve, (solid line), for $C = 2\,300\text{pF}$, assuming $V_{be} = 0{\cdot}2\text{V}$. Experimental results are seen to lie very near the latter.

Resistor Tolerances

The output pulse duration is more dependent on the values of resistors R_c and R_e than on other resistors in the circuit. To assess the effects of small variations in these resistor values, write:

Fig. 6. Theoretical and experimental responses for a practical case.
The dashed curve shows the theoretical response for ideal transistors
The solid curve is the theoretical response for $V_{be} = 0{\cdot}2V$



$E_2/E_1 = r$ (as before), and

$R_p/R_e = r + \Delta$

where Δ represents a small deviation from r. Also, let t_0 be the pulse duration for $\Delta = 0$. Then equation (5) becomes:

$$t/CR = \ln\left[\frac{r(r-y)+1}{1+y} + \frac{(r-y)\Delta}{1+y}\right]$$

so that, for small Δ,

$$(t - t_0)\,/CR = \ln\left[1 + \frac{(r-y)\Delta}{r(r-y)+1}\right]$$

$$\simeq \left[\frac{r-y}{r(r-y)+1}\right]\Delta = k\Delta \quad \text{say}$$

Using equation (7):

$$\text{at } t = 0,\ k = \frac{r}{r^2 + r + 1}$$

and this varies between 0·333 for $r = 1$ and 0·286 for $r = 2$. Using equation (8), at saturation $k = 1/(r+1)$, varying between 0·5 and 0·333 over the same range in r.

Hence the effect of a small mismatch Δ in resistance ratio is to shift the curves of Fig. 2 by a nearby constant amount which is of the order $0{\cdot}4\Delta$ in practical circuits.

To illustrate the effect of Δ, consider the experimental circuit, and assume that R_e is reduced from 2·8kΩ to 2·7kΩ (a reduction of 3·6 per cent). From Fig. 3 the ideal mid-range value of y is 0·7 where:

$$t_0 \simeq 0{\cdot}25\,CR$$

For this value of y, and for $r = 1{\cdot}5$, $k = 0{\cdot}364$.
Also $R_p/R_e \simeq 1{\cdot}445$ so that $\Delta = -0{\cdot}055$, and at mid-range, $(t - t_0)/RC = 0{\cdot}02$.

Hence:

$$\frac{t - t_0}{t_0} = 8 \text{ per cent}$$

This result was confirmed experimentally. For $t_0 = 45\mu$sec, reducing R_e from 2·8kΩ to 2·7kΩ resulted in a reduction of t_0 by 3·5μsec. Similarly shunting R_e by 47kΩ (i.e. reducing its value by near 4 per cent) resulted in an increase of t which was near constant at 4μsec for all values of v.

Alternative Ratio R_p/R_e

The ratio R_p/R_e in equation (1) was arbitrarily put equal to the ratio $E_2/E_1 = r$, and hence equation (5) is in the simplest form possible. The question now is whether anything is gained by putting:

$$R_p/R_e = m \quad E_2/E_1 = mr$$

where m is a constant other than unity. On this assumption the following more general expressions hold:

$$t/CR = \ln\left[\frac{mr(r-y)+1}{1+y}\right]_{VT_1 \text{ unsaturated}} \quad \text{.. (5a)}$$

$$t = 0 \text{ at } y = \frac{mr^2}{1+mr} \quad \text{.......... (7a)}$$

Transition to saturation is at:

$$y_{sat} = \frac{mr^2 - 1}{mr + 1} \quad \text{.............. (8a)}$$

so that:

$$(t/CR)_{sat} = \ln\left[\frac{mr^2 + 2mr + 1}{mr(1+r)}\right] \quad \text{....... (9a)}$$

$$\text{Linear range of } y = \frac{1}{mr+1} \quad \text{................. (10a)}$$

Now the circuit becomes unstable when y is negative, so $mr^2 > 1$ is required at all times. In the limit, when $mr^2 = 1$, using equation (9a):

$$\max.\,(t/CR)_{sat} = \ln 2 \simeq 0{\cdot}693$$

and this is independent of m.

Hence nothing is gained in respect of increasing the linear range of t by putting m at some value other than unity.

For $m > 1$ the slopes of the linear-range curves of Fig. 2 are increased, and the response is more sensitive to variations in v. The converse holds for $m < 1$. Families of curves for other values of m can be constructed, but the need to use them will arise only if a different sensitivity to variations in v is desired. For design purposes, a circuit having $m = 1$, and r of the order of 1·3 is highly linear: Also the maximum linear value of t/CR is not markedly below the maximum possible value, and a cushioning beyond the upper limit of the linear range is also inherent in the characteristic. In this region, in which VT_1 is saturated, a lowering of v produces only a small increase in t, and the effect is to partially saturate t for values of v which are still well above the onset of instability at $v = 0$. All of these are desirable characteristics in most applications.

Acknowledgment

The author is indebted to Dr. N. K. Sinha for helpful comments.

Appendix

Referring to Fig. 5(a) it is assumed that the 'break-in' base-emitter voltage for transistor VT_2 is zero and that when conducting the base-emitter voltage is V_{be} for both VT_1 and VT_2. Then for VT_1 not saturated:

$$e = \alpha\left[\frac{E_2 - v - V_{be}}{R_e}\right] R_c = \alpha r\,[r - y - b]$$

where $b = V_{be}/E_1$.

Also:

$$E_1 + v + V_{be} = (E_1 + e - V_{be}) \exp\{-(t/CR)\}$$

whence

$$(t/CR) = \ln\left[\frac{(1-b) + \alpha r\,(r - y - b)}{1 + y + b}\right] \quad \text{... (i)}$$

(a) *Slope*

The slope of t/CR against y in the linear range is:

$$S = -\frac{\alpha r}{1 - b + \alpha r(r - y - b)} - \frac{1}{1 + y + b} \quad \text{...... (ii)}$$

compared with the ideal theory slope:

$$S = -\frac{r}{1 + r\,(r - y)} - \frac{1}{1 + y} \quad \text{.............. (iii)}$$

To compare values, a numerical example will be taken. Let $r = 1{\cdot}5$, $\alpha = 0{\cdot}99$, $y = 0{\cdot}7$ (mid-range value for ideal case) and $b = 0{\cdot}03$.

Substitution in equation (iii) gives:

$$S = -0{\cdot}6818 - 0{\cdot}5882 = -1{\cdot}270$$

and substitution in equation (ii) gives:

$$S = -0{\cdot}7026 - 0{\cdot}5780 = -1{\cdot}2806$$

Hence α and b increase the first term in equation (iii) and reduce the second term. The slopes are very close to each other.

(b) $t = 0$

From equation (i) $t = 0$ at

$$y = y_0 = \frac{\alpha r^2 - b\,(2 + \alpha r)}{1 + \alpha r}$$

Writing this in terms of the ideal value:

$$y_o = \left(\frac{r^2}{1+r}\right)\left(\frac{1+r}{1+\alpha r}\right)\left[\frac{\alpha r^2 - b(2+\alpha r)}{r^2}\right]$$

$$= \left(\frac{r^2}{1+r}\right)\left[\frac{(\alpha r+\alpha) - (b/r^2)[\alpha r^3 + r(2+\alpha)+2]}{1+\alpha r}\right]$$

Now:

$$\frac{\alpha r+\alpha}{1+\alpha r} = 1 - \frac{1-\alpha}{1+\alpha r}$$

and:

$$\frac{\alpha r^2 + r(2+\alpha)+2}{\alpha r+1} = r + \frac{1+\alpha}{\alpha} - \frac{1-\alpha}{\alpha(\alpha r+1)}$$

the last term of which is negligible. Also $\frac{1+\alpha}{\alpha} \simeq 2$.

Hence:

$$y_o \simeq \frac{r^2}{1+r} - \frac{r^2(1-\alpha)}{(1+r)(1+\alpha r)} - \left(\frac{r+2}{r+1}\right)V_{be}/E_1$$

Thus y_o is reduced by the latter two terms in the above.

(c) $(t/CR)_{sat}$

At saturation, $e = E_1 + v + V_{cb}$, where V_{cb} is the collector-base saturation voltage for transistor VT_1. Transition to saturation occurs at:

$$\alpha r\left(r - y - (V_{be}/E_1)\right) = 1 + y + (V_{cb}/E_1)$$

Whence:

$$y_{sat} = \frac{1}{1+\alpha r}\left[\alpha r^2 - 1 - \left(\frac{\alpha r V_{be} + V_{cb}}{E_1}\right)\right]$$

Using these relations, one can write:

$$r - y_{sat} - (V_{be}/E_1) = \frac{1 + r - \left(\frac{V_{be} - V_{cb}}{E_1}\right)}{1+\alpha r}$$

where the right-hand side is very near unity in all cases of practical concern. With little error, then, if $b = V_{be}/E_1$

$$y_{sat} = r - 1 - b \quad \ldots\ldots\ldots\ldots\ldots \text{(iv)}$$

and hence:

$$(t/CR)_{sat} \simeq \ln\left(\frac{1+\alpha r - b}{r}\right) \quad \ldots\ldots\ldots\ldots \text{(v)}$$

where the ideal value is $\ln((1+r)/r)$. So the saturation value of t/CR is slightly lower than the value given by Fig. 2. For $(t/CR)_{sat}$ given by equation (v), the corresponding ideal value of y is obtained from the relation:

$$\frac{1-b+\alpha r}{r} = \frac{r(r-y)+1}{1+y}$$

or:

$$y = \frac{(r^3-1) + r(1-\alpha) + b}{r^2 + \alpha r + 1 - b}$$

Substituting:

$$r^3 - 1 = (r-1)(r^2+r+1)$$

dividing numerator by denominator, and then approximating the denominator to $r^2 + r + 1$:

$$y \simeq (r-1) + \frac{r^2(1-\alpha)+br}{r^2+r+1}$$

At saturation, then, the total shift in y below the ideal value is made up of the second term in the above added to b, viz.:

$$\frac{r^2(1-\alpha)}{r^2+r+1} + \frac{(r+1)^2}{r^2+r+1}V_{be}/E_1$$

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Thin Film Measurement

Because of the need, through increased miniaturization and microminiaturization, to measure the thickness of evaporated thin films used in the production of resistors, capacitors and transistors for integrated circuits, Rank Taylor Hobson has developed the Talystep step-height measuring instrument.

It is often necessary to know the thickness, in millionths of an inch and less, of thin films as this governs the final performance of the integrated circuit and the more closely the parameters of the various components can be controlled the more the circuit approaches the designer's optimum.

'Talystep 1' is a stylus instrument designed primarily for the thickness measurement of thin film deposits often formed by evaporation. Measurements are made by traversing the stylus either across a test groove formed in the deposit, or over the edge of the deposit. Vertical movement of the stylus is amplified electronically and recorded as a graphical representation of the difference in levels between the surfaces of the substrate and the deposit. To some extent the instrument will also reveal the surface roughness of a limited range of surfaces.

A low stylus force, of 1 to 30 milligrams, adjustable by a manual control, is employed to avoid the risk of damaging the softer types of deposit.

The highest magnification of the instrument, ×1 000 000 is adequate for the very thin films now being produced. Vertical resolution is generally better than 0·000 000 02 inch (5Å).

A drive mechanism traverses the stylus across the surface of the specimen in an accurate movement over a maximum traverse length of about 0·1in. Three traverse speeds give horizontal magnifications of the recorder graph of ×15 approx., ×200 and ×2000.

A simple viewing microscope is included as an aid to selecting the area of specimen to be measured. Filter circuits incorporated in the amplifier allow the higher frequency components to the signal, representative of residual roughness, to be suppressed so that the general levels of the upper and lower surfaces of the specimen under inspection can be more readily assessed. The instrument has facilities for compensating for small residual curvature of the specimen surface.

The Talystep step-height measuring instrument

The Anode-Controlled Pentode as an Electrometer

By J. H. B. Gould

In this article it is shown that an ordinary 'receiving' type valve can successfully be employed in the input stage of an electrometer. A special 'inverted' circuit is used in which the anode acts as the control electrode while the output is taken from a load resistor in the grid circuit.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

An electrometer is a device for measuring potential differences at very high impedance, or for measuring currents in the pico-ampere or femto-ampere range. There are several types of electrometer in use, but the simplest and most flexible is the valve electrometer. Basically, this is nothing more than a highly sophisticated valve voltmeter, but whereas the normal valve voltmeter has an impedance of the order of $10^9\Omega$, the valve electrometer's input impedance must be around $10^{14}\Omega$ or even better.

It is hardly surprising that impedances of this nature are beyond the capabilities of normal valves used in a normal manner, and special electrometer valves have been developed specifically for the purpose. Clearly, these valves must possess a very high input impedance; they also need to have a low capacitance between the grid and the other electrodes, otherwise, with currents around 10^{-14}A, it could take several minutes for the instrument to settle down to a reading without the assistance of negative feedback.

Yet despite all this, it is perfectly possible to use an ordinary 'receiving' valve in the input stage of an electrometer and still achieve results equal to or even better than those obtainable with many types of valve designed specifically for the purpose. But while the valves themselves are normal, their manner of use is not.

One expedient is drastically to reduce the electrode and filament voltages. Among other things, this cuts down the ion current on the grid. As an example[1], the pentode section of a 1U5 diode-pentode will draw a grid current of the order of 10^{-14}A when the filament voltage is reduced from its rated value of 1·4V to a value of 0·5V, with the anode voltage reduced to 14. Under these conditions the amplification factor is 8 when the valve is operated as a triode, and 180 when it is run as a pentode.

The 'Inverted' Valve

A more radical approach is to operate the valve with its anode acting as control electrode, taking the output across a load resistor in the grid circuit. To this effect the grid carries a positive bias, and the anode 'signal' must at all times be negative with respect to the filament. Thus, for a triode, the valve operates in a completely 'inverted' mode. If sufficient care is taken in design and construction, this configuration can realize control-electrode currents an order of magnitude or so less than they would be if the valve was operating in the conventional mode with reduced voltages. In practise, the minimum control-electrode current is limited by leakage across the surface film which inevitably exists on the outside of the glass forming the envelope. Unfortunately, there appear to be no suitable valves in which the anode is brought out to a top-cap.

A Simple Multi-Range Electrometer

Fig. 1 shows the schematic diagram of a single-stage valve electrometer, employing a pentode controlled by its anode and screen grid strapped together. A feature of this circuit is the fact that it can be assembled quickly and easily from a few readily available components. By employing a subminiature valve and 'pen-torch' cell the whole circuit apart from the multi-range test meter could be built into a shielded 'probe', which is one way of keeping the anode-circuit leakance and capacitance to a minimum. The grid circuit is far from critical in this regard, as its impedance is relatively very low indeed (which can be seen from the 'ohms' scale of the meter). Therefore, the leads connecting the valve to the meter can be of any convenient length.

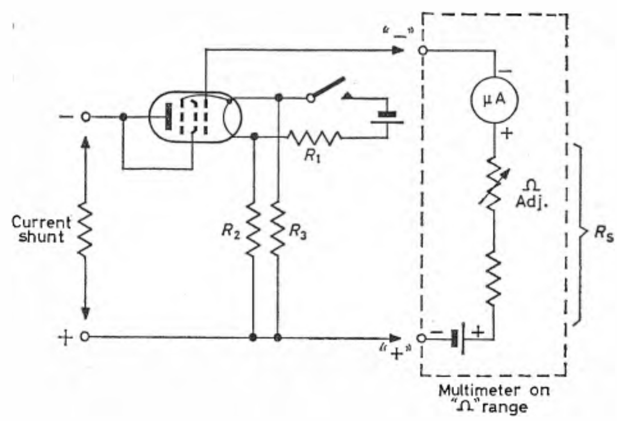


Fig. 1. A simple multi-range electrometer employing an anode-controlled pentode

The measuring range of the device can be selected simply by switching between ohms ranges without making any other adjustments whatsoever. The only proviso is that the meter movement must not be shunted in the process. This usually rules out the lowest ohms range.

Fig. 2(a) shows a family of 'calibration curves' taken on such a device. The topmost curve (15 + 0·44) was obtained using the 0 to 60MΩ range of the multimeter, energized by its internal 15V battery. The centre curve (1·4 + 0·44) was obtained on the 0 to 6MΩ range, using the internal 1·4V cell. The lowest curve shows the characteristic obtained without a 'load' resistance, and thus represents the maximum sensitivity that can be obtained from this particular arrangement; for this, the multimeter was used purely as a microammeter. The voltage to over- of measuring scale, but they arise from exactly the same conditions as battery voltage drift, i.e., a reduced battery voltage compensated by a reduced series resistance, and so they can serve as an indication of the type of effect to be expected. If desired, the voltage can be stabilized in one of several ways.

Another possible source of inaccuracy lies in the voltage of the filament cell, which can also change with time. An extreme case of this is shown in Fig. 3. Here, the voltage has been cut to half its rated value. This has naturally made it impossible to obtain the same maximum

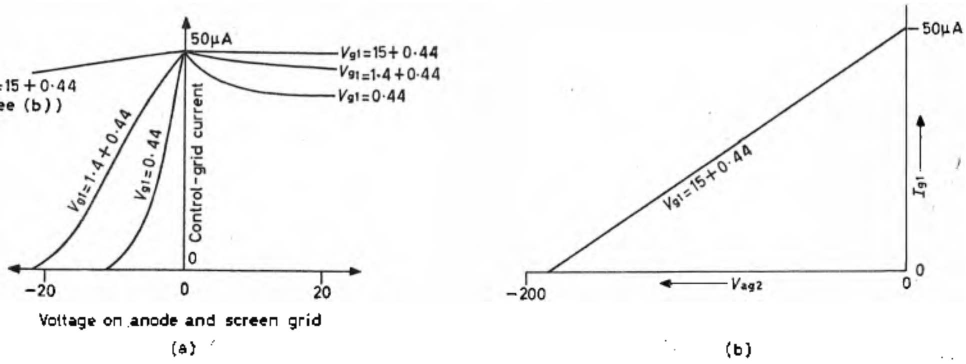


Fig. 2 (a). Characteristics obtained from the circuit of Fig. 1 using pentode type CK 505 AX at full rated filament voltage;
(b). ditto, extended to show the high-voltage range in full

come the resistance of the valve and meter movement in this case comes from the filament supply through voltage divider R_2, R_3.

The effect of the grid-load resistance R_8 is clearly shown in these curves. The higher the grid voltage and resistance, the lower is the slope of the characteristic and the smaller is the sensitivity of the electrometer. At the same time, the relationship between test voltage and meter reading is more nearly a straight line (Fig. 2(b)).

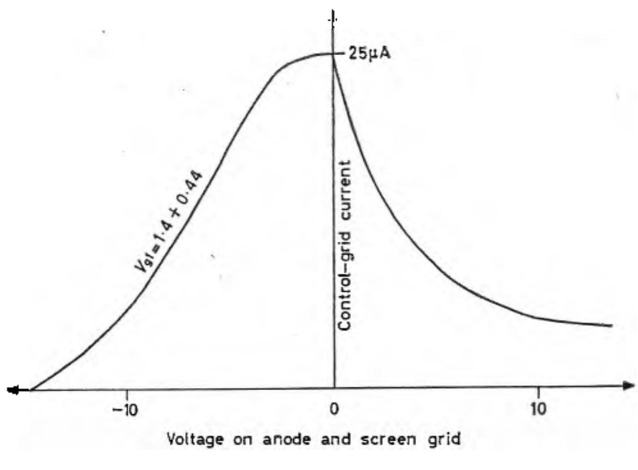
The part of the characteristic in the positive quadrant is of no significance from the electrometer point of view, as the valve draws significant anode current in this regime. It is of interest, however, in so far as the mechanism by which the anode, screen-grid voltage affects the control-grid current is entirely different in the two quadrants. In general terms, electrodes A, G_2 tend to 'steal' electrons from the control grid in the positive quadrant, whereas the effect is almost entirely one of repulsion in the negative-voltage part of the curve. The result as far as G_1 is concerned is much the same, however, as the current falls with increased input voltage, whether it be positive or negative. This fact could lead to very serious errors when the circuit is used to measure low voltages (or currents), for if the instrument was inadvertently connected into the test circuit the wrong way round, it would still be possible to get apparently reasonable readings but with an input impedance well below $10^9\Omega$.

When the electrometer is used with a normal type of multimeter switched to the ohms ranges, the voltage of the internal batteries will affect the accuracy of the readings. Long-term changes of battery or cell voltage would in practise be compensated by setting the 'ohms adjust' control on the multimeter so that the instrument has a specified reading (full-scale if possible) when the input voltage is zero. Some idea of the effect internal battery voltage has on the measuring characteristic can be gained from the three curves of Fig. 2(a). These curves are drawn for the gross changes of voltage associated with a change deflexion on the meter as with full voltage, and the characteristic is far less linear; nevertheless, the slope (transconductance) over the most linear section of the curve is almost exactly the same as it was at full filament voltage. Unfortunately, this fact is of little consolation when the device is supposed to be measuring direct voltage. Again, the voltage could be stabilized if so desired.

Space-charge depletion is illustrated clearly in the positive half of the characteristic.

The object of the voltage divider across the filament is to 'match' the grid circuit of the valve to the multimeter ohm circuit; i.e., it provides a voltage to overcome the grid resistance of the valve at zero anode voltage, so that full-scale deflexion can be obtained within the range of the 'ohms adjust' control on the meter. (This incidentally explains the '+0·44' which keeps appearing on the

Fig. 3. The effect reducing the filament voltage to half its rated value has on the (V_{g1} = 1·4 + 0·44) characteristic for the circuit of Fig. 1



characteristics.) It is tempting to try instead using this voltage to compensate for drifts in filament-cell voltage, by converting it into a negative bias. Unfortunately, the relationship between filament voltage and grid voltage for I_{g1} = const. is far from linear, and has a distressing habit of varying sharply with variations of anode voltage.

Locating a Fault on a Power Line

A completely different application for the anode-controlled valve is shown in Fig. 4. Here, the problem is to measure the charge on a capacitor without significantly affecting the charge in the process.

of the bridge consists of a microammeter scaled in zero sequence current. After a few seconds the microammeter settles down to a reading which represents the voltage on the capacitor. The time-delay element of the scheme (C_2) now causes the relay to release, which in turn releases a simple spring-loaded clamp to lock the movement of the meter. All but two of the relay contacts reset with the armature, but the transistor supply and alarm contacts remain latched. When the operator has taken his reading, he presses a button that unlocks the meter movement and resets the latched contacts to prepare the meter for the next earth fault.

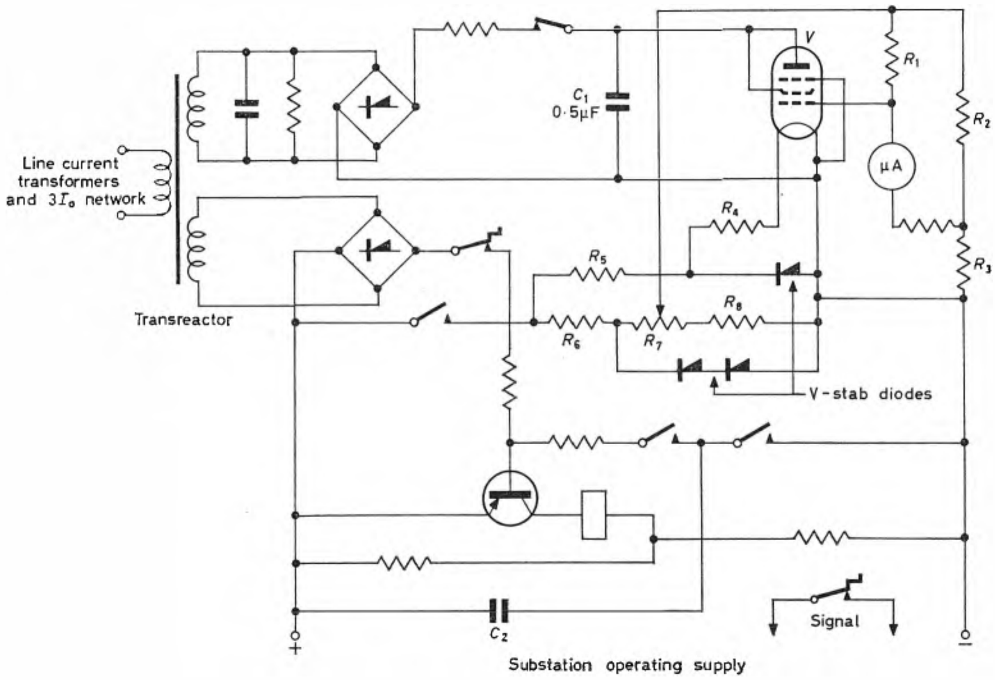


Fig. 4. Arrangement of ammeter type FIU

Fig. 4 shows the schematic diagram of ammeter type FIU, developed by the Central Laboratories of the Uzbekistan Ministry of Power[2]. The object of this device is to measure the zero phase sequence currents at the ends of a power transmission line during a fault to earth, in order to locate the position of the fault along the line. As the majority of earth faults are fleeting in nature and cease to exist once the circuit-breakers have cleared the line, the measuring process must be completed in an interval of about 0·06 to 0·1sec, which is the time it takes the line protection to bring out the breakers feeding the fault. The instrument also has to store the information until the substation staff has had a chance to read it. This it does by means of capacitor C_1, which acquires a charge proportional to the value of zero sequence current $3I_0$ obtained from the sequence network. In the event of an earth fault on the line, the high value of zero sequence current causes the relay scheme (represented in the lower part of the diagram) to operate, isolating capacitor C_1 from the transreactor and leaving it connected only to electrometer valve V, a subminiature pentode operating in the 'inverted' mode. At the same time, another contact on the relay energizes voltage-divider network R_{1-8} which supplies the grid and filament voltages to the valve.

The grid circuit of the valve, together with resistors R_{1-3}, form a bridge network which is balanced for $3I_0 = 0$. Control R_7 sets the voltage on the bridge. The diagonal

In view of the low grid voltage, the ion current from the grid is very small indeed. Thus the few hundred femto-amperes of reverse current drawn by the anode is actually less than the leakage current across the open contacts of the relay, especially after a few months of active use. The inherently low sensitivity of the anode controlled valve is no disadvantage at all in an application where the voltage to be measured is a hundred or more.

The drift in valve parameters during the warm-up period has relatively little effect in the anode controlled mode, and this effect is virtually eliminated by the fact that the interval between the valve being energized and the meter being locked always remains the same. From the point of view of thermal expansion of the electrode structure, the valve operates virtually at ambient temperature. The effect on the characteristic of variations in ambient temperature proves to be inside the limits of measurement accuracy (5 per cent on distance) even in the extremes to be found in Soviet Central Asia.

The anode controlled electrometer is also simpler and cheaper than the valve voltmeter normally used in capacitor storage fault locators.

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Low-Frequency Noise Measurements on the P-Channel M.O.S.T.

By P. W. Fry*

The noise performance of the MTO-1 p-channel m.o.s.t. operating as an amplifier has been investigated over the 12c/s to 20kc/s range.

The problem of measuring noise voltage at low frequencies was overcome by circuits enabling a count to be made of the proportion of cycles over a given amplitude. This is related to the r.m.s. noise voltage by a simple equation.

Results indicate an equivalent input noise voltage following a 1/f law, together with a noise current input which can be ignored except at low frequencies and source impedances greater than 100MΩ.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

M.O.S.T. structures, with their extremely high input resistance, lend themselves admirably to use in electrometers, light and radiation detectors and transducer amplifiers. In many such applications it is essential to know the noise properties of the device at low frequencies. The present study of the Plessey P-channel m.o.s.t. (MTO1) operated in the saturated region covers the frequency range 12c/s to 20kc/s and describes the noise generated in the device in terms of

(a) An equivalent noise generator in series with the gate.

(b) A noise current generator connected between gate and source. (Fig. 1).

The two sources were assumed to be uncorrelated.

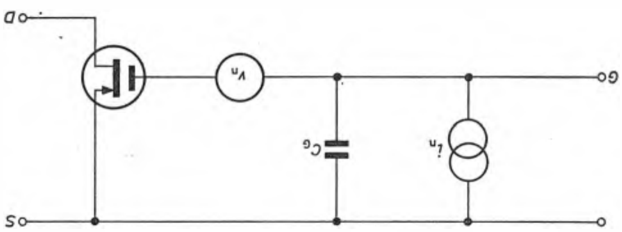

Fig. 1. Equivalent noise sources in the m.o.s.t.

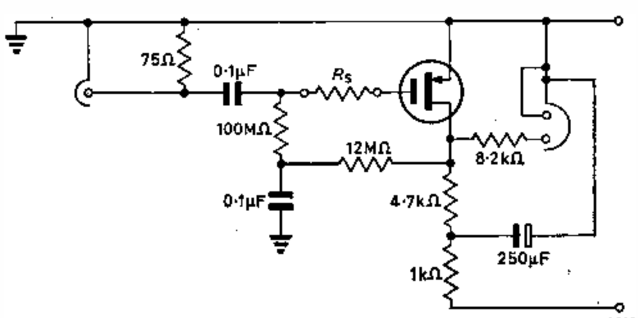

Fig. 2. M.O.S.T. pre-amplifier

Experimental Method

A pre-amplifier (Fig. 2) contained the m.o.s.t., biased in the saturated region by d.c. feedback from drain to gate, so that for all measurements $V_{DS} = V_{GS}$.

The majority of measurements were performed at a drain current of 0·85mA, except for a check on the variation of noise level with this current, when it was reduced to 0·12mA. The source resistor R_S was also included in the pre-amplifier.

The output current from the pre-amplifier was passed into a low input-impedance amplifier circuit of gain 47V/μA, followed by an active filter of the type described by Sutcliffe[1], with switched tuning covering the range 12c/s to 20kc/s in 10 steps (Fig. 3). The filter circuit Q was 100, and the gain of the filter at resonance was also 100.

Since meter readings are impossible at the lower frequencies, the output noise r.m.s. voltage was determined by the following method. One input of a very high gain, differential-input amplifier (Plessey SL701, gain approx. 3 000) was set at a suitable voltage level V. The output from the filter was connected to the other input. The output from this amplifier was limited by diodes to $\pm\frac{1}{2}$V, so that an output pulse was obtained for each cycle of noise input that exceeded the set voltage V. (Fig. 4). These pulses were counted over a period t seconds by a Racal digital frequency meter type SA505. For a narrow bandwidth noise spectrum, the number N of such cycles in a time t is given by Rice[2,3] as:

$$N = ft\,.\exp\left(-V^2/2V_0^2\right)$$

where f is the centre frequency and V_0 is the r.m.s. value of the noise voltage. For a 1 per cent bandwidth the value of t given by Rice[2] as necessary to give V_0 with 4 per cent accuracy is $10^4/f$ seconds. A graph of N/ft against V/V_0

Symbols

v_n = Equivalent m.o.s.t. noise voltage source (volts per root-cycle)
i_n = Equivalent m.o.s.t. noise current source (amperes per root-cycle)
v_N = Total equivalent input noise voltage per root-cycle from all sources
v_G = Noise generator output voltage per root-cycle
v_0 = Output voltage per root-cycle from amplifier/filter system
V_{01} = Output voltage from system with no input from noise generator
V_{02} = Output voltage from system with input from noise generator
V_{DS} = D.C. voltage, drain to source
V_{GS} = D.C. voltage, gate to source
C_G = Total input capacitance of m.o.s.t.
R_S = Source resistance

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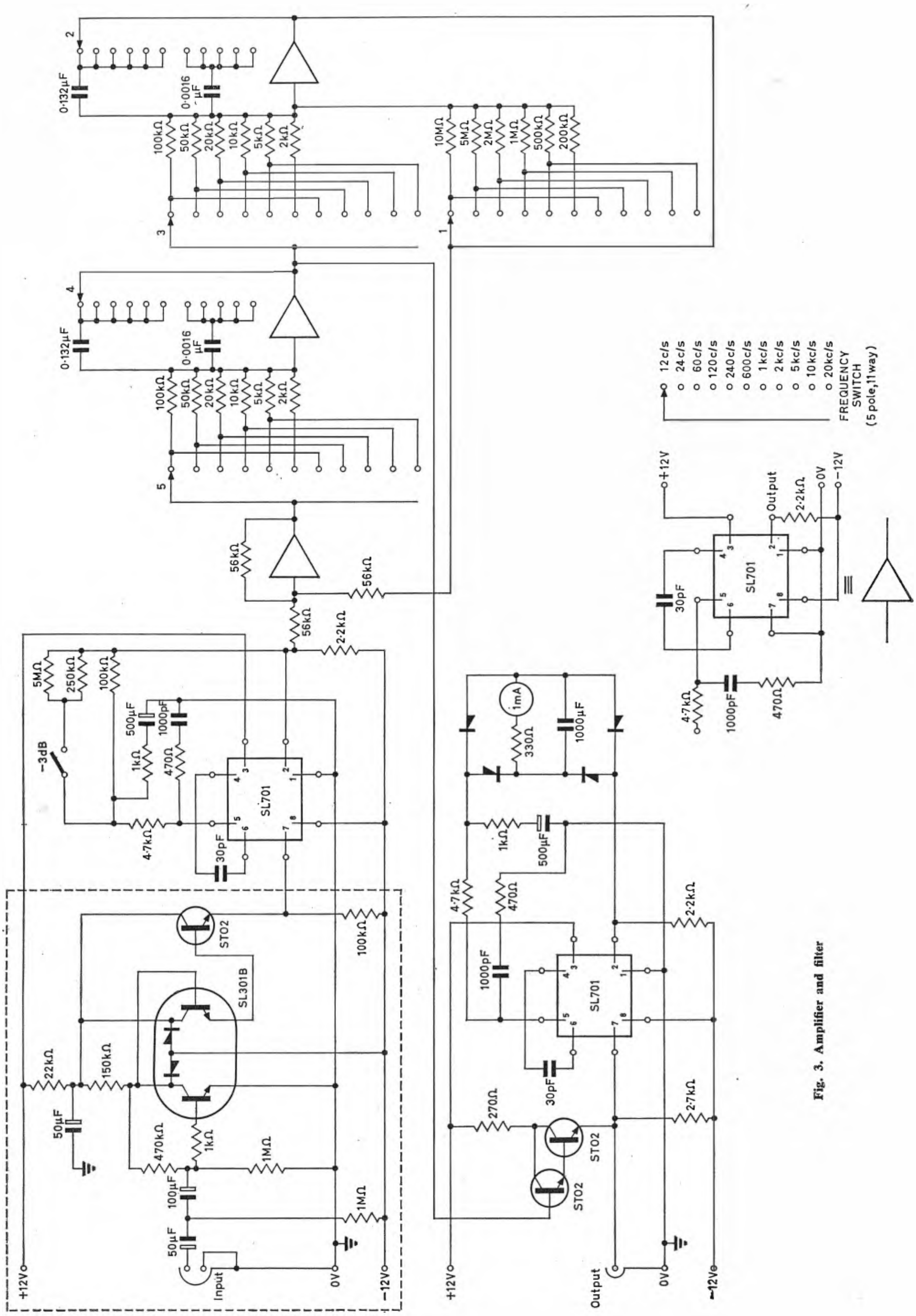


Fig. 3. Amplifier and filter

enabled easy calculation of V_0 from measured values of N. A check on the method was made by measuring the same noise signal using various values of V. The results are shown in Table 1, which indicates that for reliable measurements N/ft should be less than about 70 per cent. Above this value, errors occur due to (a) spurious switching with low-slope waveforms, and (b) low rate of change of

TABLE 1

Verification of the relationship $N = ft \cdot \exp(-V^2/2V_n^2)$ for filtered noise

Frequency 60c/s   Bandwidth 0·60c/s   t = 200sec

V	$\frac{N}{ft}$	$\frac{V}{V_0}$	V_0
0·25	0·956	0·300	0·834
0·50	0·677	0·882	0·567
0·75	0·431	1·298	0·577
1·00	0·2195	1·740	0·575
1·25	0·0946	2·172	0·576
1·50	0·0332	2·610	0·575
1·75	0·0118	2·990	0·586
2·00	0·00208	3·510	0·570

The equivalent total input noise voltage was determined by comparison with the input from a calibrated l.f. noise generator (Rohde and Schwarz Type S.U.F.). The equation linking the output voltages V_{01}, obtained from the equivalent m.o.s.t. input noise plus the source resistor thermal noise, and V_{02}, obtained from the above plus noise generator output to the m.o.s.t. pre-amplifier (v_G) is:

$$v_N = v_G \left(\frac{V_{01}}{(V_{02}^2 - V_{01}^2)^{\frac{1}{2}}} \right)$$

Since the lower limit of the generator noise spectrum was 15c/s, the input noise voltage at 12c/s had to be calculated directly from the output voltage, using the equation:

$$V_{01}^2 = \frac{A_0^2 B v_n^2 \pi}{2}$$

where

B is the experimentally checked bandwidth

A_0 is the system gain at resonance

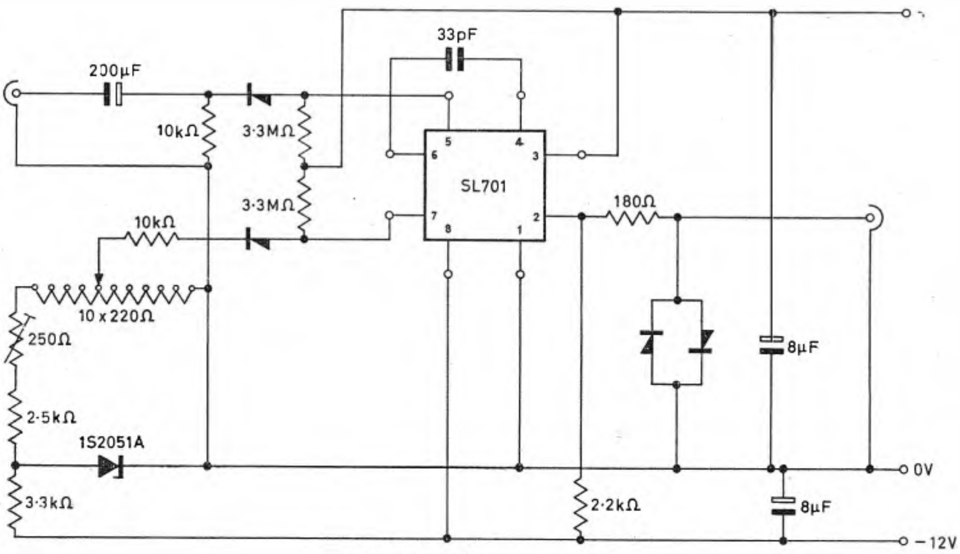


Fig. 4. Amplitude comparator

N/ft with V/V_0. Measurements of output voltage obtained by this method at higher frequencies agree well with meter readings.

Fig. 5. Equivalent noise source voltage plotted against frequency

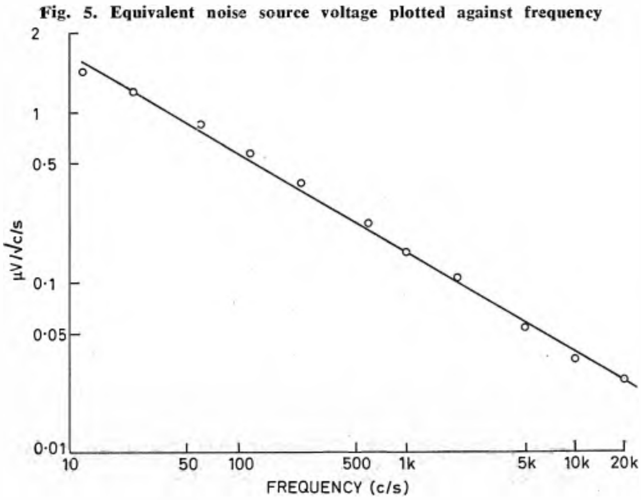
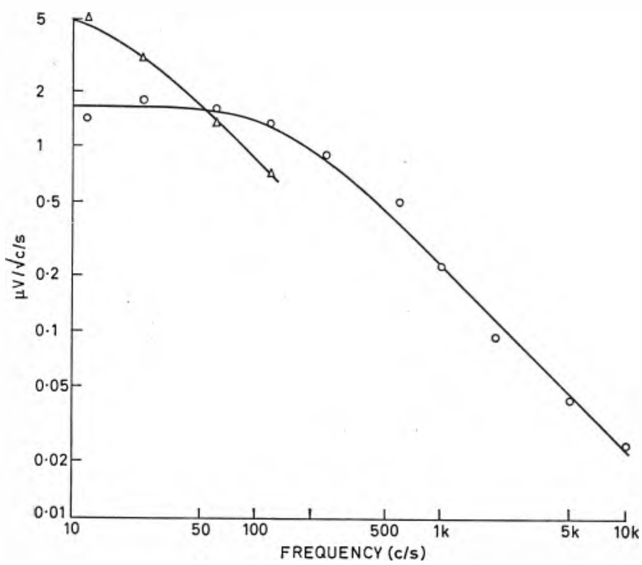


Fig. 6. $(v_N^2 - v_n^2)^{\frac{1}{2}}$ plotted against frequency



v_N is the total equivalent input noise voltage per root-cycle.

This equation is based on the transfer characteristic:

$$v_o/v_N = \frac{A_o}{(1 + (4/B^2)(f - f_o)^2)^{\frac{1}{2}}}$$

where

v_o is the output noise voltage per root-cycle, i.e.:

$$V_{01}^2 = \int_0^\infty v_o^2\, df$$

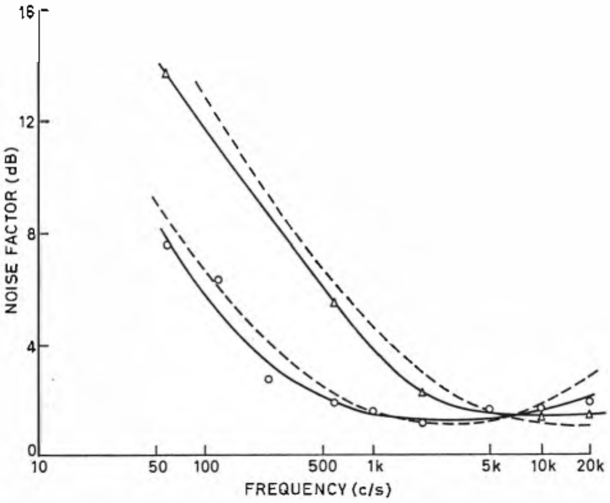


Fig. 7. Noise figures plotted against frequency, with intermediate source resistances

Measurement of the equivalent noise voltage generator v_n was made by making R_S equal to zero.

The noise current measurements were made with R_S equal to 100MΩ, and to 1 000MΩ.

From the formula:

$$v_N^2 = v_n^2 + 4kTR_S + i_n^2R_S^2$$

a value for i_n can be calculated.

Checks on the measurements were made using 1MΩ and 5MΩ source resistances. The noise figure thus measured as $v_N^2/4kTR_S$ should equal the value $1 + \dfrac{v_n^2 + i_n^2R_S^2}{4kTR_S}$ computed from the values of v_n and i_n at the same frequency.

Results

The noise voltage is seen from Fig. 5 to follow the familiar $v_n^2 = \text{const.}/f$ law associated with interface-state population fluctuations[4], over the whole range studied. These results agree well, on extrapolation, with Bozic's[5] measurements at higher frequencies on N-channel devices, showing rather lower noise levels for these P-channel devices.

The noise current source was found to be immeasurably small at other than the lowest frequencies (up to ~120c/s) since measurements depended on generating a voltage across a source resistor of upwards of 100MΩ: the capacitances in the input circuit shunt this resistor at higher frequencies. As can be seen from Fig. 6, a plot of $(v_N^2 - v_n^2)^{\frac{1}{2}}$ against f can adequately be represented as a fixed voltage attenuated by a factor:

$$\frac{1}{(1 + 4\pi^2f^2C_G^2R_S^2)^{\frac{1}{2}}}$$

where

$C_G = 11\text{pF}$.

Now simple theory gives:

$$v_N^2 - v_n^2 = (4kTR_S + i_n^2R_S^2)\left(\frac{1}{1 + 4\pi^2f^2C_G^2R_S^2}\right)$$

The input capacitance of the m.o.s.t. pre-amplifier has been calculated as 12·9pF, so that i_n can reasonably be assumed to be constant over these low frequencies. Values of i_n were calculated from the two experiments:

$$R_S = 100\text{M}\Omega: \; i_n = 11{\cdot}3 \times 10^{-15}\text{A}/\sqrt{(\text{c/s})}$$

$$R_S = 1000\text{M}\Omega: \; i_n = 4{\cdot}5 \times 10^{-15}\text{A}/\sqrt{(\text{c/s})}$$

Since the former result involves measuring a quite small increase over the noise due to R_S itself, the latter result is to be preferred.

Allowing for the m.o.s.t. input capacitance, the apparent noise figure with any source resistance R_S is given by:

$$\text{N.F.} = 1 + \frac{i_n^2R_S^2 + v_n^2(1 + 4\pi^2f^2C_G^2R_S^2)}{4kTR_S}$$

Since

$v_n^2 = \text{const.}/f$, and ignoring i_nR_S:

$$\text{N.F.} = 1 + \frac{\text{const.}}{4kTR_S}((1/f) + 4\pi^2fC_G^2R_S^2)$$

which reaches a minimum where:

$$f = \frac{1}{2\pi C_GR_S}$$

If $C_G = 11\text{pF}$

when

$$R_S = 1\text{M}\Omega, \; \frac{1}{2\pi C_GR_S} = 14{\cdot}5\text{kc/s}$$

when

$$R_S = 5\text{M}\Omega, \quad ,, \quad = 2{\cdot}9\text{kc/s}$$

These frequencies fit well the minima in the noise figure curves of Fig. 7. A plot of noise figure against frequency using a typical value of v_n from the graph of Fig. 5, with $i_n = 0$ and $C = 11$pF, is included in Fig. 7.

Reduction of I_{DS} to 0·12mA made no difference to the above values, within experimental error.

Conclusions

In amplifier applications where source resistances of 10MΩ downwards are typical, the current noise source can be entirely neglected. The voltage noise source follows the $1/f$ law over the whole range considered, and has a typical value of 2μV/√(c/s) at 12c/s.

In electrometer applications a current noise source of the order of 10^{-15} to 10^{-14}A/√(c/s) must be taken into consideration.

At frequencies above about 100c/s this current may be neglected, since it is shunted by the input capacitance of the device.

Acknowledgments

The author's thanks are due to the Plessey Co. Ltd, and to the Ministry of Defence (Naval) for permission to publish this article and to Mr. J. Hine for help in the practical work.

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A Transistorized Impedance Dialling System

By H. N. Wroe*, B.Sc.(Eng.), Ph.D., A.M.I.E.E.

Telephone lines serving h.t. power stations frequently require protection by the use of isolating transformers and normal d.c. dialling techniques are then precluded. A signalling system utilizing semiconductors, which provides a practical and economic solution to the problem, is described. Its basis is that by incorporation of the line impedance at the telephone exchange in the circuit of a low-frequency oscillator changes in the line terminating impedance at the h.t. station can permit or inhibit oscillation. The oscillator drives an impulse-correcting detector and the combination translates the changes in line impedance into normal d.c. loop signals.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

IMPEDANCE dialling was developed to permit dialling from a h.t. power station over a telephone line in which, for safety reasons, the insertion of an isolating transformer is necessary. Its principle is that by incorporating the line impedance presented to the exchange into the circuit of an oscillator located at the latter, variation of the impedance terminating the line at the h.t. station can determine whether the oscillator is active or quiescent. In this way the addition of a few very simple passive components to the subscribers' telephone instrument is all that is necessary at the h.t. station for signalling over the protected line[1].

The original impedance dialling system utilized thermionic valves and relays. For reasons of reliability the number of valves had to be kept to the minimum and, in fact, a single double-triode was employed.

Similar considerations do not apply to the use of transistors. Design is therefore less restricted, impulse correction can readily be included, and wider operating limits can accordingly be achieved. Except at two points in the circuit (forward d.c. impulsing and application of ringing current to line), where potentials beyond the range of normally available transistors are encountered, relays can be eliminated.

With a view to securing these advantages a transistorized impedance dialling system operating on a nominal frequency of 120c/s has been developed. The circuit of the system is shown in Fig. 1, and an outline of its operation is as follows:

(A) When the line is idle its impedance at the exchange is high and reactive. The oscillator composed of L_2, C_2, VT_1, oscillates and its output, rectified in the full wave bridge, bottoms VT_2. VT_6 is accordingly cut-off and permits a battery supply to the bases of VT_4 and VT_5, which are therefore bottomed and hold relay A operated. A keeps the line forward to the exchange equipment disconnected, bottoms VT_3, which cuts off VT_8, and incidentally assists in holding VT_5 bottomed.

(B) On the line being seized at the h.t. station its impedance becomes low and resistive. The consequent load on the oscillator causes oscillations to cease. VT_2 therefore cuts off, holding VT_3 and VT_5 bottomed and bottoming VT_6, which cuts off VT_4. A accordingly releases and loops the line forward. When a dial impulse arrives the line impedance becomes high, producing oscillations which bottom VT_2. VT_3 and VT_5 cut off, the former bottoming VT_4 and VT_8 and charging C_{11}. VT_6 remains bottomed because the charge on C_9 cannot leak away during the period of a dial impulse. At the end of the impulse the reduction in line impedance prohibits oscillation and VT_2 cuts off, bottoming VT_3 and VT_5. With VT_5 bottomed, and VT_4 also maintained bottomed by the discharge from C_{11}, A operates disconnecting the line forward and providing an alternative bottoming of VT_3 and VT_5 to ensure that the arrival of a further dial impulse cannot interfere with the regenerated impulse being sent forward by A. At the end of an interval, which can be adjusted by variation of RV_1, C_{11} becomes discharged and A releases. The circuit is then ready to deal with a succeeding dial impulse. VT_8 remains bottomed throughout the impulse train because of the charge stored on C_6, and provides a low resistance loop for impulsing forward in addition to eliminating cross-fire to the line winding of L_2.

When the h.t. station clears conditions become as described in (A) above.

(C) A call incoming to the h.t. station finds the circuit in the state described in (A). The normal ringing current, which is unbalanced with respect to earth, bottoms VT_7 causing relay R to operate. R therefore follows the interrupted ringing sequence and at each operation applies ringing current to line. At the first ringing interruption following the answer by the h.t. station the circuit assumes the condition of (B) above and the incoming ringing is tripped in the normal manner.

R is prevented from responding to forward line surges during dialling by the bottoming of VT_6, which short-circuits it.

(D) Adoption of a low frequency for the oscillator not only increases the permissible line length; but also helps to mitigate voice-operation possibilities. The use of the voltage-limiter pairs $MR_1 MR_3$, $MR_2 MR_4$, deals with possible trouble from excessively high voice signals, and voice-operation is not a problem.

The MR_5-MR_{11} and MR_6-MR_{12} chains protect VT_8 against high voltage surges between lines.

The most convenient supply voltage is 50V—a value too high for cheap transistors. This situation was turned to advantage by incorporating a voltage stabilizer composed of R_9, MRZ_5, MRZ_6, to produce a 25V supply. Voltage stabilization eliminates the effect of supply voltage variations on the output impulse ratio.

It is difficult, without affecting impulsing performance, so to smooth a low-frequency input to a transistor that the latter is bottomed throughout a half-cycle. To meet this MRZ_1, MRZ_2, MRZ_3 have been introduced to provide suitable voltage thresholds. MRZ_4 ensures that slight departures of VT_6 from bottoming do not affect VT_4 and VT_5.

Variation of oscillator gain is obtained by adjustment of RV_2.

The tuned circuits L_1-C_1, L_3-C_3 ensure that while the

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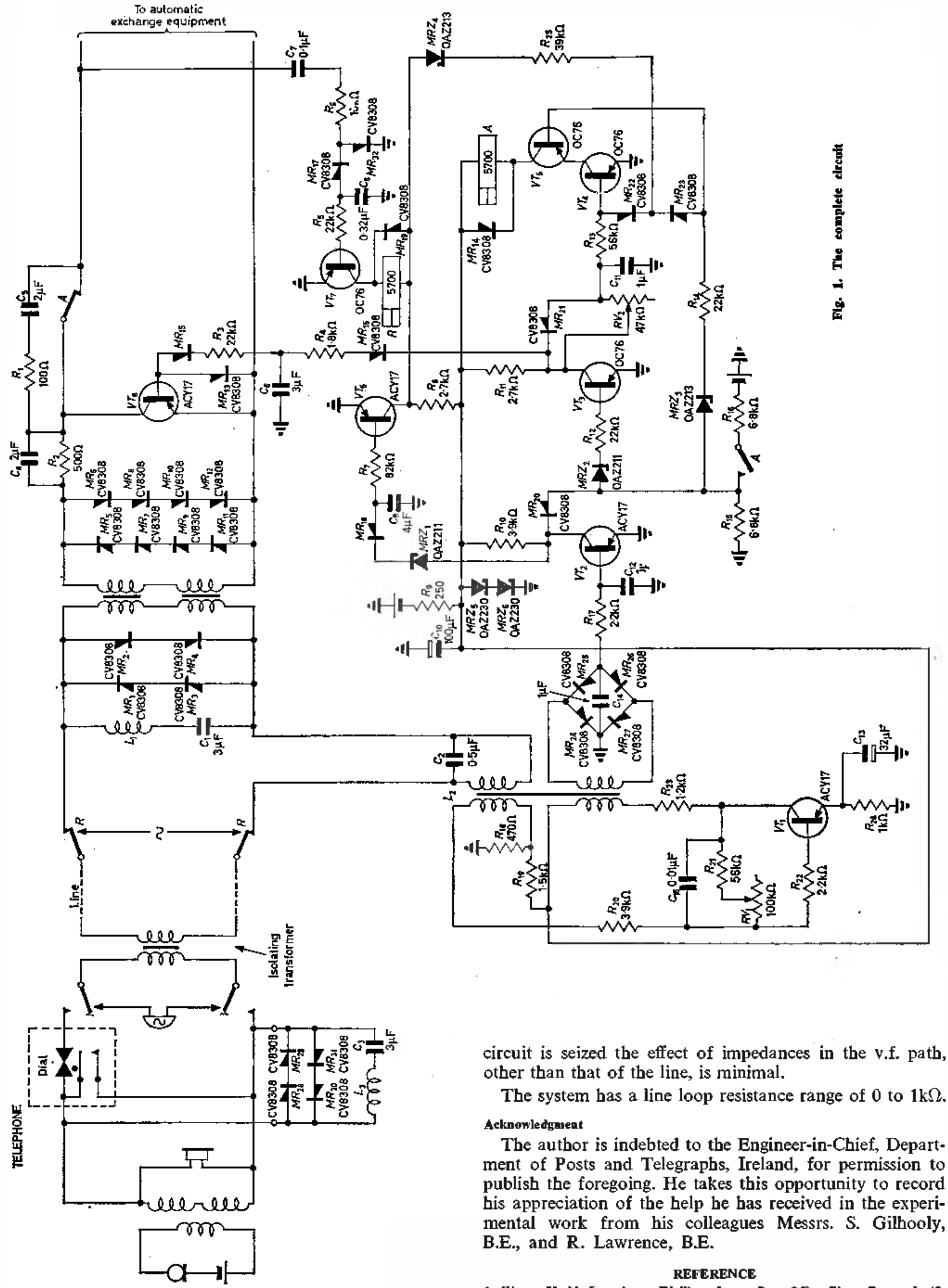

Fig. 1. The complete circuit

circuit is seized the effect of impedances in the v.f. path, other than that of the line, is minimal.

The system has a line loop resistance range of 0 to 1kΩ.

Acknowledgment

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A V.H.F. Power Modulator

By M. B. Clark*, B.Sc.(Eng.)

A modulator circuit is devised to give broadband output signals within the frequency range of 66 to 184Mc/s at 2·5mW and 50Ω impedance.

These bandwidths are achieved with a level ouput to within ±1dB and a typical insertion loss of 8dB. The conventional ring modulator would not operate at the required power level and a modified circuit is used. The minimum carrier level is derived for a given signal level ($V_c = V_s$ and $I_c = 2I_s$). This is proved in practice. For the 25mW input signal the minimum carrier power is 50mW, for the correct modulation.

The ratings of the diodes are related to the signal and carrier levels and the maximum signal levels are calculated for two types of diode. Practical details of the modulator transformers and diodes are discussed.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

THE modulator was to be capable of working with either of two carrier frequencies, thus giving a wide band of output frequencies.

Input signal: 19 to 43Mc/s. 25mW 50Ω.

Carrier drive: 109 and 140Mc/s.

Output signal: 66 to 184Mc/s. 50Ω.

A double balanced output was required with unwanted spurious signals 20dB down. It was desirable for the insertion loss to be low and the output level over the band to be within 1dB. Input and output impedances were 50Ω.

Modulator Circuit

A modified ring modulator circuit[1] is used to achieve the required power level. It is necessary to employ separate switching paths and resistive carrier drives. Thus the input transformer has two secondary windings (Fig. 1) and the carrier drives are fed through the resistors R_2. The configuration reduces to a conventional ring modulator when $R_2 = 0$ (Fig. 2) in which case only one secondary winding is required. The operation of the modified ring modulator is similar to the conventional ring modulator, but has the advantage that the signal and carrier voltages are not limited by the forward voltage of the diodes as explained below.

Theoretical Operating Levels

The modified ring modulator is examined to establish the following:

- the minimum carrier level
- the value of R_2
- the ratings of the diodes.

It is necessary to consider the condition when both carrier and signal, voltages and currents, are at instantaneous peak value (Fig. 3, V_c, I_c, V_s, I_s).

The following assumptions are made:

V_s is in phase with I_s

V_c is in phase with I_c

the voltage V_F across the diodes is constant for all values of forward current

for the diodes $Z_f = 0$.

$Z_b = \infty$

The peak conditions shown are for MR_1 and MR_2 'on' so that MR_3 and MR_4 are 'off'.

The characteristic impedance of the modulator need not be Z_o, for transformers with turns ratio n the characteristic impedance is $Z = n^2 Z_o$. Thus for the peak condition shown $V_s/I_s = Z$.

To determine the carrier voltage consider the voltage V_{bb}

$$V_{bb} = V_{bk} + V_{ka} + V_{ab}$$
$$= +(V_s/2) + V_F - (V_s/2)$$
$$= +V_F \quad \ldots\ldots (1)$$

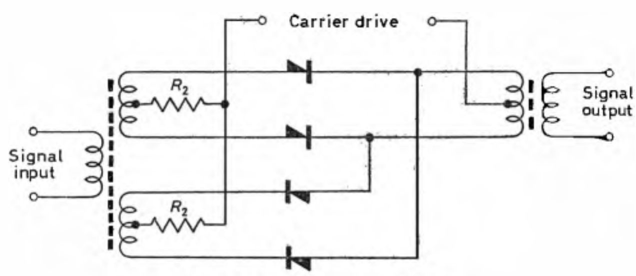
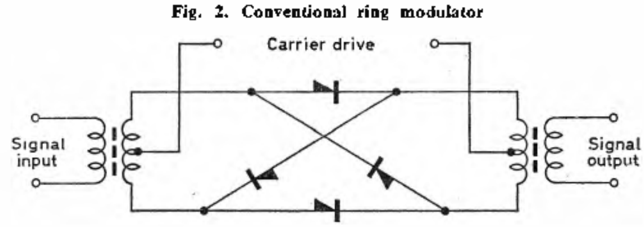


Fig. 1. Modified ring modulator

Fig. 2. Conventional ring modulator



SYMBOLS

V_c	=	Instantaneous peak carrier voltage.
I_c	=	Instantaneous peak carrier current.
V_s	=	Instantaneous peak signal voltage
I_s	=	Instantaneous peak signal current at impedance Z.
Z_o	=	50Ω = Characteristic impedance of system.
Z	=	n^2Z_o = Characteristic impedance within modulator.
n	=	Turns ratio of transformer.
V_F	=	Diodes forward voltage drop.
Z_f	=	Diodes forward impedance.
Z_b	=	Diodes reverse impedance.
$\Phi(t)$	=	Modulating function.
A	=	Amplitude of $\Phi(t)$.

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Hence the carrier voltage

$$V_c = V_{hl} = I_c R_2 + V_F \quad \ldots\ldots\ldots\ldots \quad (2)$$

However, for high-power operation V_c is large and $V_F \ll V_c$, hence

$$V_c = I_c R_2 = E_c \frac{R_2}{R_1 + R_2} \quad \ldots\ldots\ldots\ldots \quad (3)$$

and so V_c can be adjusted by the value of R_2 for a given value of I_c. In the case of the conventional ring modulator $R_2 = 0$ and therefore V_c cannot exceed V_F.

Considering the currents through the diodes MR_1 and MR_2. For MR_2 to remain 'on', I_c must be such that,

$$I_c/2 > I_s \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Thus defining the minimum carrier current for a given signal.

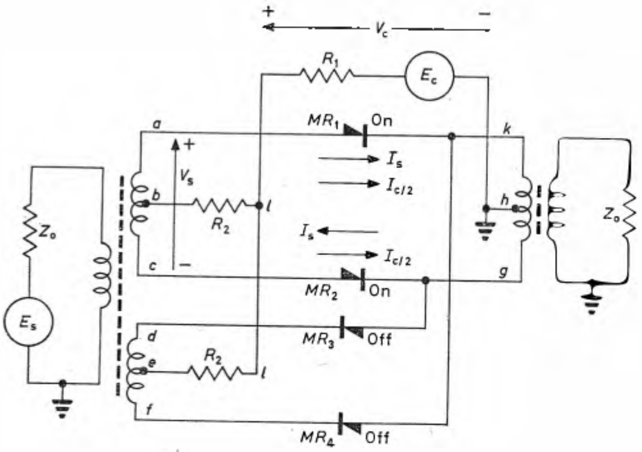

Fig. 3. Modified ring modulator—instantaneous peak conditions

The current through MR_1

$$= (I_c/2) + I_s > 2I_s \quad \ldots\ldots\ldots\ldots \quad (5)$$

defines the 'minimum' current rating of the diodes.

Consider the voltages across the diodes MR_3 and MR_4. For MR_3 and MR_4 to remain 'off' $V_{dg} < V_F$ and $V_{fk} < V_F$, i.e. the diodes must not conduct.

$$\begin{aligned} V_{dg} &= V_{de} + V_{el} + V_{lh} + V_{hg} \\ &= -(V_s/2) + 0 - V_c - (V_s/2) \\ &= -V_c - V_s \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6) \end{aligned}$$

and

$$\begin{aligned} V_{fk} &= V_{fe} + V_{el} + V_{lh} + V_{hk} \\ &= +(V_s/2) + 0 - V_c + (V_s/2) \\ &= -V_c + V_s \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7) \end{aligned}$$

Clearly for MR_4 to remain 'off' and $V_{fk} < V_F$, then,

$$-V_c + V_s < V_F \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

which is certainly satisfied if

$$V_c = V_s \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

Thus a lower bound for the peak carrier voltage is defined for a given signal level.

The reverse voltage rating of the diodes is defined by the peak voltage V_{dg} in equation (6) and when $V_c = V_s$

$$|V_{dg}| = V_c + V_s = 2V_s \quad \ldots\ldots\ldots\ldots \quad (10)$$

i.e. breakdown voltage must exceed $V_c + V_s$.

The minimum diode ratings for both current and voltage are now defined together with the desired carrier level.

With regard to R_2 the minimum required current $I_c \geqslant 2I_s$ and the minimum required carrier voltage $V_c \geqslant V_s$ gives a value of R_2.

$$R_2 = (V_c/I_c) = (V_s/2I_s) = (1/2)\,Z = (1/2)n^2 Z_o \quad \ldots \quad (11)$$

If R_2 and V_c are specified as above the carrier current I_c will also be correct.

For the conventional ring modulator $R_2 = 0$ and $V_c = V_F$.

Hence for the condition (8) it follows that V_s cannot exceed $2V_F$ without turning MR_4 'on' and shorting the signal. This is a fundamental limit on the signal level.

Theoretical Insertion Loss

The insertion loss of the modulator depends on the modulating function $\Phi(t)$. For the ring modulator this is

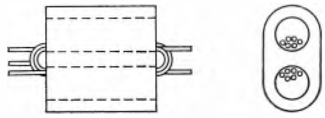

Fig. 4. Modulator transformer

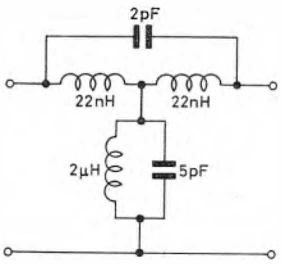

Fig. 5. Transformer equivalent circuit 2 : 2 + 2 turns referred to 2 turn primary

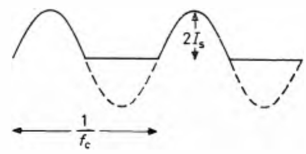

Fig. 6. Diode current waveform

a square wave of amplitude $A = 1/L$ where L is the insertion loss of the total signal[1].

$$A = \frac{Z(Z_b - Z_f)}{(Z + Z_b)(Z + Z_f)}.$$

For diodes with infinite reverse impedance and zero forward impedance $Z_f = 0$ $Z_b = \infty$ then $A = L = 1 = 0$dB.

When this is so, the insertion loss of the first order sideband is 3·9dB (Tucker[1]). But the insertion loss for the total output of all upper sidebands (or all lower sidebands) corresponding to all harmonics of the modulating frequency is 3·0dB.

Transformers

The transformers were required to have a bandwidth of 10 to 180Mc/s and a low insertion loss. To obtain the bandwidth, low leakage inductance and high shunt impedance were required, over the whole band of frequencies. The latter necessitated a high shunt inductance and a low winding capacitance.

Successful transformers were constructed using ferrite 'two hole tubes', placing the winding on the central core as shown in Fig. 4. A two-turn primary winding was required to reduce the magnetizing current and provide sufficient shunt inductance. By using a 2:1 turns ratio the impedance within the modulator was raised to 200Ω.

$$Z = n^2 Z_o = 200\Omega.$$

Thus the effect of the diode impedance Z_f was reduced ($Z_f \ll Z$) and the secondary windings consisted of four turns centre tapped.

The value of R_2 was therefore fixed by equation (11).

$$R_2 = 1/2\,Z = 100\Omega$$

A transformer bandwidth of greater than 10 to 180Mc/s was achieved with an insertion loss of approximate 1dB. A typical equivalent circuit for the reactive components is given in Fig. 5. The shunt arm is resonant at 50Mc/s which is well within the band ensuring a high impedance. The total series inductance could be reduced from 44nH to 22nH by careful winding layout. The 2pF capacitance was of no practical importance.

Diodes

The requirements for the diodes were as follows:

sufficient voltage and current ratings,

high speed to switch at 140Mc/s,

low loss for $Z_f \ll 200\Omega$,

low capacitance $Z_b \gg 200\Omega$.

Of the available diodes the only two which were found to be suitable were the FD.700 and the M54 CAY. The FD.700 showed some stored charge at 140Mc/s, but was fast enough for satisfactory modulation. The quoted reverse recovery time was 0·7nsec. The M54 CAY diode was faster but the ratings were lower.

The ratings of the diodes determine the maximum signal and carrier levels. For a given signal the peak condition is 2 I_s and 2 V_s from equations (4) and (10). It was necessary to correlate the peak condition 2 I_s to the average rating of the diodes. The average current for the diode is given by the alternate half cycles of the sine wave of amplitude 2 I_s (Fig. 6).

The average current = 0·32 (2 I_s).

However the peak voltage was the limiting factor for both types of diodes at the impedance level 200Ω. The maximum signal input power is calculated in Table 1 for the condition:

2 V_s = Peak voltage rating of diode.

Maximum power $P_m = (V_s/\sqrt{2})^2\, 1/R$

where $R = 200\Omega$.

TABLE 1

DIODE	PEAK VOLTAGE RATING (V)	MAX. AV. CURRENT RATING (mA)	MAX. INPUT POWER P_m (mW)
FD 700	20	50	250
M54 CAY	10	30	62

Modulator Performance

The modulator operated correctly with both types of diode up to the limits of signal input power specified. This was checked by plotting the output power in a first order sideband, against the voltage into a 50Ω load with the modulator removed. The relationship was linear.

If the carrier level was reduced the modulator became over-driven before the above limit was reached and the output was limited.

However for the required specification, the input signal power was only 25mW thus the voltage,

$$V_s = \sqrt{2}\ (200 \cdot 2{\cdot}5/1\,000)^{\frac{1}{2}}$$

$$= \sqrt{10} = 3{\cdot}16\text{V}.$$

and hence the required peak carrier voltage from equation (9)

$$V_c = V_s = 3{\cdot}16\text{V}.$$

In practice the carrier level was measured by a valve-voltmeter to the r.m.s. value $1/\sqrt{2}\ V_s = 2{\cdot}2$V, and the carrier drive required was 2·2V r.m.s. into $R_2 = 100\Omega$, giving a power of 50mW.

With 25mW input signal and the minimum carrier level 2·2V r.m.s. and using FD.700 diodes, the insertion loss to each sideband was typically 8dB compared to the theoretical 3·9dB. The 8dB loss could be reduced somewhat by using higher carrier drive, thus operating the diodes at higher current. The insertion loss was measured over the frequency band for both 109Mc/s and 140Mc/s carrier frequencies. Variations were within ±1dB.

An idea of the modulation input impedance was obtained by measuring the input voltage with a 50Ω load on the output. The input impedance was approximately 80 to 90Ω over the band.

It was of interest to note that the conventional ring modulator was linear up to an input level of 2·5mW. This corresponded to V_s = 1·0V, which agrees with the relation $V_s \simeq 2\ V_F$. However, the insertion loss was lower being 6dB, due to the larger carrier currents. In fact, it was difficult to control the current with $R_2 = 0$ and consequently the diodes could be inadvertently destroyed.

Acknowledgments

The author is grateful for permission to publish this article given by the Director, Royal Aircraft Establishment, Farnborough, and by the Directors of the Plessey Company Limited.

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Remote Control for Inter-Colliery Conveyors

As a result of the recent installation of a G.E.C. Electronics remote control system on conveyors carrying coal between Haigh and Woolley collieries, near Barnsley, the Yorkshire Division of the National Coal Board has achieved a valuable saving in manpower.

The two collieries are linked by three roller-driven belt conveyors installed in series and covering a total distance of one mile. Up to 3 000 tons of coal a day are conveyed from Haigh to a preparation plant at Woolley.

Before G.E.C. Electronics' 'Televeyor' telemetric control and monitoring system was introduced, three men were required during each shift to control and supervise the points at which coal is transferred from one conveyor to the next and discharged into the preparation plant. Now, all three points are supervised remotely by one operator from a control console at Woolley.

Seven conditions at each point are monitored continuously by the G.E.C. equipment and displayed on a series of illuminated indicators on the console. A telephone system, integrated with the control equipment, enables the operator to communicate with the conveyor patrolman if a fault arises at any point. Local control consoles displaying the various alarms are aslo provided at each point and the system is designed to permit manned operation of any or all of the conveyors while retaining automatic monitoring of safety devices.

The G.E.C. 'Televeyor' system is specifically designed for use in damp and dust-laden environments and is immune against the vibration of adjacent machinery. It consists of frequency division multiplex equipment and time division scanning equipment using six-core communication cables to carry control and monitoring signals back to the central control console from each of the three points under supervision.

One of the prime purposes of 'Televeyor' is to monitor fire detection instruments, consisting of 'Minerva' alarms which distinguish between 'live' smoke and coal dust; transducers which record the temperature of the belt as it passes over the driving drums; and 'Graviner' bimetallic thermostats fitted to the bearings of the main drum at each drive head.

In addition the system monitors pre-set centrifugal switches which are actuated if the belt slips on the rollers over the driving drum. Similar supervision is exercised over micro-switches which are triggered if the belt tears or frays or if the edge of the belt exceeds a pre-determined sideways movement.

Excessive accumulations of coal in the chutes at the transfer and discharge points are detected by ultrasonic beams. If a chute becomes blocked, the beams are interrupted and a signal is transmitted over the telemetry link to the local and main control consoles. G.E.C. chose an ultrasonic rather than optical chute blockage detector since the latter would have been unable to differentiate between coal and coal dust.

Low Drift Design of Differential D.C. Amplifiers

By S. K. Bhola*, M.Sc.(Hons), and R. H. Murphy*, B.Sc.(Eng.)

The performance of d.c. amplifiers is limited by their drift with temperature. Alternative approaches to reducing this drift are by chopper stabilization and by employing emitter coupled, differential, designs. The former has limitations as regards to the offset voltage and temperature coefficient of offset voltage, in addition to noise and bandwidth restrictions whereas the latter demand careful attention to matching problems. A better understanding of the factors responsible for drift has resulted in the development of devices and circuit techniques for differential amplifiers which reduce the drift below 1μV/°C and offer small size and noise free operation. An attempt has been made to demonstrate the low drift capabilities of these amplifiers using newly available devices and techniques.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

IN practice there are many occasions when one needs a d.c. amplifier because of very low frequency signals being involved or where the d.c. component of an a.c. signal is to be retained or for purely d.c. signals. Nevertheless, the performance of d.c. amplifiers for such applications depends very much on the level of the input signals involved. As the signal level decreases, the complexity and cost of providing a suitable d.c. amplifier also increases. This is due to the fact that the operating conditions of a d.c. amplifier change with time as well as with temperature. This causes the output of a d.c. amplifier to drift away from zero for no signal conditions. When a signal is being amplified, drift changes the output from its true value, thus acting as an error signal. It is this drift, often referred to as equivalent to an error signal, as referred to the input, which limits the level of the signal to be amplified or the accuracy of the signal.

Another factor, which, of course, is not so serious, is the noise generated in the transistors and resistors. The noise voltage generated in the amplifier also limits the performance in the same way except that this tends to remain constant.

The drift is due to two reasons:

(1) *Time*

This is a slow process which occurs due to ageing of the device and changes in the effective parameters of the device. Usually the techniques used for drift reduction due to temperature also take care of drift due to time.

(2) *Temperature*

This is a more important parameter causing drift. The main factors giving rise to drift are:

(a) Variation of base-collector leakage current I_{CBO} with temperature.

(b) Variation of d.c. gain h_{FE} with temperature.

(c) Variation of forward base-emitter voltage drop V_{BE} with temperature.

The leakage current I_{CBO} as a function of temperature is given by the following equation:

$$I_{CBO} = A \,.\, 10^{(T-T_m)/30°C}$$

where I_{CBO} = Base collector leakage current at a temperature T°K

A = Measured value of I_{CBO} at T_m°K

The initial value of I_{CBO} is dependent entirely upon the device. In silicon planar epitaxial devices, it can be as low as 2×10^{-9}A at 25°C and can be neglected for all practical purposes.

The d.c. gain decreases with temperature and the component of drift due to this factor is given by:

$$\frac{\Delta h_{FE}}{h_{FE}} (I_{CBO} + I_B) \,.\, \Delta T$$

The contribution of this component can be reduced by decreasing the working currents. It is true that h_{FE} decreases with current, but the reduction of h_{FE} by a factor of two might be accompanied by a hundredfold decrease of I_B. Modern devices like the 2N930 have a d.c. gain of 100 to 500 at 10μA of collector current. In practice, it is the leakage current and noise which set a limit to the minimum operating base currents. D.C. gain does not pose a big problem because it can be stabilized by negative feedback.

The base emitter voltage of a transistor is approximately given by:

$$V_{BE} = V_{BE}' + I_b r_b + I_e r_e$$

where V_{BE} = Actual base-emitter voltage

V_{BE}' = Ideal base-emitter voltage

I_b = Base current

r_b = Base resistance

I_e = Emitter current

r_e = Emitter resistance

The ideal base-emitter voltage V_{BE}' has a negative temperature coefficient of approximately −2·5mV/°C. The actual base-emitter voltage V_{BE} is due, in addition to V_{BE}', to the bulk resistance in series with the base-emitter diode and this bulk resistance has a positive temperature coefficient. This resistance is not closely defined and therefore the actual V_{BE} and its temperature coefficient are also not well defined. The temperature coefficient of V_{BE} is also dependent upon current, being close to the ideal at low currents. With the planar method of transistor manufacture, resistivities are more uniform and consequently V_{BE} and its temperature coefficient are more uniform.

In low level differential amplifiers, the residual drift has been found to be due to the difference in the temperature coefficients of base-emitter voltages of two transistors in the input differential stage. A study of this effect and its dependence upon transistor parameters has led to designs having very high performance.

The collector current in a transistor is given by:

$$I_C = AT^3 [\exp(-qE_g/kT)] [\exp(qV_{BE}/kT) - 1] \quad \dots\dots\dots\dots (1)$$

* *Transitron Electronic Ltd.*

where A = Constant for the semiconductor device
T = Absolute temperature
q = Electronic charge
k = Boltzmann's constant
E_g = Energy gap between the valence and conduction bands in the semiconductor material
V_{BE} = Base-emitter voltage.

In silicon planar transistors, leakage current can usually be neglected, hence equation (1) reduces to:

$$I_C = AT^3 \exp [(V_{BE} - E_g)\, q/kT]$$

or:

$$V_{BE} = kT/q\, [\ln I_C - 3 \ln TA] + E_g \quad \ldots\ldots\ldots (2)$$

Differentiating the above equation with respect to temperature gives:

$$(\delta V_{BE})/\delta T = k/q\, [\ln I_C - 3 \ln AT] + E_g - (3k/q) \quad \ldots\ldots (3)$$

If $(\delta V_{BE1})/\delta T$ and $(\delta V_{BE2})/\delta T$ are the temperature coefficients related to the two transistors in the input stage, then: : : :

$$(\delta V_{BE1})/\delta T = (k/q)\,(\ln I_{C1} - 3 \ln AT_1) + E_1 - (3k/q) \quad \ldots\ldots\ldots\ldots (4)$$

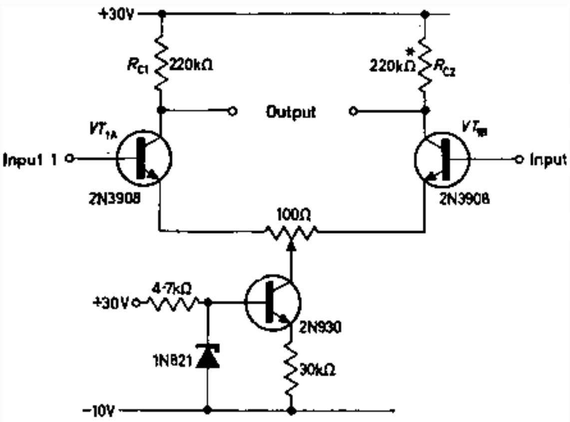

Fig. 1. Circuit for driving high input impedance systems
** See text*

$$(\delta V_{BE2})/\delta T = k/q\, (\ln I_{C2} - 3 \ln AT_2) + E_g - 3k/q \quad \ldots (5)$$

Now the differential temperature coefficient of base emitter voltage is:

$$(\delta V_{BE12})/\delta T = (\delta V_{BE1})/\delta V - (\delta V_{BE2})/\delta T$$
$$= k/q\, [(\ln I_{C1} - \ln I_{C2}) - 3(\ln T_1 - \ln T_2)]$$

If the two transistors are maintained at very nearly the same temperature, i.e. if good thermal matching is employed, then $T_1 = T_2 = T$ and:

$$\delta V_{BE2}/\delta T = (k/q) \ln (I_{C1}/I_{C2}) \quad \ldots\ldots\ldots\ldots (6)$$

This equation indicates that the net differential temperature coefficient can be reduced to zero by proper adjustment of the two collector currents. Assuming this adjustment is possible with a high degree of accuracy, the drift then depends upon the constancy of the two collector currents, which tend to change with temperature as well as with load. Thus the minimum drift achievable in practice is a function of the possible closeness of thermal and collector current matching.

Fig. 1 shows one of the practical circuits suitable for driving high input impedance systems such as digital voltmeters. The transistors are fed from a constant current source comprising a current stabilized transistor in the common emitter supply. The current stability of this transistor with temperature depends upon temperature coefficient of Zener voltage and that of the base-emitter voltage. With a Zener of 0·01 per cent/°C temperature coefficient (0·6mV/°C for 6V Zener), this yields a change of 0·05 per cent for the worst case when both temperature coefficients have the same sign. With component values shown in Fig. 1, the amplifier typically has a gain of 460, which can be set to any desired value by employing negative feedback, and drifts $-3{\cdot}5\mu$V/°C from 21°C to 38°C.

Now for initial balance (i.e. zero differential output for no input) the voltage drops across the collector resistors are equal, so:

$$R_{C1} \,.\, I_{C1} = R_{C2} \,.\, I_{C2}$$

or:

$$I_{C1}/I_{C2} = R_{C2}/R_{C1} = \exp[q/k \,.\, \delta V_{BE12}/\delta T] \quad \ldots\ldots (7)$$

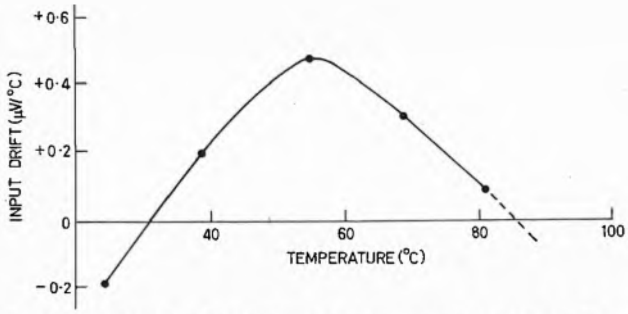

Fig. 2. Variation of drift voltage with temperature for circuit of Fig. 1

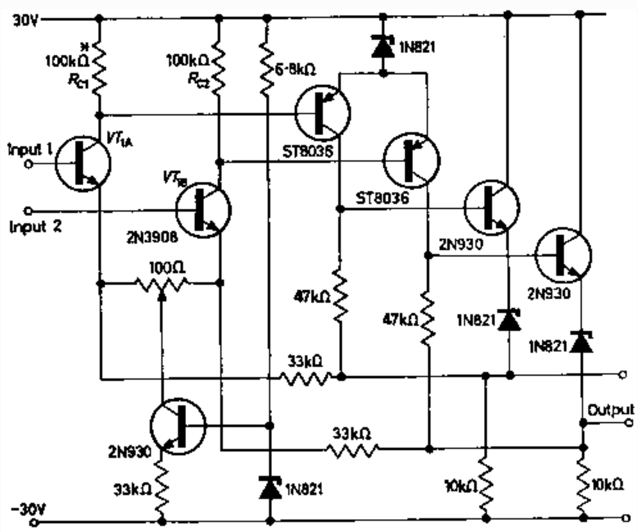

Fig. 3. Circuit for driving relatively low impedance

For the above example $\delta V_{BE12}/\delta T = -3{\cdot}5\mu$V/°C

$$I_{C1}/I_{C2} = R_{C2}/R_{C1} = \exp\left[\frac{6{\cdot}31 \times 10^{-19}}{1{\cdot}38 \times 10^{-23}}(-3{\cdot}5 \times 10)^{-6}\right] = 0{\cdot}96$$

$$R_{C1}/R_{C2} = I_{C2}/I_{C1} = 1{\cdot}04$$

Hence in order to compensate this drift, I_{C1} should be increased, I_{C2} decreased, or R_{C2} increased by 4 per cent, the last method being the simplest to arrange. A 10kΩ resistor added to R_{C2} did in fact reduce the amplifier drift to $-0{\cdot}4\mu$V/°C between 25°C and 60°C. However, it was found that with a 13·3kΩ resistor added to R_{C2}, the amplifier has a drift of $+0{\cdot}1\mu$V/°C between the same temperature range. The variation of drift voltage with temperature was as shown in Fig. 2. The amplifier had an 8 hour stability of $\pm 0{\cdot}2\mu$V when placed in an oil bath at (18 ± 2)°C.

For amplifiers suitable for driving relatively low impedance loads, e.g. in cases where a load current of about 250μA is required, the circuit shown in Fig. 3 gives an excellent performance. It operates as a current amplifier and employs current feedback to raise the internal resistance. Collector current in the first stage is about 64μA. When cycled between 17°C and 34°C, it exhibits a typical drift of $-0{\cdot}5\mu$V/°C. According to the compensating scheme outlined above, it was found that:

$$R_{C2}/R_{C1} = 1{\cdot}007$$

or:

$$\Delta R = (R_{C1} \,.\, 1{\cdot}007 - R_{C2}) = 700\Omega$$

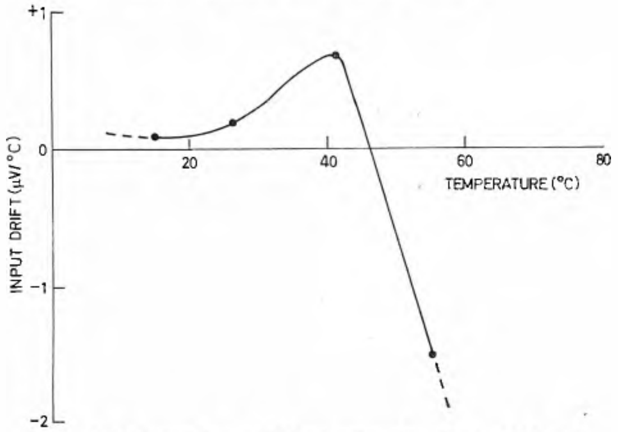

Fig. 4. Drift variation with temperature for circuit of Fig. 3

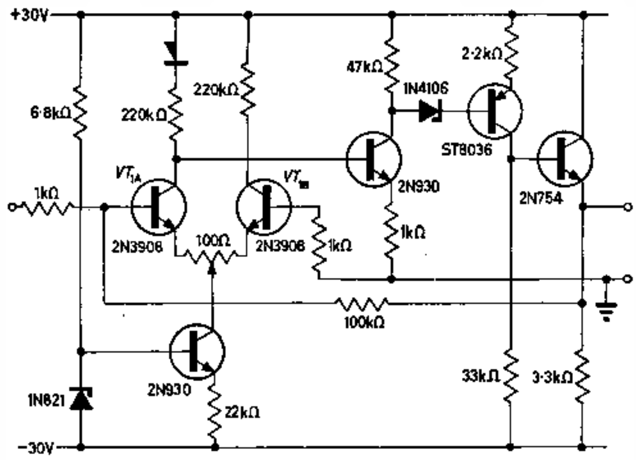

Fig. 5. Operational amplifier

When a 680Ω resistance was added to R_{C1} the drift was reduced to $0{\cdot}2\mu$V/°C over the same temperature range. Because the amplifier's noise was also of the same order of magnitude referred to input it was not practicable to reduce the drift still further. The drift variation with temperature of this amplifier is given in Fig. 4.

Both of the circuits described above are double ended and it is therefore difficult to employ them as operational amplifiers in analogue computing applications. Fig. 5 shows a suitable circuit for such arrangements. It has an open loop gain of 6 500 which was reduced to 100 by feedback resistors. Over the temperature range from 25°C to 55°C, it exhibited a drift of -12μV/°C. In this example, the compensating technique, as discussed above cannot be used so some other technique must be adopted. Other compensating methods include the use of a forward biased silicon diode or a thermistor. Temperature coefficient of forward voltage drop of a diode can be derived from equation (1) which can be matched with the temperature coefficient of base-emitter voltage of the transistor. However, without any matching, introduction of a silicon diode in the collector circuit typically reduces the drift to -4μV/°C between 24°C to 32°C. The variation of drift with temperature is shown in Fig. 6.

Examination of Figs. 2, 4 and 6 shows that drift is not uniform with temperature. In the last case, it is wholly due to mismatching of temperature coefficients of diode and transistor which can be improved to a certain extent. In the previous two circuits, the non-uniformity is the result of various factors, including the difference between the temperature coefficients of the collector resistors. The result of this mismatch of the collector resistors predominates at lower temperatures while at higher temperatures, the contribution due to difference in leakage currents becomes significant.

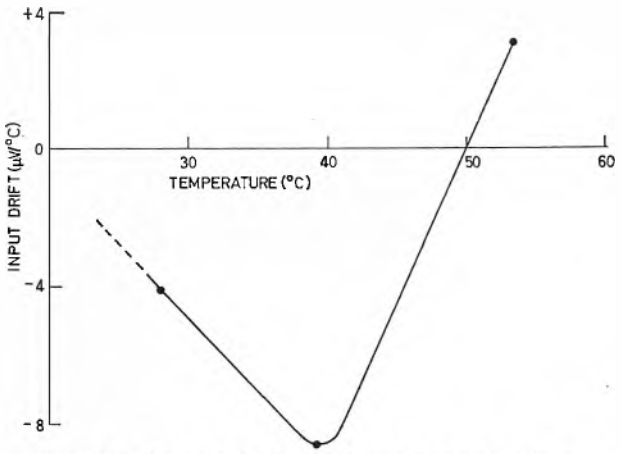

Fig. 6. Variation of drift with temperature for circuit of Fig. 5

In spite of non-uniformity, drift characteristics have regions of minimum value and have crossover points where the drift changes sign. The absolute magnitude of drift near the crossover points remains quite small. By adjusting these crossover points, it is possible to obtain an extremely low drift performance over any specific temperature range of interest.

It has been mentioned earlier that apart from drift, noise generated within the amplifier also limits the performance. In high quality amplifiers, the components should be so chosen as to introduce a minimum of noise. The sources of noise are transistors, Zener diodes, and resistors. In transistors, the flicker or low frequency noise is most important and varies inversely with frequency. It also depends upon the operating collector current, usually increasing with decreasing collector current. Modern low noise transistors like the 2N3908 or 2N930 have a noise voltage of 8×10^{-8}V at 120c/s with collector currents of 10μA. The resistors used were metal oxide film resistors having a noise voltage of $0{\cdot}1\mu$V/V. Zener diodes introduce more noise than resistors and transistors, so very good Zener diodes should be used. The 1N821-7 and other ultra stable Zener diodes usually have a noise voltage of about 5 to 10μV which can appear as noise voltage of 50 to 100μV at output, giving a noise of 0·1 to $0{\cdot}2\mu$V as referred to input in Fig. 1.

Acknowledgment

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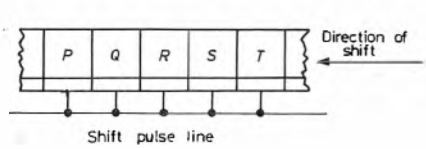
A Binary-Decimal Convertor Using Integrated Circuits

By K. J. Dean*, B.Sc., C.Eng., M.I.E.R.E., A.Inst.P., A.M.I.E.E.

The article describes the design of a binary-decimal convertor which is based on the logical control of a shift register. The design is carried out using integrated circuits. An interesting feature of the design is that the complexity of the circuits only increases linearly with the number of bits to be decoded.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

BECAUSE of the limitations of ten-state devices arithmetical operations in small digital machines are usually carried out in one of the binary codes. Nevertheless, such calculations are best inserted into the machine and results



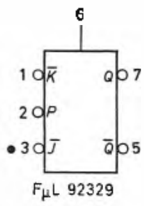


Fig. 1 (left). *P, Q, R, S* **and** *T* **are five stages of a shift register whose logical states must be controlled**

Fig. 2 (right). The logical symbol for an SGS-Fairchild Industrial Micrologic Flip Flop FμL 92329

obtained and displayed in decimal form[1]. The use of a binary-decimal code in the machine frequently increases the complexity of the logic to such an extent that on grounds of cost and speed it becomes uneconomical. Thus, it is most convenient to operate in simple binary code and use binary-decimal conversion at input and output.

A number of binary-decimal convertors have been described[2,3,4] which are relatively simple when the number of bits is small. These frequently consist of a diode matrix[5,6] which produces a ten-line output for each decimal digit, but the rate of increase of complexity of the matrix is not constant, becoming much greater for large numbers of bits.

A more simple approach is suggested here in which logically controlled shift registers are used for code conversion. This method increases linearly in complexity with the number of bits since the length of the register must be increased linearly to accommodate them.

In this article the design of a shift register will be discussed in which conversion takes place from binary code into a binary decimal code. A very simple diode matrix is then required to convert this into a ten-line output.

Principles of Operation

The binary-decimal code which will be used is 8421 b.c.d. This takes the same sequence as simple binary code from 0 to the count of 9. The following *six* binary states (10 to 15) are not used. Hence to convert a number such as 15 (binary 1111) to b.c.d. this filler of 6 must be added. The result (10101) when split into groups of four bits from the least significant digit is 15 in 8421 b.c.d. A similar method can be used for larger numbers as shown in the following example. Assume that the number is being shifted into a register and examined after each shift.

100's	tens	units		
			1 1 0 1 1 0 1 = binary 109	
		1	1 0 1 1 0 1	
		1 1	0 1 1 0 1	
		1 1 0	1 1 0 1	
		1 1 0 1	1 0 1	Greater than 9, add 6
	1	0 0 1 1	1 0 1	shift one place left
	1 0	0 1 1 1	0 1	shift one place left
	1 0 0	1 1 1 0	1	add 6 to units column
	1 0 1	0 1 0 0	1	shift one place left
	1 0 1 0	1 0 0 1		add 6 to tens column
1	0 0 0 0	1 0 0 1		stop
1	0	9		

In this example the numbers were examined commencing with the most significant digit. When any of the groups of four digits exceeded nine, the filler of six was added.

A further simplification can be made when it is realized that since the filler (110) has a '0' in its least significant place, the addition can be carried out on all codes greater than four, the filler being reduced to three (11). The example is repeated in this way.

		1 1 0	1 1 0 1	Add 3 to units column
		1 0 0 1	1 1 0 1	Shift one place left
	1	0 0 1 1	1 0 1	Shift one place left
	1 0	0 1 1 1	0 1	add 3 to units column
	1 0	1 0 1 0	0 1	shift one place left
	1 0 1	0 1 0 0	1	add 3 to tens column
	1 0 0 0	0 1 0 0	1	shift one place left
1	0 0 0 0	1 0 0 1		stop
1	0	9		

The complete convertor, therefore, consists of a shift register which receives information in binary form, and whose combinational logic is so arranged that three is added on each shift only when required by the states held in the register. The addition and the shifting are, in fact, combined in the following design.

Design Method

The design of the shift register may be carried out using Karnaugh[7] maps to determine the input conditions. These conditions will be determined here for one decade only but the extension to other decades is entirely repetitive. Five stages, *P, Q, R, S* and *T*, of the shift register are shown in Fig. 1.

The number is shifted in to the register at *T* and the states of *Q, R, S* and *T* are examined and when a count of 5 (0101) or greater occurs, the number 3 (0011) is

* *Letchworth College of Technology.*

added at the next shift pulse. Table 1 shows all the possible states of the elements before and after a shift pulse is applied. It should be noted that states corresponding to 10 and above cannot occur before the pulse, since such states would have been dealt with one pulse earlier. For example, binary 10 (1010) would have been modified one pulse earlier from 5 (0101) to 8 (1000).

TABLE 1

STATES BEFORE SHIFT PULSE				STATES AFTER SHIFT PULSE				
Q	R	S	T	P	Q	R	S	
0	0	0	0	0	0	0	0	
0	0	0	1	0	0	0	1	
0	0	1	0	0	0	1	0	
0	0	1	1	0	0	1	1	
0	1	0	0	0	1	0	0	
0	1	0	1	1	0	0	0	
0	1	1	0	1	0	0	1	
0	1	1	1	1	0	1	0	3 added
1	0	0	0	1	0	1	1	
1	0	0	1	1	1	0	0	

Before the design can proceed further a shift pulse element must be selected. Here an SGS-Fairchild Industrial Micrologic JK flip flop $F\mu L$ 92329 will be used. The connexions to this element are shown in Fig. 2. Its truth table is given in Table 2. In the table Q_n is the output after n clock pulses and Q_{n+1} is the output after $n + 1$ pulses.

TABLE 2

P	$\overline{J}$	$\overline{K}$	Q_{n+1}
0	X	X	Q_n
1	0	1	1
1	1	0	0
1	1	1	Q_n
1	0	0	$\overline{Q_n}$

X denotes an immaterial condition

One must now determine those conditions which are necessary and sufficient in order to steer the element correctly when a shift pulse is applied. For example, it can be seen from Table 2 that if a flip-flop is to be switched from '0' to '1' the state of $\overline{K}$ is immaterial, so long as J is at the '0' level. Table 3, which summarizes the conditions for all the possible transitions, is derived from Table 2.

TABLE 3

FROM	TO	NECESSARY CONDITIONS	
		$\overline{J}$	$\overline{K}$
0	0	1	X
0	1	0	X
1	0	X	0
1	1	X	1
X	0	1	0
X	1	0	1

From the conditions given in Table 3 it is now possible to draw up maps to derive the input conditions for P, Q, R and S based on the states of Q, R, S and T. The maps, in Table 4, give the logic required to implement the conditions of Table 1. In this table and subsequently the symbols j, k, etc will be written for $\overline{J}$, $\overline{K}$, etc where these occur as suffixes.

TABLE 4

ST \ QR	00	01	11	10
00	0	4	X	8
01	1	5	X	9
11	3	7	X	X
10	2	6	X	X

P_j

ST \ QR	00	01	11	10
00	1	1	X	0
01	1	0	X	0
11	1	0	X	X
10	1	0	X	X

P_k

ST \ QR	00	01	11	10
00	0	0	X	1
01	0	1	X	1
11	0	1	X	X
10	0	1	X	X

Q_j

ST \ QR	00	01	11	10
00	1	0	X	X
01	1	1	X	X
11	1	1	X	X
10	1	1	X	X

Q_k

ST \ QR	00	01	11	10
00	X	X	X	0
01	X	X	X	1
11	X	X	X	X
10	X	X	X	X

R_j

ST \ QR	00	01	11	10
00	1	X	X	0
01	1	X	X	1
11	0	X	X	X
10	0	X	X	X

R_k

ST \ QR	00	01	11	10
00	X	0	X	X
01	X	0	X	X
11	X	1	X	X
10	X	0	X	X

S_j

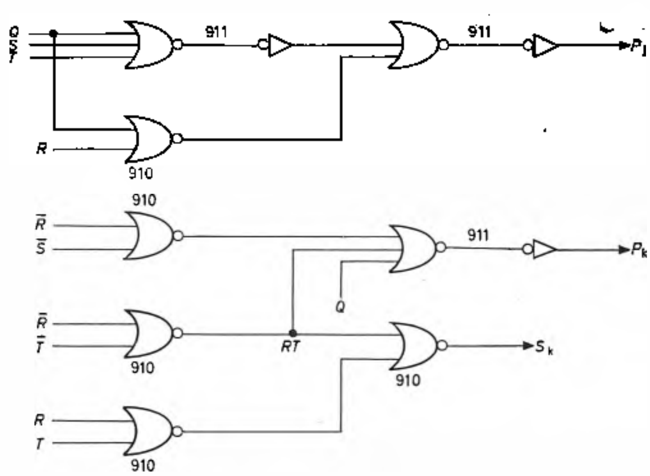
ST \ QR	00	01	11	10
00	1	1	X	0
01	0	1	X	1
11	X	X	X	X
10	X	X	X	X

S_k

ST \ QR	00	01	11	10
00	X	X	X	X
01	X	X	X	X
11	1	0	X	X
10	0	1	X	X

$P_j = \overline{Q}\overline{S}\overline{T} + \overline{Q}\overline{R}$; $P_k = Q + RT + RS$; $Q_j = \overline{R} + T + S$; $Q_k = T$; $R_j = \overline{Q}\overline{S} + QT$; $R_k = ST$; $S_j = \overline{Q}\overline{T} + RT + QT$; $S_k = \overline{R}T + R\overline{T}$.

Fig. 3. Arrangement of logic gates to control the inputs of the flip flop, P



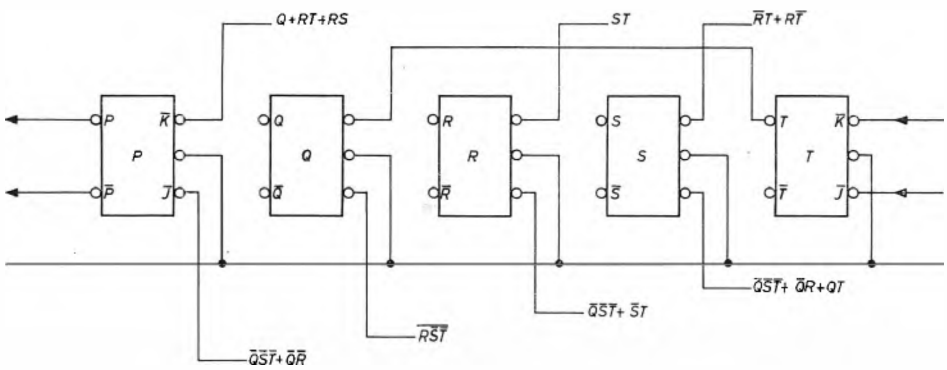


Fig. 4. The elements of the binary/decimal convertor. P, Q, R, S and T, are shown together with the logical conditions which must be generated to control them

design is completed by using these gates to derive the input conditions for each of the flip-flops. Some of these gates are illustrated in Fig. 3 and the outline of the complete register is shown in Fig. 4.

In this article the design of a binary to b.c.d. convertor has been considered. A somewhat similar procedure can be adopted for b.c.d. to binary conversion, except that provision must be made for subtracting three, instead of adding it.

It should be noticed that by avoiding complete minimization less functions need to be generated. This is because the minimization of the complete system must be carried out rather than the minimization of the input conditions of each flip-flop independently of those for any of the other flip-flops. For example, if $R_J=\overline{Q}\overline{S}\overline{T}+\overline{S}T$, the first term can be used for P_J and S_J also.

The SGS-Fairchild range of integrated circuits used here includes 2-input and 4-input gates which have the functions OR and NOR in the positive logic convention. The

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Release Techniques for Injection Moulding

The Wiring and Connectors Division of the Plessey Components Group has successfully employed advanced release techniques to overcome production problems on the injection moulding of silicone rubber components, and achieve a better finish on the parts. They have also substantially reduced the number of rejects previously experienced.

The Plessey Wiring and Connectors Division, at Swindon, is one of the largest producers of professional connectors in the world. Every year the Division manufactures tens of thousands of complete wiring harnesses, looms and junction boxes, and millions of connectors with their associated piece parts. It has developed advanced versions of previous connector ranges, as with the UK-AN series, and has designed many new and unique connectors and interconnexions—such as the Mark II for mining equipment and interconnexions made by UNIcrimp tools. The Division has also pioneered in the utilization of many new materials and production techniques for high-performance connector inserts.

Products manufactured by the Division include plugs and sockets, electrical connectors, fuse and terminal blocks, cable assemblies, looms, bonding leads, control panels, junction boxes, starter cables, suppressor leads, preformed wiring connexions, ignition harnesses and crimping tools.

The injection moulding of rubber inserts is undertaken by the Division's Rubber Moulding Section, which serves the Connectors Unit. Two principal moulding processes are employed—the loose pin tooling method and the semi-automatic rotary machine method. With the first method, the moulds are made by hand—the required number of pins being inserted into them by operatives—and then placed on the moulding stations of an adapted Pilot hydraulic press. The press has four sliding moulding stations, situated under the injection positions, and when the moulds are in place, the silicone rubber is fed from the injection cylinder into the mould.

Rotary machines are employed in the production of inserts when they are manufactured in large quantities. Large, semi-automatic machines designed by the Division are used for this purpose. They have ten or twelve moulding stations, and are capable of achieving one moulding per minute.

The silicone rubber used for inserts is purchased by the Division in gum form, and then modified to meet the required specifications. It is compounded and blended in the Division to meet specific performance requirements. Where resistance to degradation by kerosene is required a fluorinated silicone gum stock is employed.

Each week the Division's Rubber Moulding Section completes some 6 000 to 7 000 inserts of a wide variety of sizes. One crucial problem in the moulding process is the release of the mould from the silicone rubber. A typical insert, for example, may have a diameter of ¾in and as many as 24 holes of 1/16in diameter. A clean release is therefore essential.

To solve these release problems, the Moulding Section tested a number of release agents, but without achieving any noticeable success. Recently, however, tests were carried out with Rocol MRS (Mould Release Spray). This product, manufactured by Rocol Ltd, of Leeds, is specially designed to facilitate quick and easy release of plastic and rubber products from all types of moulds, formers, dies, rolls and presses.

Rocol MRS, applied with an aerosol spray, deposits a fluorocarbon film on both metallic and non-metallic surfaces. Because Rocol MRS's fluorocarbon is a telomer rather than a polymer, it can be used as a spray without the addition of resin to bond the surface of the film. Consequently the release agent does not migrate into plastics, rubber, or other materials to pose subsequent production problems. Furthermore, it is colourless, non-staining, and non-toxic, has a high melting point (275°C), and is unaffected by cold, water, solvents, or chemicals while repelling oil, water and dirt.

The tests carried out by the Moulding Section with Rocol MRS have proved highly successful, and showed that it is not necessary to apply the product to the moulds as frequently as was required with other types of release agents. A much cleaner release has been achieved, and Rocol MRS does not build up on the pins of the mould and so leave 'rivers', or tear the walls of the holes. It also withstands the high temperatures under which the moulding takes place. For these reasons, it is regularly used whenever a new mould is adopted, or whenever there is any evidence of a release problem on existing moulds.

Analysis of the Complementary Pair Emitter-Follower

By P. Mars*, B.Sc.

A matrix analysis of the complementary pair emitter-follower is performed and overall circuit parameters developed. The circuit has better characteristics than the conventional single transistor emitter-follower and possesses good d.c. quiescent stability.

(Voir page 695 pour le résumé en français; Zusammenfassung in deutscher Sprache auf Seite 702)

A CIRCUIT is often required which provides the normal single transistor emitter-follower properties but possesses improved performance characteristics. Simply cascading single transistor emitter-followers does not always provide a satisfactory performance. For example the cascaded arrangement will have a definite minimum value for the output impedance[1]. An alternative solution is provided by the compound complementary pair configuration[2].

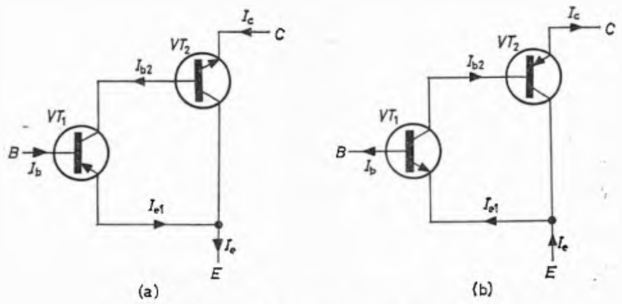

Fig. 1 (a). PNP complementary pair
(b). NPN complementary pair

Fig. 1(a) shows a complementary superbeta configuration which in effect synthesizes an equivalent pnp transistor. VT_1 determines the effective polarity of the superbeta pair since its emitter follows the base signal. Transistor VT_2 acts merely as a current driver. On interchanging the positions of VT_1 and VT_2 so that they appear as shown in Fig. 1(b) an equivalent npn transistor is synthesized. In order to demonstrate the superbeta property of the complementary pair configuration consider for example Fig. 1(a).

The overall beta is given by:

$$\beta = I_c/I_b = I_{e2}/I_{b1} = \frac{I_{b2}(1+\beta_2)}{I_{b1}}$$

But

$$I_{b2} = I_{c1} = \beta_1 I_{b1}$$

hence:

$$\beta = \frac{\beta_1 I_{b1}(1+\beta_2)}{I_{b1}} = \beta_1(1+\beta_2) \simeq \beta_1\beta_2$$

Circuit Analysis

Fig. 2 shows the pnp complementary pair connected in the common collector configuration. The a.c. equivalent circuit of Fig. 2 is shown in Fig. 3. For analysis the circuit of Fig. 2 is split into elementary four terminal networks as shown. In the figure R_L represents the parallel combination of the resistor R_4 and the external load.

Fig. 4 shows transistor VT_2 with the resistor R_3 connected between base and emitter. For the purpose of matrix analysis a fictitious collector lead is introduced as shown[3].

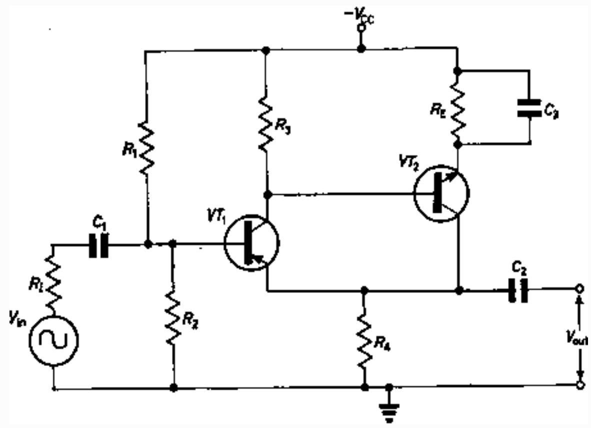

Fig. 2. PNP complementary pair emitter-follower

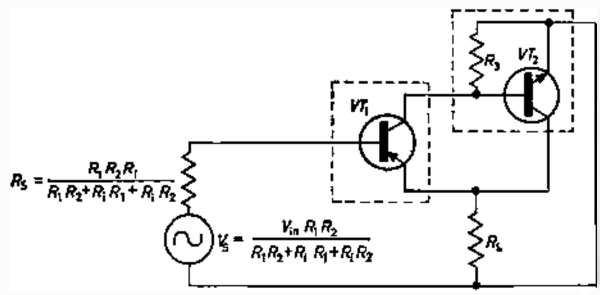

Fig. 3. A.C. equivalent circuit of Fig. 2

The Y parameters for the transistor are given by:

$$[Y]_T = \begin{bmatrix} 1/h_{ic2} & -h_{rc2}/h_{ic2} \\ h_{fc2}/h_{ic2} & \Delta h_{c2}/h_{ic2} \end{bmatrix}$$

Hence for the resistor transistor combination:

$$[Y]_2 = \begin{bmatrix} (1/h_{ic2}) + Y_3 & (-h_{rc2}/h_{ic2}) - Y_3 \\ (h_{fc2}/h_{ic2}) - Y_3 & (\Delta h_{c2}/h_{ic2}) + Y_3 \end{bmatrix}$$

Here the subscripts 1 and 2 will be used to represent parameters associated with transistors VT_1 and VT_2 respectively.

Converting from y parameters to h parameters gives the equivalent $[H]$ matrix.

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$$[H]_2 = \frac{1}{h_{ic2} + R_3}\begin{bmatrix} h_{ic2}R_3 & h_{ic2} + h_{rc2}R_3 \\ h_{fc2}R_3 - h_{ic2} & h_{oc2}R_3 + (1 + \Delta h_{c2} - h_{fc2} - h_{rc2}) \end{bmatrix}$$

Let this matrix be represented by:

$$\begin{bmatrix} h_{ic2}' & h_{rc2}' \\ h_{fc2}' & h_{oc2}' \end{bmatrix}$$

The equivalent $[A]$ matrix is given by:

$$[A]_2 = 1/h_{fc2}'\begin{bmatrix} -\Delta h_{c2}' & -h_{ic2}' \\ -h_{oc2}' & -1 \end{bmatrix}$$

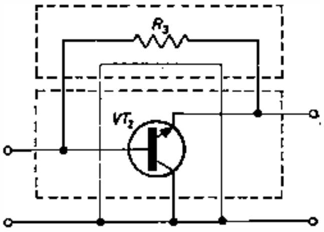

Fig. 4. Circuit for matrix analysis of VT_2 and R_3 combination

The $[A]$ matrix for transistor VT_1 is given by:

$$[A]_1 = 1/h_{fe1}'\begin{bmatrix} -\Delta h_{e1}' & -h_{ie1}' \\ -h_{oe1}' & -1 \end{bmatrix}$$

where in the case of Fig. 2.

$$h_{ie1}' = h_{ie1} \qquad h_{re1}' = h_{re1}$$

$$h_{fe1}' = h_{fe1} \qquad h_{oe1}' = h_{oe1}$$

$$\Delta h_{e1}' = \Delta h_{e1} = h_{ie1}h_{oe1} - h_{re1}h_{fe1}$$

Fig. 5 shows the circuit of Fig. 2 modified in order to reduce the shunting effect of the potential divider R_1, R_2 on the circuit input impedance. Fig. 6 shows the a.c. equivalent circuit of Fig. 5. R_L now represents the parallel combination of R_4, the external load and R_1 and R_2 in parallel.

In order to obtain the $[A]$ matrix of transistor VT_1 with input shunted by R_5 it is merely necessary to transpose the matrix $[A]_2$ from the common collector to the common emitter configuration.

This gives:

$$h_{ie1}' = \frac{h_{ie1}R_5}{h_{ie1} + R_5} \qquad h_{re1}' = \frac{h_{re1}R_5}{h_{ie1} + R_5}$$

$$h_{fe1}' = \frac{h_{fe1}R_5}{h_{ie1} + R_5} \qquad h_{oe1}' = \frac{\Delta h_{e1} + h_{oe1}R_5}{h_{ie1} + R_5}$$

$$\Delta h_{e1}' = \frac{\Delta h_{e1}R_5}{h_{ie1} + R_5}$$

The overall $[A]$ matrix for the cascade connexion of VT_1 and VT_2 is thus given by:

$$[A]_1[A]_2 = \frac{1}{h_{fe1}'h_{fc2}'}\begin{bmatrix} h_{ie1}'h_{oc2}' + \Delta h_{e1}'\Delta h_{c2}' & \Delta h_{e1}'h_{ic2}' + h_{ie1}' \\ h_{oc2}' + h_{oe1}'\Delta h_{c2}' & 1 + h_{oe1}'h_{ic2}' \end{bmatrix}$$

The equivalent $[H]$ matrix is:

$$\frac{1}{1 + h_{oe1}'h_{ic2}'}\begin{bmatrix} h_{ic2}'\Delta h_{e1}' + h_{ie1}' & h_{re1}'h_{rc2}' \\ -h_{fe1}'h_{fc2}' & h_{oc2}' + h_{oe1}'\Delta h_{c2}' \end{bmatrix}$$

This is essentially the $[H]$ matrix for the complementary pair with respect to the emitter terminal. Transposing the above matrix to the common collector configuration produces the following $[H]$ matrix.

$$\begin{bmatrix} \dfrac{h_{ie1}' + \Delta h_{e1}'h_{ic2}'}{1 + h_{oe1}'h_{ic2}'} & 1 - \dfrac{h_{re1}'h_{rc2}'}{1 + h_{oe1}'h_{ic2}'} \\ \dfrac{h_{fe1}'h_{fc2}'}{1 + h_{oe1}'h_{ic2}'} - 1 & \dfrac{h_{oc2}' + h_{oe1}'\Delta h_{c2}'}{1 + h_{oe1}'h_{ic2}'} \end{bmatrix}$$

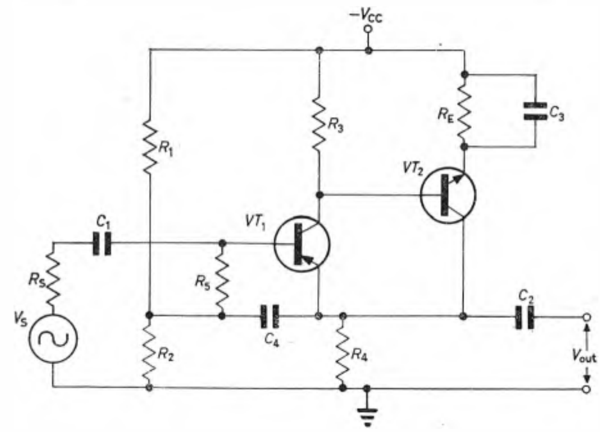

Fig. 5. Circuit of Fig. 2 modified to reduce shunting effect of potential divider

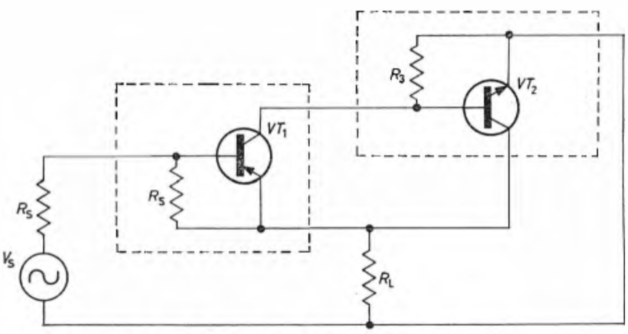

Fig. 6. A.C. equivalent circuit of Fig. 5

For convenience of analysis let the above matrix be equivalent to:

$$\begin{bmatrix} h_{ic} & h_{rc} \\ h_{fc} & h_{oc} \end{bmatrix}$$

Parameter Approximations

$$h_{ic} = \frac{h_{ie1}' + \Delta h_{e1}'h_{ic2}'}{1 + h_{oe1}'h_{ic2}'} = \frac{h_{ie1}' + \Delta h_{e1}'h_{ie2}'}{1 + h_{oe1}'h_{ie2}'}$$

where the parameters h_{e2}' represent the common emitter equivalents of the common collector parameters h_{c2}'.

If

$$\Delta h_{e1}'h_{ie2}' \ll h_{ie1}'$$

$$h_{ic} \simeq \frac{h_{ie1}'}{1 + h_{oe1}'h_{ie2}'}$$

$$h_{rc} = 1 - \frac{h_{re1}'h_{rc2}'}{1 + h_{oe1}'h_{ic2}'} = 1 - \frac{h_{re1}'(1 - h_{re2}')}{1 + h_{oe1}'h_{ie2}'}$$

when

$$1 + h_{oe1}'h_{ie2}' \gg h_{re1}'(1 - h_{re2}')$$

$$h_{rc} \simeq 1$$

$$h_{fc} = \frac{h_{fe1}'h_{fc2}'}{1 + h_{oe1}'h_{ic2}'} - 1 = \frac{-h_{fe1}'(1 + h_{fe2}')}{1 + h_{oe1}'h_{ie2}'} - 1$$

If

$$h_{fe1}'(1+h_{fe2}') \gg 1 + h_{oe1}'h_{ie2}'$$

$$h_{fc} \simeq \frac{-h_{fe1}'(1+h_{fe2}')}{1+h_{oe1}'h_{ie2}'}$$

$$h_{oc} = \frac{h_{oe2}' + h_{oe1}'\Delta h_{e2}'}{1+h_{oe1}'h_{ie2}'} = \frac{h_{oe2}' + h_{oe1}'(1+h_{fe2}'+\Delta h_{e2}'-h_{re2}')}{1+h_{oe1}'h_{ie2}'}$$

when

$$\Delta h_{e2}' - h_{re2}' \ll 1 + h_{fe2}'$$

and

$$h_{oe1}'(1+h_{fe2}') \gg h_{oe2}'$$

$$h_{oc} \simeq \frac{h_{oe1}'(1+h_{fe2}')}{1+h_{oe1}'h_{ie2}'}$$

Hence final $[H]$ matrix becomes:

$$\begin{bmatrix} \dfrac{h_{ie1}'}{1+h_{oe1}'h_{ie2}'} & 1 \\ \dfrac{-h_{fe1}'(1+h_{fe2}')}{1+h_{oe1}'h_{ie2}'} & \dfrac{h_{oe1}'(1+h_{fe2}')}{1+h_{oe1}'h_{ie3}'} \end{bmatrix}$$

Introduce term Δ

where

$$\Delta = h_{ic}h_{oc} - h_{fc}h_{rc}$$

$$= \frac{(1+h_{fe2}')\,[h_{ie1}'h_{oe1}' + h_{fe1}'(1+h_{oe1}'h_{ie3}')]}{(1+h_{oe1}'h_{ie2}')^2}$$

If

$$h_{fe1}'(1+h_{oe1}'h_{ie2}') \gg h_{ie1}'h_{oe1}'$$

$$\Delta \simeq \frac{h_{fe1}'(1+h_{fe2}')}{1+h_{oe1}'h_{ie2}'}$$

Circuit Characteristics

Substitution in the normal equations used for the calculation of emitter-follower input impedance, voltage gain, current gain and output impedance produces the following relationships:

Input impedance

$$Z_{IN} = \frac{h_{ic} + R_L\Delta}{1+h_{oc}R_L} = \frac{h_{ie1}' + R_L h_{fe1}'(1+h_{fe2}')}{1+h_{oe1}'[h_{ie2}' + R_L(1+h_{fe2}')]} \qquad \ldots\ldots\ldots\ldots (1)$$

If

$$R_L h_{fe1}'(1+h_{fe2}') \gg h_{ie1}'$$

and

$$h_{oe1}'[h_{ie2}' + R_L(1+h_{fe2}')] \ll 1$$

then:

$$Z_{IN} \simeq R_L h_{fe1}'h_{fe2}'$$

Voltage Gain

$$A_v = \frac{-h_{fc}R_L}{R_L(h_{oc}R_s + \Delta) + h_{ic} + R_s}$$

$$= \frac{h_{fe1}'R_L(1+h_{fe2}')}{R_L(1+h_{fe2}')[R_s h_{oe1}' + h_{fe1}'] + h_{ie1}' + R_s(1+h_{oe1}'h_{ie2}')} \qquad \ldots\ldots (2)$$

with

$$R_s = 0 \text{ and } R_L h_{fe1}'(1+h_{fe2}') \gg h_{ie1}'$$

$$A_v \simeq 1$$

$$\text{Current Gain} = A_i = \frac{h_{fc}}{1+h_{oc}R_L}$$

$$= \frac{h_{fe1}'(1+h_{fe2}')}{1+h_{oe1}'[h_{ie2}' + R_L(1+h_{fe2}')]} \qquad \ldots\ldots\ldots\ldots (3)$$

with

$$h_{oe1}'[h_{ie2}' + R_L(1+h_{fe2}')] \ll 1$$

$$A_i \simeq h_{fe1}'h_{fe2}'$$

Output Impedance

$$Z_{OUT} = \frac{R_s + h_{ic}}{R_s h_{oc} + \Delta}$$

$$= \frac{h_{ie1}' + R_s(1+h_{oe1}'h_{ie2}')}{(1+h_{fe2}')(R_s h_{oe1}' + h_{fe1}')} \qquad \ldots\ldots (4)$$

If

$h_{fe2}' \gg 1$ and $h_{fe1}' \gg R_s h_{oe1}'$

$R_s \gg h_{ie1}'$ and $h_{oe1}'h_{ie2}' \ll 1$

$$Z_{OUT} \simeq \frac{R_s}{h_{fe1}' \,.\, h_{fe2}'}$$

Typical Design

Consider the design of the complementary pair emitter-follower shown in Fig. 5.

Suppose the circuit has the following component values:

$R_1 = R_2 = 10\text{k}\Omega$, $R_s = 50\text{k}\Omega$, $R_4 = 3{\cdot}3\text{k}\Omega$,

$R_3 = R_5 = 5{\cdot}2\text{k}\Omega$, $C_1 = 0{\cdot}1\mu\text{F}$,

$R_E = 2{\cdot}2\text{k}\Omega$, $C_2 = C_3 = C_4 = 50\mu\text{F}$

Let $V_{cc} = 10\text{V}$ and assume the transistors used have the following parameters:

VT_1	$h_{ie1} = 1{\cdot}5\text{k}\Omega$	$h_{re1} = 6 \times 10^{-4}$
	$h_{fe1} = 72$	$h_{oe1} = 30\mu\Omega^{-1}$
VT_2	$h_{ie2} = 2{\cdot}1\text{k}\Omega$	$h_{fe2} = 45$

Substitution of the above values in equations (1), (2) and (3) and (4) gives the following circuit characteristics:

$Z_{IN} = 1{\cdot}43\text{M}\Omega$, $A_v = 0{\cdot}97$, $A_i = 710$, $Z_{OUT} = 27{\cdot}8\Omega$

Using the same pnp transistor a conventional single transistor emitter-follower would produce the following characteristics:

$Z_{IN} = 140\text{k}\Omega$, $A_i = 69$, $A_v = 0{\cdot}73$, $Z_{OUT} = 690\Omega$

Conclusions

The complementary pair emitter-follower provides a higher input impedance, lower output impedance, and a voltage gain nearer unity than that obtainable from a single transistor emitter-follower. Experimental results were found to correspond closely to the theoretical predictions. Inspection of the various expressions shows that to increase the input impedance and current gain and decrease the output impedance the resistors R_3 and R_5 should be as high as possible. However it should be observed that the resistor R_5 determines the d.c. stability of Fig. 5 and R_3 limits the collector current of VT_2.

Acknowledgments

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Simple Grounded Base Stage for use with Coaxial Cables

By B. L. Hart*, B.Sc., A.M.I.E.R.E.

In both high frequency analogue and digital circuits it is sometimes necessary to feed a signal to a transistor stage via a coaxial cable. In this article a suitable circuit is described which employs a double-transistor. An analysis of the design and performance is given.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

IN high frequency analogue and high speed digital circuits, there are a number of occasions when it is required to couple a signal (which may be positive or negative with respect to an earth reference level) on a coaxial cable into a transistor stage, the transfer characteristic of which is well-defined. D.C. coupling might be desirable in order to overcome problems of pattern sensitivity which can arise if the signal is a.c. coupled via a

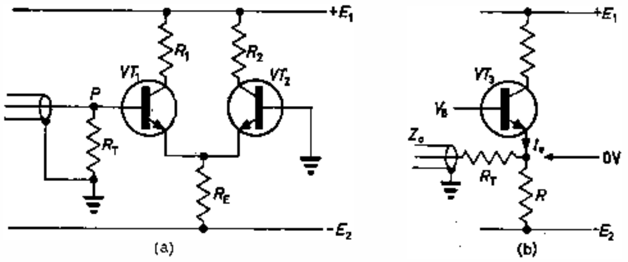

Fig. 1 (above) (a). Input stage using a differential amplifier.
(b). Grounded base input stage

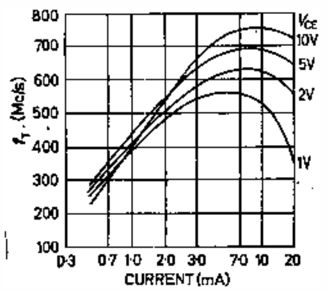

Fig. 2. (left). Typical f_T contours for transistor type SL 301-A

(Courtesy of Semiconductors Ltd)

capacitor. However, difficulties can occur with d.c. coupling because a minor deviation from earth potential at the input can cause significant d.c. loading on the driving unit and thus upset the bias arrangements of that stage.

Circuit Details

One method by which the object might be achieved is shown in Fig. 1(a) in which a conventional differential amplifier uses npn transistors. Point P will differ from earth by a very small amount if matched transistors are used. Furthermore, the variation in potential with temperature is very small.

Disadvantages of this arrangement are, firstly, that the input impedance looking into the base of VT_1 may become comparable with R_T at high frequencies. These effects can be partially offset by using small resistors in the emitter leads of VT_1 and VT_2 to give emitter degeneration. An attractive alternative solution is the use of a grounded-base configuration. In Fig. 1(b) the base of VT_3, driven from a low impedance d.c. source, is arranged to be such that its emitter is at earth potential.

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Then $|I_e| = E_2/R$, in the absence of an input signal, and $|I_c| = \alpha E_2/R$ where α = d.c. common-base current gain of VT_3 (I_{co} is neglected). The low frequency impedance looking into VT_3's emitter is $(KTR/qE_2) + r_{bb'}/(\beta + 1)$ where $r_{bb'}$ = incremental extrinsic resistance of the base and the other symbols are defined in the Appendix. For $T = 300°K$ this becomes $\simeq (26R/E_2)\Omega$ if R is in kilohms and E_2 in volts, and if $r_{bb'}/(\beta + 1)$ can be neglected.

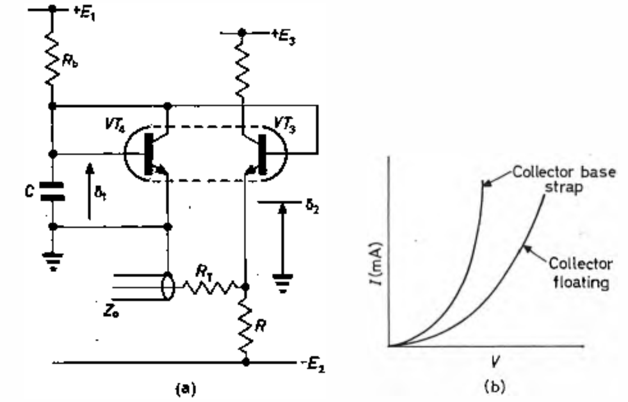

Fig. 3 (a). Grounded base to stage using integrated transistor pair
(b). Diode behaviour of a transistor

Input stage design involves the choice of I_e such that

(a) $Z_o = R_T + (26R/E_2)$ and

(b) I_e gives a suitable value of f_T.

Most manufacturers of high frequency transistors give graphical information relating to f_T contours and these are shown, for the particular transistor type used, in Fig. 2.

Operating at, say, 5mA ensures that the transistor l.f. emitter input impedance is small compared with the Z_o obtaining with most coaxial cables. Thus small variations in I_e due to the input signal will have little effect on input matching. The problem remains of biasing the base of VT_3 suitably such that its emitter is at zero volts for zero input signal, and ensuring that the variation of this potential with respect to temperature is very small. Using a potentiometer on the base means a manual adjustment and a variation of $\simeq 2$mV/°C in emitter potential. A more elegant circuit is shown in Fig. 3(a). In this VT_3, VT_4 are two transistors on the same slice of semiconductor material and mounted on the same header (e.g. SL301-A). This helps to promote essentially matched static and dynamic behaviour.

VT_4 with its collector-base strap has enhanced performance over the arrangement where the collector of VT_4 is floating or where a diode is used. This is shown in Fig. 3(b), (actually there is a 'kink' in the I, V_b curve at high currents, e.g. 30mA, due to transistor saturation effects but the transistor is not usually strapped for use as a diode at these levels).

Calculations show that, in principle, R_b may be selected so that for a given I_e, the d.c. output level at VT_3's emitter, i.e. δ_2, is zero. This arises because the base-emitter voltage drops of VT_3, VT_4 are both equal. Such a state of affairs is amenable to calculation if it is known which of VT_3, VT_4 has the higher V_{be} for a given emitter current. However, this is not usually known unless measurements are made and this defeats the object of the exercise. Thus the technique adopted is to operate VT_3, VT_4 at the same emitter current level and accept the small departure (e.g. $\pm$ 5mV) from zero that such a procedure entails. It is shown in Appendix 1(a) that for fixed E_1, E_2 it is required that:

$$R_b/R \simeq \left(\frac{E_1 - \delta_1}{E_2}\right)\left(\frac{\beta_2 + 1}{\beta_2 + 2}\right) \qquad (1)$$

and in Appendix 1(b) that small percentage changes in the ratio R_b/R cause very small variations in δ_2. Furthermore,

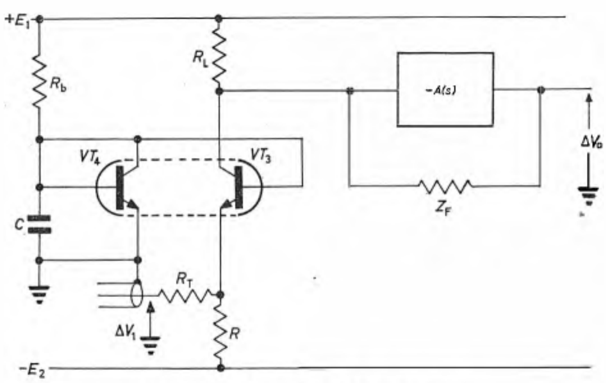

Fig. 4. Complete amplifier arrangement

it is shown in Appendix 2 that variations in δ_2 with T are very small.

Equation (1) is of interest since in both semiconductor integrated circuits and thin film arrangements it is relatively simple to produce a precise resistor ratio (whereas production of a resistance, by itself, to a close tolerance is not so easy). Thus no manual adjustment is needed in the base circuit of VT_4 unless it is necessary for δ_2 to be closer to earth than, say, $\pm$5mV.

Fig. 4 shows how a grounded base stage may be used in conjunction with a current-in-to-voltage-out amplifier (unit $-A(s)$ together with feedback impedance Z_F) to give a circuit with a well-defined voltage transfer function $\Delta V_o/\Delta V_i = -Z_F/R_T$ for small input signals.

Most of what has been discussed so far concerns the d.c. performance. From the point of view of h.f. performance it is interesting to note that Coli, Lupini and Silvestrini[1], have used a grounded base circuit, with a terminating resistor connected to the emitter but a potentiometer on the base, as the input to a wideband amplifier extending from d.c. to 175Mc/s and having a voltage gain of $\times$ 10 flat to 40Mc/s. At much higher frequencies a more complex input network than a series resistor is mandatory for good matching to the cable.

Results

Calculations for the Semiconductors Ltd SL301-A [two transistors matched to $\pm$5mV at $I_e = 1$mA, $V_{CE} = 5$V] at 5mA gave $R_b = 1{\cdot}81$kΩ for $R = 2$kΩ.

R_b and R were set up on 0·05 per cent resistance boxes and δ_2 observed as +5mV. Then R_b was decreased by 5 per cent and at the same time R was increased by 5 per cent; δ_2 increased to +8mV. Considering the fact that a digital voltmeter was used (a 4 or more digit type being, of course, preferable if available) this gives good agreement with the predicted change of approximately 2·6mV. The theory is based on the assumption that the matched transistors VT_3, VT_4 have the 'same' V_{be} at the same I_e. This is strictly true only for the case when their V_{cb}'s are also equal. However, with modern planar transistors base width modulation factors are very small so the fact that V_{cb} of VT_3 was 0V while that of VT_4 was +5V caused an insignificant difference.

Appendix

(1(a)) Calculation of R_b/R

Fig. 5 shows Fig. 3(a) redrawn in order to incorporate a d.c. model of the transistors VT_3, VT_4 which is sufficient for the present application.

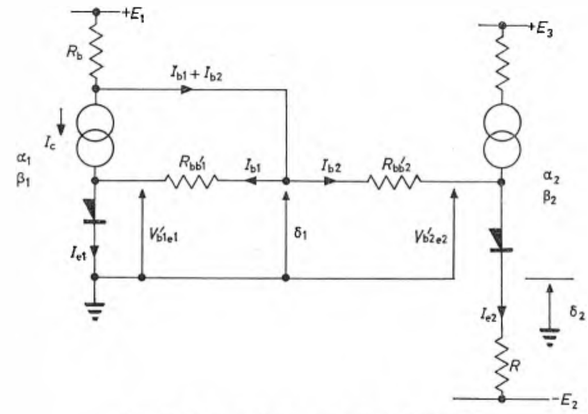

Fig. 5. D.C. equivalent circuit of Fig. 3(a)

The d.c. base resistance $R_{bb'}$ and α are both functions of I_e. It follows from the equations of Ebers & Moll[3] that if a homogenous base npn transistor, in which base width modulation effects are neglected, is operated in the forward active mode.

$$I_e \simeq I_s\left[\exp\left(qV_{b'e}/nKT\right) - 1\right] \qquad (2)$$

Thornton *et al*[4] have shown that for $I_e \gg I_s$, equation (2) may be rewritten:

$$I_e \simeq I_s\left[\exp\left(qV_{b'e}/nKT\right)\right] = \frac{qAD_pP_{bo}}{W}\left[\exp\left(qV_{b'e}/nKT\right)\right] \quad (3)$$

where:

A = Area of emitter

W = Width of base region

D_p = Electron diffusion constant in the base.

P_{bo} = Equilibrium electron density in the base

q = Magnitude of electronic charge

$V_{b'e}$ = Internal base-emitter voltage drop

K/q = 86·2μV/°C

T = Absolute temperature

n = Constant dependent on injection level. Usually over most of used current range $n \simeq 1$.

n may be determined for a particular transistor type by measuring I_e against V_{BE}. At low currents the effect of $R_{bb'}$ is negligible and the slope of the curve gives the information required. For the SL301-A over the range $1\mu A < I_e < 1mA$ then $n \simeq 1$ within the limits of experimental error. A knowledge of $R_{bb'}, n$, is required if the V_{BE} of a transistor is required at some current other than that specified on the data sheet.

At high current levels voltage drop in $R_{bb'}$ causes deviation from a straight line relationship, while at low current levels depletion layer recombination may cause n to be other than 1. Although equation (3) specifically applies to homogenous base types the exponential relationship applies to non-homogenous bases. However, there is a different grouping of device constants, though, what is most important for the present application, the quantity A stays as a multiplying factor.

From Fig. 5 and equation (3) with appropriate subscripts and $n = 1$:

$$\delta_1 = KT/q \ \ln (I_{e1}/I_{s1}) + \frac{I_{e1}}{(\beta_1 + 1)} R_{bb'1} \ldots\ldots (4)$$

$$(\delta_1 - \delta_2) = KT/q \ \ln (I_{e2}/I_{s2}) + \frac{I_{e2}}{(\beta_2 + 1)} R_{bb'2} \ldots . (5)$$

subtracting equation (5) from equation (4):

$$\delta_2 = KT/q \ \ln (I_{e1}/I_{e2}) + KT/q \ \ln (I_{s2}/I_{s1}) + \frac{I_{e1}}{(\beta_1 + 1)} R_{bb'1} - \frac{I_{e2}}{(\beta_2 + 1)} R_{bb'2} \ldots\ldots (6)$$

Also from Fig. 5:

$$\frac{E_1 - \delta_1}{R_b} = I_{e1} + I_{b2} = I_{e1} + \frac{I_{e2}}{(\beta_2 + 1)} \ldots\ldots\ldots (7)$$

and:

$$\frac{E_2 + \delta_2}{R} = I_{e2} \ldots\ldots\ldots\ldots\ldots\ldots\ldots (8)$$

Equations (7) and (8) yield:

$$I_{e1}/I_{e2} = \left[\left(\frac{E_1 - \delta_1}{R_b}\right)\left(\frac{R}{E_2 + \delta_2}\right) - \frac{1}{(\beta_2 + 1)}\right] \ldots (9)$$

Hence substituting in equation (6):

$$\delta_2 = KT/q \ln\left[R/R_b \cdot \frac{E_1 - \delta_1}{E_2 + \delta_2} - \frac{1}{(\beta_2 + 1)}\right] + KT/q \ln (I_{s2}/I_{s1}) + \frac{I_{e1}}{(\beta_1 + 1)} R_{bb'1} - \frac{I_{e2}}{(\beta_2 + 1)} R_{bb'2} \ldots\ldots (10)$$

Now if two transistors are manufactured close to each other on the same slice of semiconductor it is reasonable to suppose that the dominant differences between their I_s's will be due to masking errors producing different values of emitter area, i.e. it can be assumed that no significant differences arise due to inequalities in the diffusion process:

$$I_{s2}/I_{s1} = A_2/A_1 \ldots\ldots\ldots\ldots\ldots (11)$$

If it is arranged that $I_{e1} = I_{e2}$ then:

$$R_b = \frac{(E_1 - \delta_1)}{(E_2 + \delta_2)} \cdot \frac{(\beta_2 + 1)}{(\beta_2 + 2)} \simeq \left(\frac{(E_1 - \delta_1)}{E_2}\right)\frac{(\beta_2 + 1)}{(\beta_2 + 2)} \ldots (12)$$

since $\delta_2 \simeq 0$.

Under the conditions of equation (12) the actual value of δ_2 is given by:

$$\delta_2 = \Delta = KT/q \ln (A_2/A_1) + I_e\left[\frac{R_{bb'1}}{(\beta_1 + 1)} - \frac{R_{bb'2}}{(B_2 + 1)}\right] \ldots\ldots\ldots (13)$$

Δ is specified on the data sheet for a specified I_e. For operation at some other I_e, Δ will be different because $R_{bb'}/(\beta + 1)$ is a function of current and not identical for both transistors. For a 10 per cent difference between the terms $R_{bb'}/(\beta + 1)$ for each transistor the second term in equation (13) contributes an amount $\simeq \frac{R_{bb'}}{10(\beta + 1)} \times I_e$.

In the case $R_{bb'} = 250\Omega$, $\beta = 50$(av), this amounts to 0·5mV for each 1mA emitter current different from that given at the specification point.

Letting $R_b/R = x$, then for $E_1 \gg \delta_1$:

$$(\partial x/x) \simeq -(\partial\delta_1/E_1) \ldots\ldots\ldots\ldots\ldots (14)$$

Thus a small uncertainty in δ_1, does not mean a greater change in the required resistor ratio.

(1(b)) Variation of δ_2 with respect to (R_b/R)

Using equation (10) with $R/R_b = y$, assuming $E_1 \gg \delta_1$, $E_2 \gg \delta_2$, $\beta_2 \gg 1$, and ignoring terms in $R_{bb'}$:

$$d\delta_2/dy \simeq +(KT/q) \cdot (1/y) \ldots\ldots\ldots (15)$$

i.e. a 10 per cent change in y means a change in δ_2 of about 2·6mV at $T = 300°K$.

(2) Variation of δ_2 with T

An exact calculation of $(d\delta_2/dT)$ is complex. However, an engineering approximation may be based on the following argument. The emitter currents of VT_3 and VT_4 are sensibly circuit-determined. Hence it is assumed that changes in δ_1, δ_2 have an insignificant effect on these currents; this fact is used to calculate the required variation.

Thornton *et al*[3] show that equation (3) may be rearranged in the following form to show the temperature dependence of I_e:

$$I_e \simeq K\,AT^c \exp\left[-(V_{go} - V_{b'e})\,q/nKT\right] \ldots . (16)$$

where K = A constant, independent of T.

c = A material constant ($\simeq 3$ for Si)

V_{go} = Apparent energy gap at $T = 0°K$.

Now:

$$dI_e = \left.\frac{\partial I_e}{\delta T}\right|_{V_{b'e}} dT + \left.\frac{\partial I_e}{\partial V_{b'e}}\right| dV_{b'e} \ldots\ldots (17)$$

If $dI_e = 0$

$$\frac{dV_{b'e}}{dT} = \frac{-\left.\frac{\partial I_e}{\partial T}\right|_{V_{b'e}}}{\left.\frac{\partial I_e}{\partial V_{b'e}}\right|_T} \ldots\ldots\ldots\ldots\ldots (18)$$

Evaluating equation (18) using equation (17):

$$\frac{dV_{b'1e1}}{dV} = -\left[\frac{V_{go} - V_{b'1e1}}{T} + \frac{cnKT}{b}\right] \ldots\ldots (19)$$

and thus:

$$d/dT\ (V_{b'1e1} - V_{b'2e2}) = \left(\frac{V_{b'2e2} - V_{b'1e1}}{T}\right) \ldots\ldots (20)$$

This will be of the order of microvolts per degree Celsius. To this should be added the variation of:

$$\left[\frac{R_{bb'1}}{(\beta_1 + 1)} - \frac{R_{bb'2}}{(\beta_2 + 1)}\right]$$

with T, but this will be small also, since for the type of matched transistors considered, operated in the same temperature environment, the $R_{bb'}$'s and β's will track with temperature.

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Amplitude Density Distribution Function Measuring Equipment

By D. W. H. Hampshire*, B.Sc., A.M.I.E.E.

This article describes a very simple method of measurement of the amplitude density distribution function of any waveform using a minimum of equipment. Measurements showing the power of this method in estimating very small amounts of distortion are given.

(Voir page 695 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 702)

WHEN carrying out statistical investigations into the properties of continuous waveforms it is sometimes desirable to measure the amplitude density distribution function (a.d.d.f.) This is the fraction of the total time that a voltage waveform has values in the range $V \pm \delta V$, as a function of V. This article describes a very simple method of making these measurements with the minimum of special apparatus and using other instruments normally available in the laboratory.

Circuit Description

This system is based upon the use of an oscilloscope, a phototransistor and a Schmitt trigger circuit, the layout of which is shown in Fig. 1. An opaque mask is mounted across the face of the oscilloscope and a phototransistor, VT_1, is mounted so that any light shining through a small hole in the mask strikes its sensitive area. The resistor between base and emitter is added to reduce the effect of phototransistor leakage current. This of necessity reduces its photo-sensitivity and the value is something of a compromise. If the longest period for which the waveform is in the range $V \pm \delta V$ is known to be less than about 0·01sec, then it would be advantageous to replace this resistor with an inductance of 1H or more. However, as the basic device was to be as versatile as possible it was left as a resistor. To increase the sensitivity, so that it is not necessary to operate the oscilloscope at harmful brightness levels, the ouput from the photocell is amplified by VT_2 before being fed to the Schmitt trigger circuit VT_3 and VT_4. Conditions are arranged so that with no light falling on the phototransistor VT_4 is cut off, but with the phototransistor illuminated VT_4 passes a fixed current. This current is determined by the resistors in the circuit and is independent of the characteristics of the transistors and the brightness of the tube. This current is passed through an ammeter, the reading of which will be the mean value of the current passed by VT_4 and will thus give a measure of the fraction of the time for which the phototransistor is being activated by light from the face of the c.r.t.

Setting up Procedure

It is first necessary to adjust the position of the phototransistor and the conditions in the circuit to account for variations in leakage currents of VT_1 and VT_2. This can conveniently be carried out in the following manner, provided the oscilloscope is a double beam type.

The Y shift is adjusted until one of the beams transverses the hole in the mask. The second beam is used to display the output from the collector of VT_2. The photocell is adjusted in position until the deflexion produced as the spot traverses the aperture is as large as possible. It may be necessary to alter the collector load resistor of VT_2 to ensure the reliable triggering of the Schmitt circuit. This can be established by examining the waveform at the collector of VT_4. This triggering should take place without any pronounced halation of the tube. The timebase of the oscilloscope is switched off, and with no signal applied the spot is moved by means of the X and Y shift controls until it is opposite the centre of the hole in the mask. The signal is now coupled to the Y amplifier of the oscilloscope by way of the circuit shown in Fig. 2. This enables the d.c. level of the signal to be varied by a measured amount before being connected to the d.c. amplifier of the oscilloscope. The gain of the amplifier is adjusted so that a convenient displacement is obtained. This is conveniently the full height of the tube. The d.c. level of the signal is adjusted so that the tip of the vertical

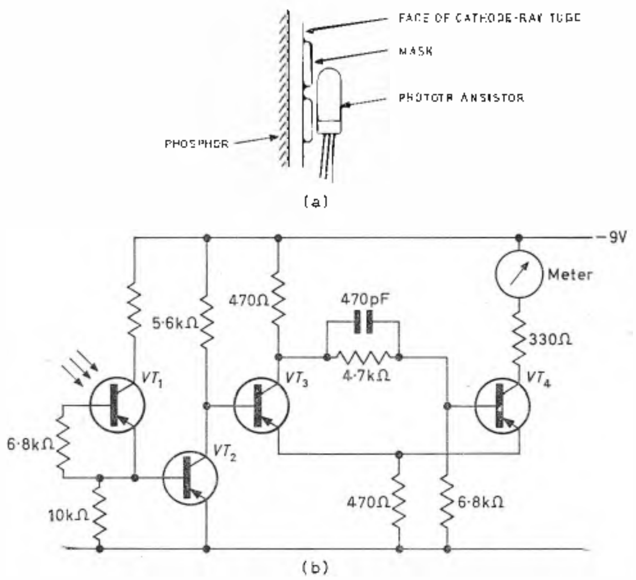
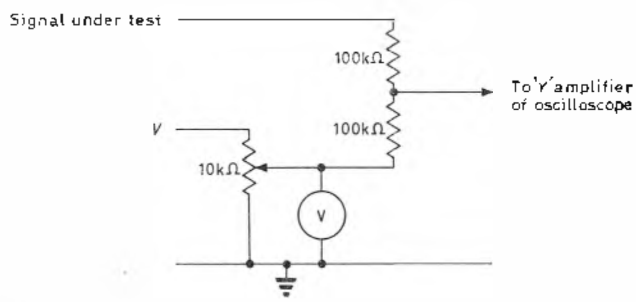


Fig. 1 (a). Mounting of the phototransistor (b). Circuit details

Fig. 2. Oscilloscope coupling circuit



* Portsmouth College of Technology.

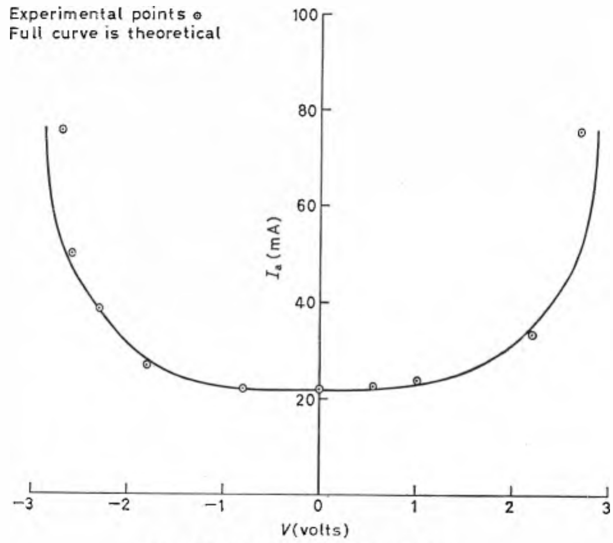


Fig. 3. A.D.D.F. of 50c/s sine wave

line just comes within the aperture of the mask. The d.c. level is noted together with the reading of the meter. The d.c. level is now progressively varied until the whole of the trace has been traversed, the two readings being noted each time. The points when plotted give the required a.d.d.f.

Test Results

In order to test the system, the functions of a number of well-known signals were measured and compared with the theoretically derived values. As can be seen from Figs. 3 to 6 there is good agreement considering the simplicity of the apparatus. These figures show the results for a sine wave, a sawtooth, a square wave and a differentiated square wave. In each case the full curve is a theoretically derived one, which has been normalized to give the same value as the experimental results, at one value of V.

It is interesting to comment that the nominal square wave analysed and shown in Fig. 5 had a mark-to-space ratio of 4·8 : 4·5, i.e. 1·07 : 1. The relative peaks on the a.d.d.f. were 5·4 : 5·0, i.e. 1·08 : 1, which is surprising agreement.

The power of this method in measuring distortion is exhibited in Fig. 7 which is the a.d.d.f. for a sine wave of the same frequency as that shown in Fig. 3. The generator used to provide the waveform shown in Fig. 3 was a decade oscillator with distortion of less than 0·1 per cent. The generator used in the second case is basically a sawtooth oscillator whose output is shaped by diodes to produce a distortion of less than 1 per cent. It can be seen that the two a.d.d.f.'s differ considerably and yet the difference in the waveform is still very small, as can be seen in Fig. 8(a) and (b).

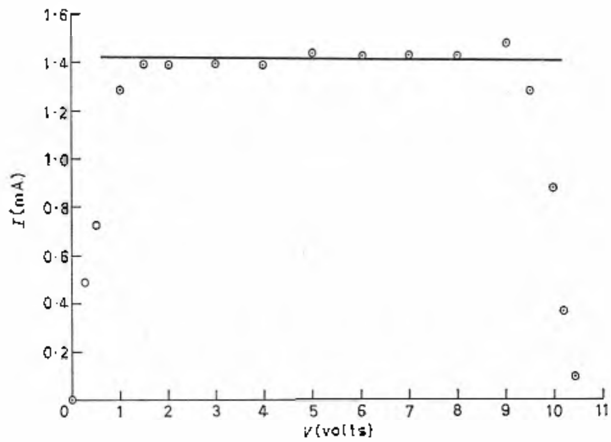


Fig. 4. A.D.D.F. of 50c/s sawtooth

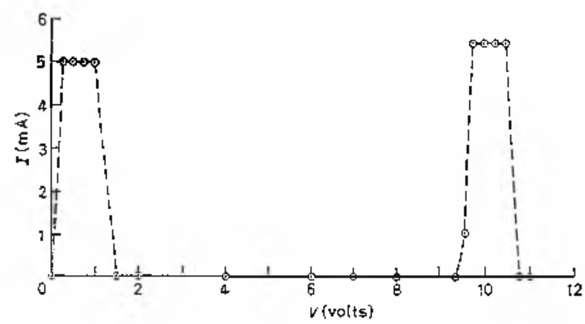
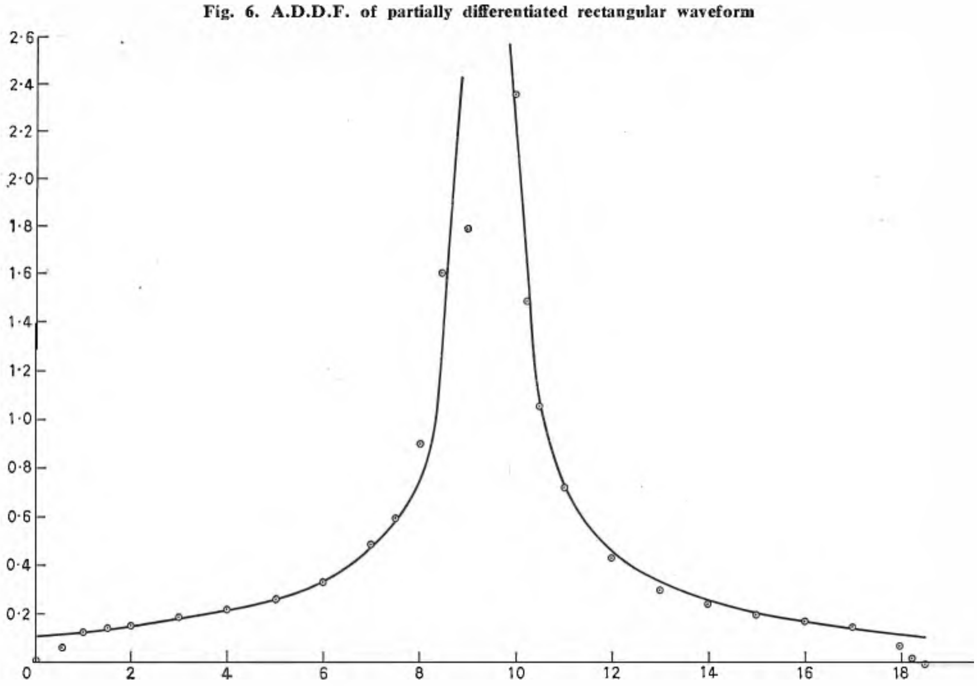


Fig. 5. A.D.D.F. of 50c/s rectangular wave; mark-to-space ratio 4·8 : 4·5

Fig. 6. A.D.D.F. of partially differentiated rectangular waveform



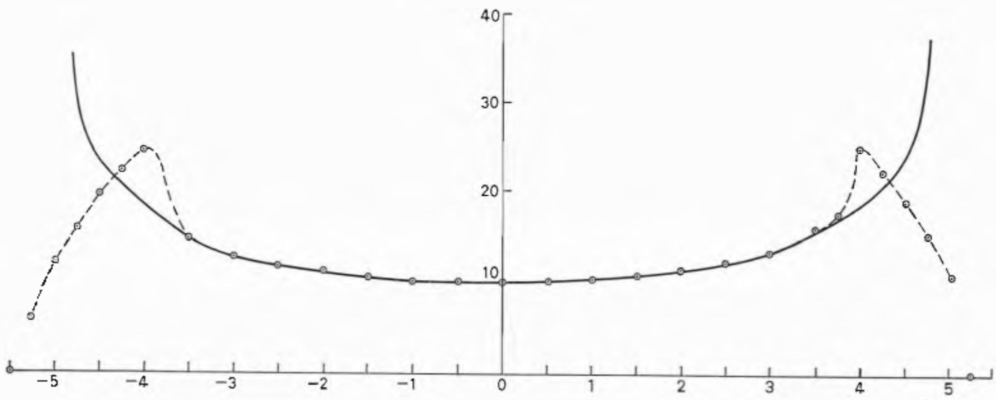


Fig. 7. A.D.D.F. of 50c/s sine wave derived from sawtooth generator and shaping diodes

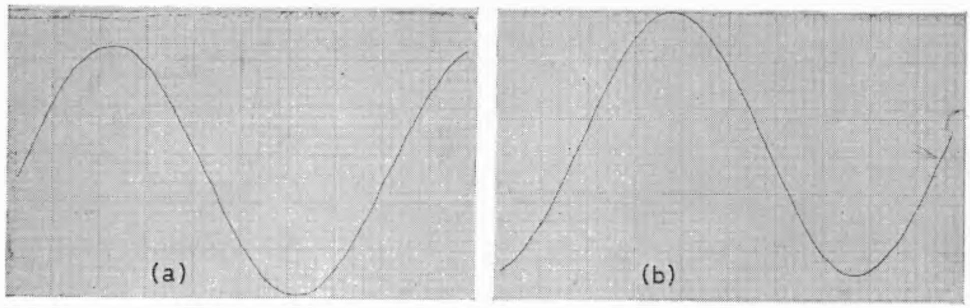


Fig. 8 (a). Waveform analysed in Fig. 7 : 1 per cent distortion
(b). Waveform analysed in Fig. 3 : 0·1 per cent distortion

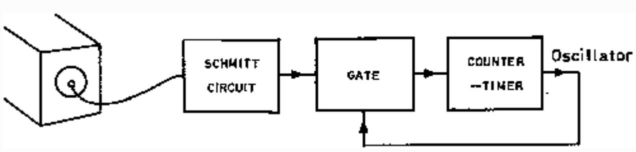


Fig. 9. Counter system for v.l.f. waveforms

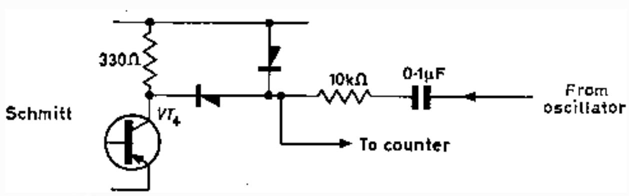


Fig. 10. Details of gate

Frequency Limitations

It was found that results were repeatable for sine waves having frequencies between 5c/s and 1kc/s. Above 1kc/s the response falls off due to 'smearing' effect of the phosphor and below about 5c/s the meter becomes difficult to read because it no longer averages the current because of its finite time-constant. However, the frequency response can be taken down in frequency almost indefinitely by using digital counting techniques as shown in Fig. 9. Instead of the Schmitt trigger being used to operate a meter, the pulse produced at the collector of VT_4 is used to open a gate, Fig. 10, which allows pulses from an oscillator to be counted for a fixed period of time. The oscillator is most conveniently that within the counter which is used to determine this time period. Under these circumstances the readings are independent of the accuracy of the oscillator.

In order to increase the frequency range in the upper direction, use can be made of a sampling oscilloscope or of one of the sampling attachments which are now available for certain oscilloscopes.

Conclusion

A very simple system has been shown to be successful in measuring the a.d.d.f. of various waveforms, having frequency components in the range 5c/s to 1 000c/s. With the addition of other pieces of standard equipment its range can be extended to both higher and lower frequencies.

Appendix

Amplitude Density Distribution Function of a Sine Wave

Consider the waveform

$$V = V_{(pk)} \sin \omega t$$

$$\therefore \quad dV/dt = V_{(pk)}\omega \cos \omega t$$

$\therefore$ Taking small changes,

$$\delta t = \frac{\delta V}{V_{(pk)}\omega \cos \omega t}$$

$$= \frac{\delta V}{\omega \sqrt{(V_{(pk)}^2 - V^2)}}$$

Hence the length of time during each quarter of a cycle that V is in the range

$$V \pm (\delta V/2) \text{ is } \frac{\omega \sqrt{(V_{(pk)}^2 - V^2)}}{\delta V}$$

Therefore the total time in each cycle that V is in the range

$$V \pm (\delta V/2) \text{ is } \frac{\delta V}{\omega/2\sqrt{(V_{(pk)}^2 - V^2)}}$$

The fraction of time per cycle is thus

$$\frac{\delta V}{\pi \sqrt{(V_{(pk)}^2 - V^2)}}$$

Hence the amplitude density distribution function for a sine wave is

$$D(V) = \frac{\delta V}{\pi \sqrt{(V_{(pk)}^2 - V^2)}} \text{ for } -V_{(pk)} \leqslant V \leqslant V_{(pk)}$$

A.D.D.F. of an Exponential Waveform

By a similar argument it may be shown that for an exponential voltage given by

$v = V_1 e^{-t/cr}$ which terminates at the voltage V_2 at time T the a.d.d.f. is given by:

$$D(V) = (\delta t/T) = \frac{\delta v}{v \,.\, \ln(V_1/V_2)}$$

SHORT NEWS ITEMS

The Fifth International Television Symposium, organized by the Swiss P.T.T. and the town of Montreux, is to be held at Montreux, Switzerland, on 22 to 27 May 1967. Among the subjects to be dealt with will be satellite communications, colour television, and the scientific applications of television including education.

The International Supplement (second edition) to the British Standard Institution's Year book has just been published. It contain details of the publications of international standards bodies—ISO (International Organization for Standardization); IEC (International Electrotechnical Commission); CEE (International Commission on the Rules for the Approval of Electrical Equipment) and CISPR (International Special Committee on Radio Interference)—complete to 31 March 1966.

Each entry gives the title of the publication, a brief summary of its contents and an indication of its relationship with the equivalent British Standard. In addition to an index there is a table giving in numerical order the British Standards which have an international counterpart, with appropriate cross references.

Copies may be obtained from the B.S.I. Sales Branch, 2 Park Street, London, W.1. Price 7/6d each (postage 9d extra will be charged to non-subscribers).

The London and Home Counties Regional Advisory Council for Technological Education of Tavistock House South, Tavistock Square, London, W.C.1, has published its 1966-67 Bulletin of Special Courses in Higher Technology, Management Studies and Commerce (Price 7/-).

This bulletin gives details of the special advanced courses held in London and the Home Counties Region which do not regularly appear on college calendars or prospectuses as part of a grouped course.

Included in the courses are automatic control, communications, computers, data processing, electronics, instrumentation, mathematics, semiconductors etc.

The Hughes International Consortium has been chosen, as already announced by the North Atlantic Treaty Organization, to build the NATO Air Defence System (NADGE) stretching from Norway to Turkey. The system proposed by the Hughes Consortium includes a wide range of British-designed equipment covering the new requirements for both primary and secondary radar. In the latter field the chosen system includes secondary surveillance radar equipment designed and manufactured by Elliott-Automation and incorporating the latest digital techniques. Elliott-Automation is already a supplier to the Ministry of Defence, the Ministry of Aviation and the Eurocontrol Organization for this type of equipment.

Elliott-Automation signal simulation equipment is also included in the Hughes Consortium proposal. The full simulation system which it is providing for the vast integrated United Kingdom Air Defence/Air Traffic Control System is now in an advanced state of construction.

Pye TVT Ltd of Cambridge has received a further contract valued at more than £500 000 for the supply and installation of transmitters to equip eight more high-power u.h.f. stations for the latest expansion of the BBC2 television service.

The transmitters, scheduled for delivery during 1968, will have peak sync outputs of 25kW (3 stations) and 10kW (5 stations). They are provisionally scheduled for installation at the BBC2 stations to serve Bristol and Somerset, Hampshire, Sussex, Staffordshire, Flintshire, Londonderry, Angus and East Lothian and Fife.

Technically the transmitters will be of interest as conventional valves are practically eliminated. The sound and vision power amplifiers will utilize vapour phase cooled klystrons and an automatic fall-back system will be employed where, in the event of a failure of either klystron, the sound and vision input signals are automatically combined and fed to the remaining tube, thus providing a lower power emergency service without separate standby equipment.

The Computer Group of the Electrical Research Association at Leatherhead has now installed an integrated computer system to enable it to carry out its studies and research into problems of power systems analysis.

Among the studies now being carried out are the following:

High Voltage Cable Networks
Inductive Interference Levels
Induction Motors
Problems of Harmonic Analysis

The National Computing Centre has placed an order valued at about £400 000 with English Electric Leo Marconi Computers Ltd for a KDF9 computer to be installed next April at the Centre's Headquarters in Manchester.

The computer will have the Egdon operating system which was originally designed for KDF9 computers supplied to the United Kingdom Atomic Energy Authority and has since been recommended by the Flowers Committee for use with similar computers installed in eight British universities.

The National Computer Centre will provide computer users with information about programs already available and will develop and sponsor the development of general programs to reduce wasteful duplication among users. These standard software systems will be made available to all users and manufacturers; for example, one of its first jobs will be to develop a billing system that all electricity boards can use. The Centre will also provide and encourage training in systems analysis, programming principles and computer applications. The first action in the education field of the Centre has been to sponsor a three-year course in systems analysis, to be held at Imperial College. The first course will commence in October of this year. Research will be promoted into programming and operating methods and the influence of these on the design of computers.

G.E.C. (Telecommunications) Ltd has received an order valued at £3·2M from the Australian Post Office for a microwave communications system to connect Eastern and Western Australia.

The link—between Perth and Adelaide—will be the longest civil communication system of its kind in the world and will be in service in 1969, by which time telephone traffic over the route will be about one million calls a year. The growth of traffic is expected to necessitate doubling the number of circuits every six years or so.

In the east, the G.E.C. link will terminate at Port Pirie, from where the traffic will be carried to Adelaide and beyond by the existing radio relay network. In the west, the system will terminate at Northam and from there will be extended to Perth on G.E.C. radio relay equipment which is already installed.

Between Port Pirie and Kalgoorlie, G.E.C. will provide one main channel carrying 600 telephone circuits, and one standby channel. Between Kalgoorlie and Northam there will be one 600-circuit main channel and a television channel with its associated sound programme, together with a common standby protection channel. All-semiconductored microwave equipment in the 2000 Mc/s frequency band will be used throughout. Eventually, the system will be capable of carrying up to four additional radio channels of the same high capacity.

Some 60 repeater stations, spaced at approximately 25-mile intervals, will be installed along the route.

The Solartron Electronic Group Ltd has received an order valued at £140 000 from the British Aircraft Corporation for the supply of airborne digital tape recorders and associated ground replay decks, to be used on prototype flight trials for the Concord project.

The equipment will operate with an error rate of less than one word in 10^6 under severe vibration and climatic stress. Using 33 tracks on 1in tape, it will record digital data generated from engine and aircraft parameters by an independent data processing system. The ground replay tapes can be computer commanded to run at any of 16 speeds.

The tape decks have been designed by Tolana, Solartron's sister company in France and a co-member of the parent international concern, Schlumberger Ltd.

Standard Telephones and Cables Ltd has received a contract valued at over £875 000 from the Ministry of Communications of the Republic of Iraq for the supply and installation of comprehensive airport services, including radio navigation, communication, air traffic control, meteorological and ancillary equipment at the new international airport at Baghdad.

The radio navigation aids include instrument landing systems (i.l.s.), v.h.f. omni-directional radio range (v.o.r.), non-directional beacons (n.d.b.), and v.h.f. direction-finding equipment. The i.l.s. is the company's STAN 7/8/9, four of which are in use at London (Heathrow) Airport.

Among the communication equipment ordered for Baghdad are h.f., v.h.f. (extended range ground-to-air and land mobile), ground-to-air selective calling systems, internal and external telephone systems, and teleprinter systems. The h.f. communications equipments are the STC DS.12 transmitter and RX.10 and RX.11 receiver which will be used for point-to-point communication between Baghdad and other Middle Eastern aeronautical centres. The extended range v.h.f. equipment consists of the DU.6 transmitter and the RX.25 receiver. Other items to be supplied are a Plessey weather radar, a meteorological broadcast centre, an intercommunication system, speech recorders, and air traffic control consoles for the control tower and a.t.c. centre.

Elliott-Automation has won a contract for an advanced new electronic 'head-up' display system, which utilizes digital computing techniques, for use in a United States military aircraft.

The system will form part of the Integrated Light Attack Avionics System (ILAAS). The first ILAAS installation will be in the new Ling-Temco-Vaught A-7 Corsair 2.

The contract has been awarded by the Sperry Gyroscope Company Division of the Sperry Rand Corporation, of Great Nock, New York, which is responsible for the development of the complex airborne navigation and attack system on behalf of the U.S. Navy.

The Elliott equipment enables the pilot to see both instrument information and the outside world at the same time without his having to move his head or re-focus his eyes. The airborne digital computer techniques are adapted from the low-cost Elliott MCS 920M microminiature digital computer incorporating internationally standardized integrated micro-electronic circuits manufactured at the company's factory at Glenrothes in Fife. Production will involve the supply of integrated circuits at the rate of at least a quarter of a million per year.

These displays are already standard equipment in British military attack and transport aircraft. Aircraft already equipped include all Hawker Siddeley Buccaneer attack aircraft, the RAF Belfast transport, and a number of development installations, including the B.L.E.U. Comet, Sea Vixen and a French Noratlas tactical transport. The first deliveries to Sweden for the Viggen attack aircraft, of which 800 may be built, have already been made.

The Elliott-head-up display for ILAAS will incorporate a wide-angle optical system and the unique Elliott all-ceramic cathode-ray tube. The optical system was developed in association with The Wray Optical Company of Bromley, which will manufacture the lenses for this contract.

ILAAS is a new approach to aircraft instrumentation, control and navigation. Instead of employing a number of independent equipments, each performing its own function independently of the others, ILAAS is the first fully integrated automation system for aircraft, which employs digital computers and a common digital language throughout. The ILAAS system is applicable to a wide range of aircraft.

Telefunken of West Germany has supplied electronic traffic supervision equipment in the new Coen Motorway Tunnel which crosses the North Sea Canal, northwest of Amsterdam.

This tunnel which is one mile long, consists of two dual lane motorways, and in each motorway thirty radar selectors have been installed at 165ft intervals and which are set at different speeds.

When a traffic jam begins to build up, data is transmitted to the control centre and at the same time a television camera is switched on in the particular tunnel section, enabling the control staff to take the appropriate action.

RCA Colour Tubes Ltd is the title of a new company set up to manufacture RCA colour television picture tubes in Britain.

The new company is owned two-thirds by RCA Great Britain Limited, the wholly owned British subsidiary of RCA, and one third by Radio Rentals Limited. It will have production facilities at Skelmersdale, Lancashire, in a new factory to be completed for the start of operations in mid-1967. The company's initial product line will include 25-inch and 19-inch rectangular colour television picture tubes similar to those now manufactured in quantity by RCA in the United States.

Two on-line moisture meters have been ordered from AEI Electronics Ltd, by the State Committee for Building in Moscow, the Soviet authority that bought four portable moisture meters last November. This £3000 order follows the display of these microwave units at the recent British Trade Fair in Moscow and AEI Electronics understand that this is an initial order representing the second stage in the Russian development programme. Success of the tests could lead to substantial orders.

In Russia asbestos cement sheet is an important building material and the AEI equipment will be used to control the amount of moisture present during the production process, improving the quality and output.

Specto Avionics Ltd—a subsidiary of Smiths Industries Ltd has been awarded a 'key' contract by the Ministry of Aviation for the supply of head-up display systems for the Hawker P1127 VTOL strike aircraft, it follows the award in January 1964 of a M.o.A. contract for the development of advanced systems; the two contracts are worth considerably in excess of £1M. A number of N.A.T.O. Air Forces are similarly interested.

This new system provides a complete presentation of primary flight instruments, information, both command and situation, projected into the pilot's view ahead. The display is focused at infinity so that the information presented and the view through the windscreen are seen by the pilot as a composite image. An important feature of the system is that it presents only the information required for the particular flight mode at any instant.

The Specto Avionics system is claimed

to be the most advanced in the world; it is based on a concept proposed by R.A.E., Farnborough and has achieved general acceptance at home and overseas by aircraft constructors and governments as primary instrumentation for their future aircraft and important contracts have been placed.

The Digital Systems Department of Ferranti Ltd has received an order worth several millions of pounds from the Ministry of Defence (Navy Department) to develop and supply training equipment in conjunction with the Admiralty Surface Weapons Establishment. This equipment includes five Ferranti F1600 general purpose computers and two cathode-ray tube display systems developed by Associated Electrical Industries Ltd.

A complete action speed tactical trainer utilizing a suite of three computers and in excess of 80 cathode-ray tube displays is included as part of this training equipment. This trainer will be used in the exercising of present and future commanders in rapidly changing tactical situations involving the mutual interaction of realistically represented ships, aircraft, submarines and weapons, with all the modern shipborne complex of radar, sonar, data automation and communication equipment.

Under the control of the computers, information in the form of labelled plan displays, letters, figures, special symbols or tabulated data is displayed on the c.r.t.s for as long as it is needed. The students and instructors communicate with the computers by means of special switches for recurring operations, and simple keyboards for general purposes. The computers work out the effects of decisions taken by the students and modify the displayed information as required. From this modified information the student must then make the next series of decisions. The power and speed of the computers ensures that the constraints imposed by the physical system being simulated are accurately reproduced and yet the instructor can at any time intervene to ensure that the right lesson is learned.

Take-off and overshoot manoeuvres of large jet aircraft require the utmost concentration from the pilots. Variations in critical speeds due to weight and altitude, undercarriage and flap selections and the need for accurate control movements, all contribute to the workload. Failure to keep to the best flight path will result in excessive take-off distance or loss of height on overshoot, and if these errors are serious, stalling or diving in might be caused.

To provide pilots with guidance indication to achieve the best possible pitch direction under these critical conditions, Smiths Industries Aviation Division has developed the STO.6 take-off and overshoot director as the latest addition to the Series 6 flight control system.

Studies have shown that by using only two simple inputs all relevant data including weight, altitude, temperature are accounted for. Pitch attitude and angle of incidence are therefore used as the basic references in the STO.6 director system. A command signal is fed to the flight director to indicate to the pilots the best angle to pitch the aircraft.

With a total weight of only 7½lb the system consists of two units, a mode selector and a duplicate monitored computer. The mode selector is a panel-mounted instrument with one knob to select either take-off or overshoot mode, and a second knob to set rotation speed on a digital read-out counter. The computer is housed in a ¼ ATR short case.

Signal levels conform to ARINC recommendations throughout, so that the take-off and overshoot system may be used with any flight director, as well as Smiths series 6 flight control system.

Flight testing of the director system has been carried out in Smiths Industries own HS.748. The system has been flown by a number of pilots who have found it simple to use, the signal being very stable and of natural form to which pilots are accustomed.

The President of the Institution of Electrical Engineers for session 1966/67 will be J. A. Ratcliffe, formerly Director of the Radio and Space Research Station of the Science Research Council.

The new Chairman of the three Divisions of the Institution have also been elected, as follows:

S. S. Carlisle: Control and Automation Division
Director of the British Scientific Instrument Research Association

P. A. T. Bevan: Electronics Division
Consultant Engineer of the Independent Television Authority

E. S. Booth: Power Division Member for Engineering, Central Electricity Generating Board.

An order for television equipment for two transmitting stations and a production studio has been placed with the Marconi Co. Ltd by the Ministry of Information of the Hashemite Kingdom of Jordan. The stations will be on the air next year, and will provide the first television service in the Holy Land.

One transmitting station, broadcasting in Band III, will be located in Jersualem, to give coverage of the Holy City and the surrounding area. The other will be on the outskirts of Amman, and will broadcast in Band I. The Jordan Television Corporation studios will be in the city of Amman itself, and the programmes will be fed to Jerusalem by a microwave link.

The order includes the supply of 10kW transmitters and associated equipment for the two transmitting stations. Four broadcast vidicon cameras, a 16mm telecine unit and electronic equipment for the control rooms will be provided for the studios, together with a film camera and suitable storage and processing equipment.

British Overseas Airways Corporation has awarded a contract worth about £2½M to Ferranti Ltd for terminal equipment as part of the airline's £33M computer system, BOADICEA (British Overseas Airways Digital Information Computer for Electronic Automation). The contract is for 700 television-type displays and 50 of the latest Argus microcircuit computers.

The television displays, which will be installed in BOAC offices and stations throughout Europe and North America, will be linked directly with the central IBM BOADICEA processors in London.

On the screen, orange-yellow letters on a matt-black background will provide information on flights and on seat availability at a rate of 333 characters/sec in response to interrogations from a silent keyboard.

Essentially the displays will accept passenger reservations, confirm them, and automatically record the consequent reduction in seats remaining on that flight. They will also provide visual information on maintenance and many other facets of BOAC's operation. All this will be done without paperwork—the information being fed into and stored in the BOADICEA system.

BBC-2 trade test transmissions are now transmitted at the following times daily (except Sundays) from all BBC-2 transmitters.

0900-1200
1400-1915 approximately

The gap from 1200 until 1400 is necessary for the maintenance and testing of transmitters and programme distribution circuits.

The extension of trade tests to 1915 will provide an additional period for the radio trade for the purposes of installation adjustment and demonstration of receivers.

During the trade test periods, subject to programme commitments and any alterations that may be necessary because of engineering maintenance, test card will be radiated continuously and will be accompanied by sound as follows:

0900-0904 440c/s tone

0904-0905 no sound

0905-0930 recorded music

This sequence will be repeated until 1200 and again from 1400 until 1915 approximately.

NEW BOOKS

Introduction to Laser Physics

By A. Lengyel. 311 pp. Med. 8vo. John Wiley and Sons Inc. 1966. Price 68s.

IN spite of the prodigious flow of letters and papers relating to lasers and their applications, there are very few books on this topic and Professor Lengyel's 'Introduction to Laser Physics' is a welcome addition. It is an expanded and improved version of his earlier volume entitled Lasers[1].

The new book fills a gap which existed between Heavens' short monograph[2], still one of the best in its class, and Birnbaum's[3] more elaborate exposition. Although the general physical background as presented by Lengyel and Birnbaum is somewhat similar, the more recent publication date of Lengyel's book made it possible for him to include such topics as electron beam pumping, dissociative excitation and the development of ionized noble gas and molecular transition lasers.

The book starts with a brief discussion of Planck's radiation law, spontaneous and stimulated emission, spatial and temporal coherence and the basic ideas of quantum mechanics necessary to the proper understanding of atomic spectra. For most electrical engineers this very lucid chapter is likely to constitute a welcome addition to the book, even though the author does not use the m.k.s. system of units everywhere. Chapter 2 deals specifically with the physical theory of the laser and covers such topics as the derivation of the threshold condition, the rate equations and the radiation modes; here the problems associated with losses, mechanical tolerances, mode competition and line width are also considered. Solid-state, fluid-state and gas lasers are respectively discussed in Chapter 3-5. It starts, appropriately enough, with a fairly detailed description of the ruby laser which is followed by the introduction of four level lasers containing rare earths or uranium ions. Semiconductor lasers are also considered, logically enough, under the heading of "Solid-State Lasers", the discussion starting with the description of a GaAs laser. Fluid-state lasers are considered largely in terms of rare earth chelates, although the general properties of such lasers are briefly outlined. This is followed by a fairly extensive treatment of gas lasers, starting with the most popular of them all, the He-Ne laser. Other gas lasers, including the more recent types based on dissociation excitation, ionization processes and molecular transitions are also considered. Chapter 6 deals with the irregularities of the output from pulsed lasers ("spiking") and the discovery of giant pulses, using externally controlled optical shutters and bleachable absorbers. The non-linear phenomena, such as harmonic generation, the stimulated Raman effect and the problem of multiple-phonon absorption are finally discussed in Chapter 7. A very brief summary of laser applications is then presented in Chapter 8.

On the whole the book strikes the right balance between excessive specialization and dangerous superficiality, although, as the title implies, it is primarily concerned with the physical side of the problem, the constructional and engineering aspects being largely ignored. However, very much to his credit the author of the book decided to consider, in simple language, some of the more specialized topics, e.g. light coherence and the basis of Russell-Saunders and Racah notation in spectroscopy. This greatly enhances the value of the book to electrical engineers who, reasonably enough, may not necessarily be familiar with such topics.

In the reviewer's opinion the book is well suited to the requirements of most electrical engineers and it can be recommended as a good introduction to the physical principles governing the rapidly developing field of laser technology.

P. A. LINDSAY

1. LENGYEL, B. A. Lasers, Generation of Light by Stimulated Emission. (John Wiley & Sons Inc., New York, 1962).
2. HEAVENS, O. S. Optical Masers. (Methuen & Co. Ltd, London, 1964).
3. BIRNBAUM, G. Optical Masers. Advances in Electronics and Electron Physics, Supplement 2. (Academic Press, New York, 1964).

Applied Underwater Acoustics

By D. G. Tucker and B. K. Gazey. 244 pp Crown 8vo. Pergamon Press. 1966. Price 27s. 6d.

THE Cumulative Index to the Journal of the American Acoustical Society for the ten years 1949 to 1958 required only five pages to list the contemporary literature on 'Underwater Acoustics', whereas the index for the five years 1959 to 1963 requires 17 pages, an above-average example of the literature explosion. The greatly increased interest is due to the development of long range signalling systems, weapon detection systems, the glamour of oil and gas exploration and the widespread use of the more mundane fish finding devices.

Unfortunately for the reputation of British engineers, very little of the technical literature has originated in England though a great deal of work is in progress. Outside the Service establishments, Birmingham University is the only organization actively interested in the problems. 'Underwater Acoustics' is a Pergamon Press book covering the material presented at the University to final-year and post-graduate students in acoustics under the guidance of Dr. D. G. Tucker. It follows the standard pattern of University based texts of presenting the basic principles of the subject with passing reference to actual practice.

It is divided into six chapters, the first being a general survey of underwater communication systems. The second chapter deals very concisely with the basic problems of any communication system, the factors that control bandwidth requirements and the limits to the achievable range set by the deterioration in the signal-to-noise ratio.

Two chapters are devoted to the propagation of sound and the special problems of the transmission of sound in water. Transducer devices for transferring the signal between the electrical circuits and the sea are then discussed in a long chapter, followed by the final chapter dealing with the fundamental factors that control the directivity of an array of transducers, a subject to which Dr. Tucker has made many significant contributions. The achievement of adequate and controllable directivity is always a problem in free space communication systems but, because of the relative proximity and the roughness of the boundaries to the media in which the signal is propagated, the problem of directivity presses more heavily on the underwater system designer than it does on others.

The book easily achieves its purpose as a clear presentation of the principles of underwater communication systems. It is unique as a lecture text but it is also exceedingly useful to the practising engineer as a rapid source of reference to information on the basic problem of underwater communication.

Each chapter ends with a list of relevant references and the book includes a series of problems and solutions that are of real value to the student working alone.

The book is well printed, the line diagrams are excellent and the book can be recommended without hesitation to any student or engineer working in the field.

J. MOIR

Solid State Communications

Edited by J. R. Miller. 365 pp. Crown 4to. McGraw-Hill. 1966. Price £5.

'SOLID State Communications' is not a text-book which presents its teachings in a series of well-ordered, logically arranged chapters. In fact the twenty-two chapters have only too clearly been written by eleven different authors. Although all written around the application of solid-state components to communications, there is apparently little cohesion between adjacent chapters and no development of a theme is apparent.

Neither is this a reference book in the generally accepted sense. One essential for any reference book is a good index. A text of 357 pages can hardly be adequately indexed in six and a half pages and when put to the test on a few random samples, the index was, as might be expected, not very successful.

The Preface gives a clue that "The objective of the work is to give the communications circuit designer as much useful and current information as can be presented in a work of less than 400 pages". Much information is indeed crammed in, but whether eleven consecutive pages of undefined graphs (pages 19-29) on which the grid lines are completely omitted, and whether when one of these (Fig. 26) has no label on one of the axes, can be put to any use by a circuit designer is problematical.

Other misprints occur. For example page 5 obviously starts in mid-sentence but the first part of the sentence is still missing, at any rate it has disappeared from the previous four pages.

Among the other editorial points, the "Tentative Data Sheet" on page 199 reproduces printing the size of which will try most eyes and yet a large amount of white space surrounds the figure.

It is doubtful if even the most enthusiastic do-it-yourself circuit engineer would attempt to build himself a 2Gc/s cavity from the drawings on the last few pages but what other purpose can be found for these fairly detailed, dimensioned, production drawings?

Like the curate's egg there are good parts and several chapters do provide useful information both on theoretical and practical grounds. It is a pity that as much thought and care which have obviously been given to those chapters has not been extended to the rest of the book.

R. A. BONES

Audio Systems

By J. L. Bernstein. 409 pp. Demy 8vo. John Wiley & Sons Inc. 1966. Price 60s. (cloth), 34s. (paper).

IN January 1956 a book of a similar nature, by Maurice Rettinger (also of RCA) was reviewed in this journal. Now, ten years later, we find an extraordinarily similar book and this is hardly surprising considering that there have been virtually no new developments in the principal subject matter during this period. Mr. Bernstein (again, of RCA), is a teacher, whereas Mr. Rettinger is a practising engineer. For this latter class of person, should they have Mr. Rettinger's book, they will not need Mr. Bernstein's. But this author has a neat and deft touch in explaining knotty calculations with plenty of worked examples—useful ones, too—plus many problems for the reader to solve.

Naturally enough, since the whole field of audio has been thrashed out by so many authors, Mr. Bernstein has concentrated on his forté—impedance matching and mismatching, attenuators, equalizers and multiple mixing of all kinds. He also brings good commonsense to bear on loudspeaker crossover network systems. These matters occupy 181 pages out of 400 odd. No one knows more about these subjects than RCA, with their vast experience of recording and broadcasting over the notoriously difficult United States.

At the same time, it is difficult to place this book in any particular category. Microphones are dismissed in a bare 6 pages, but there is quite a good chapter on amplifiers and a little information on tape recording.

It seems most likely to appeal to the student of line communication, and if this is the correct viewpoint, then it provides a wealth of information on how to solve all kinds of multiple input and output circuits. To this extent, the price of 60s. is justified, but for readers interested in other aspects of this art (e.g., room acoustics, calibration of required loudspeaker powers, the effect of standing waves etc.), then it would hardly be a profitable purchase. The book is excellently written and the production is up to the very high standard associated with Wiley books.

A. DOUGLAS

Semiconductors: Vol. 1 Physics and Electronics

By E. J. Cassignol. 310 pp. Med. 8vo. Philips Technical Library. 1966. Price 72s.

THE author intends this volume to be the first of a series of three, the other two being devoted to transistor circuits. The present volume consists of two halves, the first concerned with the physics of semiconductors and devices and the second with transistor characteristics and equivalent circuits.

Although some other semiconductor devices are mentioned, there is no serious attempt to discuss any device besides the junction transistor, except for a section on the tunnel diode which seems a little out of place. This restriction in subject matter allows the author to describe the classical theory of diode and transistor action in some mathematical detail and to discuss the various equivalent circuits for the transistor at length, even repetitively.

However it is disappointing that the author makes so little attempt to relate the characteristics of the transistor to its construction. The author implies that all transistors behave in accordance with the analysis he has presented, which is appropriate for an alloy transistor. He mentions some of the more recent designs of transistor but ignores the fact that these incorporate a lightly doped collector which affects the operation of the device, particularly in saturation. Most of the analysis is valid for all transistors, but the reader should have been warned about departures from the elementary theory. In fact there are occasions throughout the book where bald statements are made that are only true in a restricted range, and which could cause confusion, as the opposite condition is quite frequently encountered in practice. These simplifications could be justified if the presentation was consistently lucid, but the explanations of diffusion and tunnelling are muddled, the latter being discussed under the Zener diode, and the tunnel diode as though there were no connexion between them.

When the author gets on to the electronic sections he appears to be on firmer ground, and the discussions of equivalent circuits, biasing and thermal runaway are straightforward. Many formulae are derived and quoted which would be useful for reference purposes, although the student, for whom the book is primarily intended, would find it boring to read.

As a whole the book must be criticized as an unbalanced account of the physics and characterization of semiconductor devices. The omission of the central theme, the structure of the device, also results in a lack of vitality, because it precludes a discussion of the approximations made in the analyses. Nevertheless, much of the mathematics in the book is well presented and may be useful for reference, particularly in connexion with the circuit volumes of the series.

J. R. A. BEALE

Semiconductor Circuits: theory design and experiment

By J. R. Abrahams and G. J. Pridham. 310 pp. Pergamon Press. Crown 8vo. 1966. Price 30s.

Both authors of this book have spent a number of years teaching engineering undergraduates and senior technicians and their book is concerned with the design of transistor circuits.

It is divided into three main parts, Part 1 deals with the physical theory of semiconductors and Part 2, the major portion, with complete circuit designs while Part 3 deals with some 20 demonstrations or experiments.

Ultra High Frequency Propagation

By H. R. Reed and C. M. Russell. 562 pp. Chapman & Hall. Crown 8vo. 1966. Price 30s.

A second edition of this book, originally published in America in 1953, has now appeared in paperback form. The basic material of the first edition has been retained but certain theoretical derivations have been redeveloped.

Advanced Electric Circuits

By A. M. P. Brooks. 186 pp. Pergamon Press. Crown 8vo. 1966. Price 17s. 6d.

Based on lectures given to undergraduates, this book is designed to extend the range covered by the author's earlier book "Basic Electric Circuits" so as to include all the circuit analysis required for honours degrees in Electrical Engineering where the syllabus is similar to that required for The Electrical Sciences Tripos at Cambridge University.

A number of worked examples is included.

Fundamentals of Modern Quantum Physics

By G. Heber and G. Weber. 137 pp. Demy 8vo. Asia Publishing House. 1966. Price

This is an English translation of the author's book written in German and published in 1959 entitled, "Grundlagin der modernen Quantenphysik".

Ferroelectric Ceramics

By M. Déri. 96 pp. Med. 8vo. Maclaren & Sons Ltd. 1966. Price 36s.

This is an English translation of the author's book written in Hungarian and published in Budapest in 1963 under the title "Seignette-elektromos Kerámiai Anyayok".

Electronic Time Measurements

Edited by B. Chance. 538 pp. Demy 8vo. Constable & Co. Ltd. 1966. Price 24s.

The Dover Edition of this book is an unabridged and unaltered republication of the book first published in 1949 by McGraw-Hill, and which appeared as Volume 20 in the Massachusetts Institute of Technology Radiation Laboratory Series.

Klystrons and Microwave Triodes

By D. R. Hamilton. 533 pp. Demy 8vo. Constable & Co. Ltd. 1966. Price 24s.

The Dover Edition of this book is an unabridged and unaltered republication of the book first published in 1948 by McGraw-Hill, and which appeared as Volume 7 in the Massachusetts Institute of Technology Radiation Laboratory Series.

Technique of Microwave Measurements

Vols. 1 and 2. Edited by Carol G. Montgomery. 939 pp. Demy 8vo. Constable & Co. Ltd. 1966. Price 16s. each

The Dover Edition of these two volumes is an unabridged and unaltered republication of the book first published in 1947 by McGraw-Hill, and which appeared as Volume 11 in the Massachusetts Institute of Technology Radiation Laboratory Series.

Radar Aids to Navigation

By J. S. Hall. 389 pp. Demy 8vo. Constable & Co. Ltd. 1966. Price 22s.

The Dover Edition of this book is an unabridged and unaltered republication of the book first published in 1949 by McGraw-Hill, and which appeared as Volume 2 in the Massachusetts Institute of Technology Radiation Laboratory Series.

Employment Problems of Automation and Advanced Technology

Edited by J. Stieber. 479 pp. Demy 8vo. McMillan & Co. Ltd. 1966. Price 90s.

This book contains the proceedings of a conference held at Geneva in July 1964 by the International Institute for Labour Studies. It is divided into five parts, each of which deals with the problems connected with automation and advanced technology of interest to economists, labour and management organizations.

An appendix is included consisting of a background survey paper on the literature on the social and economic effects of technological change.

Introduction to Variational Methods in Control Engineering

By A. R. M. Noton. 122 pp. Demy 8vo. Pergamon Press. 1965. Price 30s.

This monograph has been written to provide an introductory text for engineering research workers and for students, and explains the application of the Calculus of Variations, Pontryagin's Principle and Dynamic Programming in the design of automatic controls.

Packaging Directory

372 pp. Demy 4to. Tudor Press. 1966. Price 63s.

The seventeenth edition of this Directory includes additions to the sections covering the list of manufacturers, Buyers' Guide, Trade Names and the foreign firms with British agents.

Microwave Receivers

By S. N. van Voorhis. 617 pp. Demy 8vo. Constable & Co. Ltd. 1966. Price 24s.

The Dover Edition of this book is an unabridged and unaltered republication of the book first published in 1948 by McGraw-Hill, and which appeared as Volume 23 in the Massachusetts Institute of Technology Radiation Laboratory Series.

Research in Electric Power

By P. Sporn. 64 pp. Crown 8vo. Pergamon Press. 1966. Price 7s. 6d.

This book contains the three lectures given by the author to the Cornell University in May 1965.

Radio Circuits

By W. E. Miller. 228 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 35s., paper 25s.

The fifth edition of this book, originally published in 1941, has been completely revised to include the changes brought about by the use of transistors in present-day radio receivers and the further development of frequency modulation.

Vacuum Tube Electronics

By R. H. Krackhardt. 115 pp. Med. 8vo. Prentice/Hall International. 1966. Price cloth 32s., paper 16s.

This introductory textbook on vacuum tubes is one of the Merrill Electronics Series intended to cover the entire field of electronics for undergraduates.

Transistor Circuit Analysis and Design

By F. C. Fitchen. 412 pp. Med. 8vo. D. Van Nostrand. 1966. Price 40s.

The second edition of this book, originally published in 1960, includes several advanced techniques in this subject. The theory of feedback has been strengthened by the addition of a separate chapter on Gain Stability.

Logic circuits and noise problems are treated in more detail.

Introduction to Nonlinear Automatic Control Systems

By R. Tomovic. 172 pp. Med. 8vo. John Wiley & Sons. 1966. Price 50s.

The author of this book, which first appeared in Yugoslavia, is a Professor in the Faculty of Electrical Engineering at the University of Belgrade. The English version has been translated by P. Pignon.

Marketing Guide to U.S. Government Research and Development

By R. Rickles. 230 pp. Demy 4to. Noyes Development Corp. 1966. Price $20

Linear Integrated Circuit Fundamentals

240 pp. Demy 8vo. Electronic & Component Devices, R.C.A. 1966. Price $2

Alternating Current Circuits and Measurements

By C. I. Anderson et al. 367 pp. Med. 8vo. Prentice/Hall International. 1966. Price 63s.

This is a self-instructional manual, and companion volume to Direct Current Circuits and Measurements.

LETTERS TO THE EDITOR

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

Phase Inversion in an Overloaded Amplifier

Dear Sir,—This note is written to draw attention to a mal-function which can arise in d.c. feedback circuits which, while quite simple in explanation, can be difficult to identify when first encountered.

A common-emitter stage may be used with an emitter resistor which is comparable with the signal source impedance, or which may become so under overload conditions (e.g. one side of a long-tailed pair when the other side is cut off). If such a stage is driven into saturation by an excessive input signal the transistor becomes effectively a switch connecting the emitter and collector together, and both now follow the base potential. Thus, in the saturation region, the voltage gain of the stage is unity and there is no longer phase inversion as there is in the normal operating range.

When such a stage is used in a d.c. feedback loop in which the feedback is normally negative, it will be seen that under overload conditions, the stage no longer inverts and thus the feedback becomes positive. This can lead to an unwanted 'lock-out' state which may perhaps arise only occasionally (e.g. through switching transients). Once the nature of this fault is recognized, it is usually fairly simple to eliminate it by arranging that the stage cannot be overdriven.

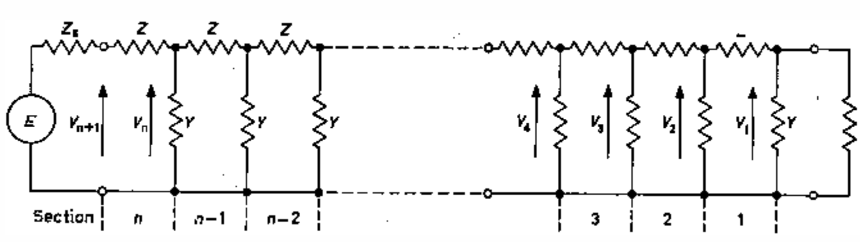
Yours faithfully,

P. C. Pugsley,
Crosfield Electronics Ltd.,
London, N.19.

Voltage Distribution Along an n-Section Uniform Ladder Network

Dear Sir,—If the number of sections in the ladder network shown in Fig. 1 is infinity, the network can be treated as transmission line and the voltage distribution along the network can be computed in terms of its characteristic impedance and propagation constant[1].

Fig. 1. *n*-section uniform ladder network energized from source *E*



However, this approach is not very convenient when the number of sections is not very large, and the analysis given below may be found more direct and useful.

Putting $ZY = x$, $ZY_L = x_1$, $Z_SY = x_2$, $Z_SY_L = x_3$, and $nC_r = \dfrac{n!}{r!\,(n-r)!}$

$$V_2 = V_1 + V_1\,Z(Y+Y_L) = V_1[(1+x)+x_1]$$
$$V_3 = V_1[(1+3x+x^2)+x_1(2+x)]$$
$$V_4 = V_1[1+6x+5x^2+x^3)+x_1(3+4x+x^2)$$
$$\cdots\quad\cdots\quad\cdots\quad\cdots$$
$$\begin{aligned}V_{n+1} = V_1[(1&+n+1C_{n-1}x+n+2C_{n-2}x^2+\ldots\\ &+2n-2\,C_2x^{n-2}+2n-1\,C_1x^{n-1}+x^n)\\ &+x_1(n+n+1\,C_{n-2}x-n+2\,C_{n-3}x^2+\ldots\\ &+\ldots 2n-3\,C_2x^{n-3}+2n-2C_1x^{n-2}+x^{n-1})]\end{aligned} \qquad \ldots\ldots\ldots\ldots (1)$$

Equation (1) will enable voltage along the network to be calculated in terms of its input terminal voltage.

Also:

$$\begin{aligned}E = V_1\,[(1&+n+1\,C_{n-1}x+\ldots\\ &\ldots 2n-1\,C_1x^{n-1}+x^n)\\ &+x_1(n+n+1\,C_{n-2}x+\ldots\\ &\ldots 2n-2\,C_1x^{n-2}+x^{n-1})\\ &+x_2(n+1+n+2\,C_{n-1}x+\ldots\\ &\ldots 2n-1\,C_1x^{n-1}+x^n)\\ &+x_1x_2(n+1\,C_{n-1}+n+2C_{n-2}x+\ldots\\ &\ldots 2n-1\,C_1x^{n-2}+x^{n-1})\\ &+x_3]\end{aligned} \qquad \ldots\ldots\ldots\ldots (2)$$

When the output is open-circuited, $x_1 \rightarrow 0$, and when it is short-circuited $V_1x_1 = V_2$. Furthermore if $x \ll 1$, all terms in equations (1) and (2) involving higher orders of x can be ignored.

Yours faithfully,

S. Ghosh,
Standard Telephones & Cables Ltd.,
Basildon, Essex.

REFERENCE

1. Rogers, F. E. The Theory of Networks in Electrical Communications. (Macdonald).

An Improved RC Bridge Circuit with High Selectivity

Dear Sir,—In the conclusions of the article 'An Improved RC Bridge Circuit with High Selectivity', published in the May 1966 issue, the author claims that a new bridge network has been found. But by redrawing the zero phase bridge

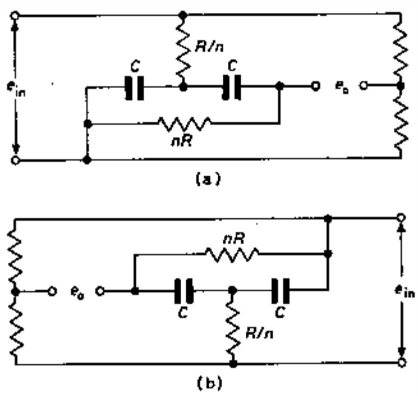


Fig. 1 (a). Zero phase bridge, Table 1(f)
(b). Sulzer's bridge, ref 8

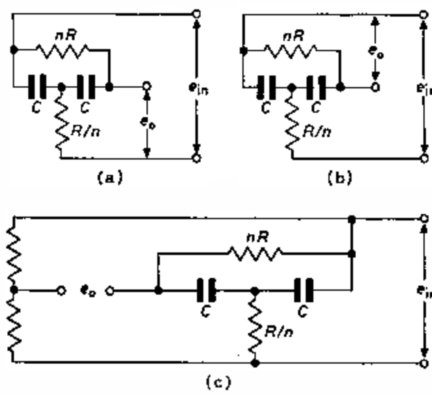


Fig. 2 (a). RC bridge-T network
(b). Zero phase shift network
(c) Zero phase bridge, or Sulzer's bridge

of Table 1(f), as shown in Fig. 1, it can easily be seen that his bridge is the same as the one described by Sulzer (Ref. 8). The results obtained—as will be expected—are also identical with Sulzer's.

Sulzer incorporated an *RC* bridged-T network in a bridge configuration by adding a resistive potential divider in parallel with the network, the output being the difference between the outputs of these two paths in parallel. The author begins with the zero phase shift network and incorporates it in the bridge configuration in the same manner as Sulzer and, probably, does not realize the fact that in the bridge configuration the difference between these two networks disappears. As a result he arrives at the same bridge as Sulzer's.

The only difference between the RC bridged-T network and the zero phase shift network (for the same component values) is that the common terminal between the input and the output is different in the two cases. (Fig. 2(a) and (b)). When these networks are incorporated in the bridge configuration by addition of a resistive potential divider in parallel, this difference is of no consequence and the same bridge is obtained with either of the networks (Fig. 2(c)).

I, however, appreciate that the author has considered one more variation of the network parameters (Table 1(d)) which leads to higher selectivity, when $n = \infty$, than that obtained in (f), and the introduction of a buffer stage to realize this condition.

I cannot agree with his statement that the twin-T network is less selective than the Wien bridge (or any other RC network) in oscillator circuits.

In RC oscillator circuits, usually, both positive and negative feedback paths are employed. One of the paths contains the RC network while the other comprises a resistive attenuator. The complete feedback network generally takes the form of a bridge. In these bridge feedback oscillators, a twin-T network provides the highest stability factor of 0·5A, where A is the gain of the unstabilized amplifier system[1]. With the bridged-T network or the zero phase shift network (even when $n = \infty$) the stability factor cannot be increased beyond this limit.

Yours faithfully,

V. B. MEHTA,
University of Manchester.

REFERENCE

1. MEHTA, V. B. Comparison of RC Networks for Frequency Stability in Oscillators. *Proc. Instn. Elect. Engrs.* 112, 296 (1965).

The author replies:

DEAR SIR,—Dr. Mehta's comments and article are most informative, and in general, I must agree with the points he brings out.

The analyses I made were for the bridge networks *per se*, and the tables were compiled in an effort to summarize the various networks found in the literature. It is true that the zero phase shift network of Table 1(f) is equivalent to Sulzer's circuit cited in my reference 8. The main point remains however, that the zero phase shift network of Table 1(d) can exhibit the greatest selectivity (as defined in my article) over the other RC networks analysed, for $n = \infty$.

If the performance of the networks in a *closed loop* configuration is analysed, the results are modified, as expected, due to the effects of feedback.

Following the procedures outlined in Dr. Mehta's article, the following results ensue for the zero phase bridge shown in Table 1(d) (Fig. A):

$$e_o/e_{in} \equiv \beta = \alpha + j\gamma$$

$$\beta = \frac{Nx^2 + jx(1 - x^2)}{(1 - x^2)^2 + x^2N^2}$$

$$N \equiv \frac{2n+1}{n}$$

$$x \equiv \omega/\omega_0 \;;\; \omega_0 = 1/RC$$

The selectivity factor: $S_F = A(d\gamma/dx)_{x=1}$ is found to be:

$$S_F = (-2A/N^2) = \frac{-2An^2}{(1+2n)^2}$$

The selectivity factor increases as n increases, and approaches 0·5A for $n = \infty$, as shown in the table below:

n	1	$\frac{1+\sqrt{2}}{2}$	1·5	2	2·5	3	10	∞
$\left\lvert\frac{S_F}{A}\right\rvert$	·222	·25	·282	·320	·348	·367	·452	·5

Since any Wien bridge configuration exhibits a maximum $S_F = 0{\cdot}25A$, (as proven in Dr. Mehta's article), the zero phase shift network of Table 1(d) can have a greater selectivity, if $n > 1{\cdot}2$.

In conclusion then, I must agree that using Dr. Mehta's analysis for an RC bridge network in a feedback loop, the twin-T has a selectivity factor of 0·5A; which is the theoretical maximum obtainable for any RC network.

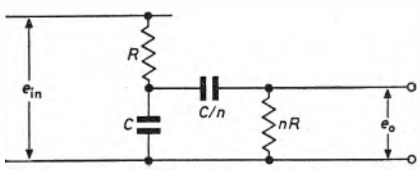


Fig. A. Zero phase bridge

However, my zero phase shift network can be more selective than the Wien bridge or bridged-T configurations. In the limit, for $n = \infty$, it will be as selective as the twin-T, since it also attains the theoretical maximum selectivity factor of 0·5A. Furthermore, the zero phase bridge lends itself better to variable frequency implementations.

S. BERNSTEIN,
The Perkin Elmer Corporation,
Connecticut, U.S.A.

Protection Circuit

DEAR SIR,—In a certain equipment it was desired to indicate whether all or none of a number of power supply isolating switches were closed, and to illuminate an appropriately coloured lamp in either case. One changeover contact was available on each switch. This problem was solved by the circuit of Fig. 1, which is self-explanatory. Such a circuit is useful whenever an indication has to be given that all or none of a number of isolating switches have been operated. Typical examples are in a complex equipment, with many different supply voltages, where it may be necessary to isolate all supplies for servicing, or in a very large equipment with interlocks, where it may be necessary to ensure that all locks are closed before switching on. In both cases a two-pole changeover switch is used, the first pole performing the switching function, and the second pole being connected as shown in my sketch. In both cases if one switch is left unoperated in the 'wrong' position, no light is shown. Naturally, such indication can be given in many different ways by relays or conventional AND gates, but I believe that my circuit is probably as simple and cheap as any, provided that relays are not used elsewhere in the interlocking.

Yours sincerely,

W. D. GILMOUR,
Glastonbury,
Somerset.

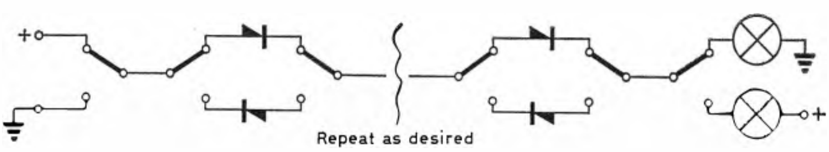
Bistable Elements for Sequential Circuits

DEAR SIR,—The E flip-flop described in my article on 'Bistable Elements for Sequential Circuits' (published in the August issue of ELECTRONIC ENGINEERING) cannot simply be made in d.c. form. The simultaneous excitation of both inputs will normally set up a critical race condition. Consequently conclusion (4) in this article is incorrect and the flip-flops in Fig. 4 must be regarded as pulsed flips-flops. Like the J-K and T flip-flops the E flip-flop must use master slave techniques in level sensitive logic circuits.

Yours faithfully,

J. J. SPARKES,
Imperial College of Science and Technology,
London.

Fig. 1. Switch indicating circuit



NEW EQUIPMENT

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

(Voir page 689 pour la traduction en français; Deutsche Übersetzung Seite 696)

ELECTRONIC COMBINATION LOCK

Dynastat Electronics Ltd, 495 Green Lanes, London, N.13

(Illustrated below)

Representing a substantial advance towards the achievement of total security, this instrument provides a new means of control not only for such obvious applications as safes, vaults, strongrooms, document stores, etc., but also on a wide field of industrial applications. For example, master security control of computers or automated machine settings, tape stores, dangerous processes—in fact, any industrial or commercial equipment which must enjoy a very high degree of security from unauthorized interference.

The instrument provides facilities for selecting a random combination of four, eight or more digits, with a memory unit to find and store the selected combination. Thereafter the condition of the controlled equipment remains secure and any attempt to release with any but the correct combination will operate the alarm circuit.

The release combination, which can be shared between two or more individuals, is selected at random at the time of locking the device and can be changed at will as desired. There is, therefore, no permanent or semi-permanent preset combination with attendant security risks.

The selected combination cannot be recalled by any process of deduction, nor will any kind of manipulation of the selector betray to any of the senses the proximity of the correct position.

The photograph shows a unit made up for demonstration purposes and employing four digit selectors. The instrument has been made up in standard modular form but can be 'tailored' to suit individual requirements. In practice, however, eight digit selectors provide a more acceptable degree of security and in fact such an arrangement gives an intruder only once chance in 10^8 of selecting the correct combination.

The selection of any other combination will not only leave the equipment completely secure, but will also operate the alarm system.

The security of the instrument cannot be breached by any form of electrical or mechanical attack, and even when it is used in positions remote from the protected equipment, cutting or short-circuiting linking wires, or feeding in any form of signal, only results in initiation of the alarm.

The instrument only uses electrical power during locking and unlocking procedures and is completely independent of mains supply during its periods of control. Mains variations or failure, therefore, have no effect on the instrument.

EE 98 751 *for further details*

PAPER TAPE READER

Electrographic, 38 Corwell Lane, Hillingdon, Middlesex

(Illustrated above right)

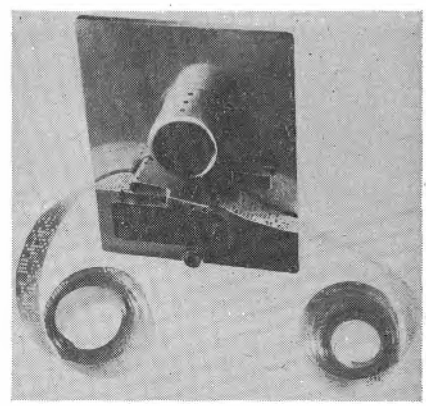
A new photo-electric punched paper tape reader is announced by Electrographic. The unit is designated the type 800 and is capable of reading 100 eight bit characters per second and stopping on any selected character.

It will accept any of the standard width tapes, 5 to 8 holes plus feed hole. The tape is moved by a sprocket driven by a stepping motor, thus ensuring positive drive and fast response.

A novel optical arrangement ensures that any paper tape may be read with high on/off ratio. A photocell is provided in the feed hole position so that accurate alignment of the tape in the read position may be determined.

The design is such that the optics cannot become fouled with dust and dirt. The code being read can be inspected visually in the read position.

A low voltage lamp with a substantial filament is employed in a very under-run



condition so its life is extremely prolonged.

The tape may be moved in either direction by simple reversal of the stepping motor drive.

A photocell buffer amplifier and motor drive board are available mounted on a chassis to fit a standard 'Imhof' printed circuit rack.

EE 98 752 *for further details*

TELEVISION DELAY LINE

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

(Illustrated on page 683)

Advantages claimed for a new pre-adjusted, ultrasonic delay line introduced by Mullard for PAL colour receivers are simplification of the receiver's decoder circuits and the elimination of in situ adjustment of the delay time.

In addition to the particular advantages of the new Mullard line, the use of a delay line in PAL receivers can help to simplify design problems in other parts of the receiver, and improve picture quality.

Production quantities of the new line, type DL1, will be available during this year, and sample quantities are already being supplied for use in development receivers.

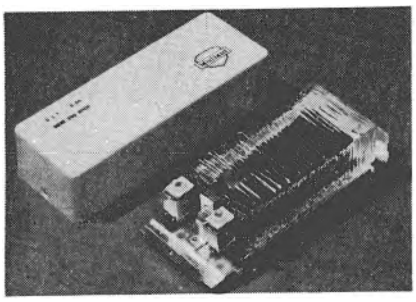
The line—a folded, reflecting, glass type—is easier to use than other lines designed for the same purpose because it gives the required delay of 63·943μsec at 4·4336Mc/s without recourse to in situ adjustment by means of additional, extraneous circuits.

Accuracy of the delay time is pre-set during manufacture, first by precision grinding of the reflecting surface of the glass block, and finally by adjusting the variable in-built inductors.

The input to the line is applied to a ceramic transducer which converts the electrical signal into an acoustic wave for transmission through the block. When reflected back to a point near the input the wave is re-converted into an electrical signal by another similar transducer. Because the line is a folded reflecting type the transducers can remain in place while the delay time is pre-set.

Because of its very low nominal insertion loss of only 10dB the DL1 will give sufficient voltage across its output terminals to directly feed a diode demodulator. The line is designed for use between sensibly non-reactive 150Ω sources.

As a further advantage to receiver manufacturers the DL1 is offered complete with circuit information to provide $(R - Y)$, $(B - Y)$ and $(G - Y)$ colour difference drive signals of constant amplitude from a simple network so



eliminating the problems associated with the spread of loss and gain between circuits.

The DL1 is fitted with connecting pins which allow it to be directly inserted into a printed board.

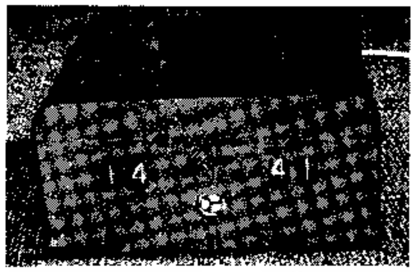
EE 98 753 *for further details*

DIGITAL CLOCK

Alma Components Ltd, Park Road, Diss, Norfolk

(Illustrated above right)

This new digital clock is designed primarily as a master programme timer. A synchronous motor supplied from the mains drives a series of Alma reed uni-selectors in cascade to provide a digital read-out in hours and minutes. Initial setting to the correct time is carried out simply by four reed push-button switches at the back of the instrument. A feature of the design is that at each and every minute in the 24-hour cycle the uni-selectors can be connected by the user to close or open an internal contact, a combination of 1 440 switch contacts, each rated at 1A 250V with a maximum power rating of 60W. The complete cycle is repeated every 24 hours or, by operating a switch, the clock can be converted for 12-hour operation. It can also be supplied having any desired time cycle, e.g. 48 hours, 60 hours, etc., or alternatively to count in minutes and seconds. Due to the flexibility in the design, special additional features can



be incorporated to meet particular applications; for example, delaying the action to allow print-out time.

The 24-hour clock is available in standard form either as a complete self-contained unit in an aluminium case or in chassis form to install in other equipment.

EE 98 754 *for further details*

CORE MATRIX ASSEMBLIES

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

(Illustrated below)

A range of inexpensive, single-plane core matrices designed for small electronic desk calculators has been announced by Mullard Ltd. Used in the stores and registers of such machines, the matrices are cheaper and smaller than semiconductor systems, and retain any data stored even when the machine is disconnected from the mains supply.

Operational stability over a wide temperature range (0 to 70°C) is achieved by the use of lithium-nickel ferrite cores. Operation at low drive current (190mA) is obtained by using 2-turn X, Y and Z windings to reduce the drive requirements and thus permit relatively simple drive and selection circuits to be used. Wiring is in accordance with the well-known MIT system.

The matrices will withstand rough handling, because the core assembly is supported in a rigid frame, with the cores secured by a special lacquer to a paper-laminated backing plate. Intended for soldering direct into printed wiring boards, the matrix pin connexions fit a 0·1in grid circuit board.

Epoxy frames capable of withstanding more humid conditions are available at a small extra cost. Matrix assembly stacks of up to four planes with series-

connected drive wires can also be supplied. The following range of planes is standard:

Type No.	*Number of cores*	*Outer dimensions*
AW3526	16×16	82×82mm
AW3527	4× 8× 8	82×82mm
AW3529	4×12×12	102×102mm
AW3531	4×16×16	122×122mm
AW3532	32×32	122×122mm

EE 98 755 *for further details*

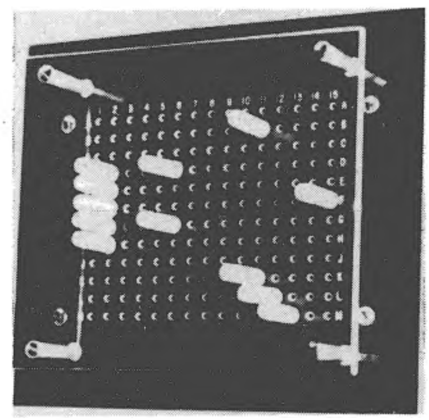
PROGRAM BOARDS

Sealectro Ltd, Farlington, Portsmouth, Hampshire

(Illustrated below)

To meet special safety requirements, Sealectro Ltd has introduced a 'Sealectoboard' equipped with a transparent cover panel that prevents unauthorized pin removal. At the same time, the protective cover provides an unobstructed view of the program and the pin arrangement.

The transparent cover panel is also provided for applications where the



program board is subject to vibration that would otherwise cause pins to shake loose from the board.

The cordless program board is available in standard and special configurations offering virtually unlimited combinations of horizontal and vertical hole electrical matrix arrangements to meet any requirements.

EE 98 756 *for further details*

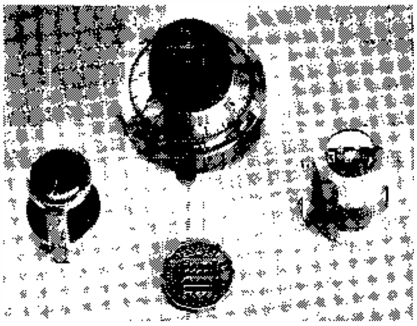
MULTI-TURN PRECISION DIALS

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

(Illustrated on page 684)

Kynmore Engineering Co. Ltd has added three new calibrated multi-turn precision dials of good appearance, manufactured by the French Atoms concern, to the Kynmore range.

The new Micro-Clockdial (left on illustration) is a ten-turn analogue dial having 50 marked divisions to each turn. It has a diameter of 0·87in (22mm) and is 1·14in (29mm) deep. The new 15-turn analogue dial, model US45 (top), has 100 marked divisions to each turn. It is 1·79in (45·5mm) in diameter and 1·20in (30·5mm) deep. The new miniature digital dial, model Flash 22 (right) is a



100-turn dial with magnified read-out having ten numbered divisions to each turn. It is 0·92in (23·4mm) in diameter and 1·39in (35·2mm) deep.

All three are designed for 0·15in (6mm) diameter shafts and fitted with brake levers, enabling them to be locked in any position.

EE 98 757 *for further details*

DIRECT WRITING OSCILLOGRAPH

Honeywell Controls Ltd, Brentford, Middlesex

(Illustrated below)

Honeywell's new "Visicorder" oscillograph 2106 has been designed and developed at the company's Lanarkshire plant to give an extremely high specification in a low cost structure, offering many features not normally associated with instruments in its price range. The compact unit has a 6in wide chart which takes records from 12 channels (plus a further two auxiliary channels).

A unique advantage in this unit is that the ultra-violet lamp can be started instantly when either hot or cold. Capable of recording a multiplicity of variables on the same chart, the 2106 also provides as standard features grid and timing lines and remote control facilities.

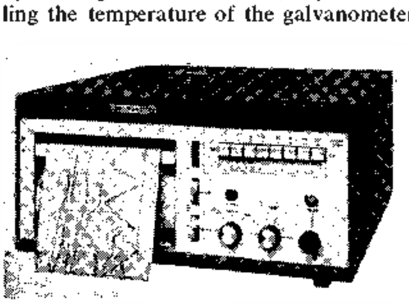
Other features include: a choice of chart speeds with gearboxes which provide an overall range of 1cm/min to 300cm/sec (0·4in/min to 120in/sec), and individual plug-in resistor boards for each galvanometer, enabling a library of input boards with different damping and matching networks to be built up, eliminating tedious and expensive setting-up time. The writing speed is greater than 125cm/msec (50 000 in/sec), and an optional record length control prevents paper wastage. The 2106 takes up to 150ft of paper, which may be narrow, down to a minimum of 3in.

Stability of performance is maintained by heating and thermostatically controlling the temperature of the galvanometer block. Any galvanometer may be inserted or removed in a matter of seconds.

A worldwide market is anticipated for this British built unit and for Honeywell's new range of series M galvanometers, which can be used with it.

The mirror of these high frequency fluid damped galvanometers is not immersed in the damping medium. Deflexion of the light image across the chart is, therefore, not limited by dispersion of the ultra-violet light in the fluid. The galvanometers combine high sensitivity and stability with flat frequency responses from d.c. to 13kc/s.

EE 98 758 *for further details*



FREQUENCY STABILITY

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated below)

Greatly improved frequency stability above 50kc/s is provided by the new Hewlett-Packard Model 8708A signal generator synchronizer.

Designed specifically to work with new versions of the HP models 606 and 608 signal generators, the new synchronizer uses phase lock techniques to provide test signals stable within two parts in 10^7 over a ten minute period. This is an improvement of more than 250 times over the stability previously available with standard signal generators operating in this frequency range.

Automatic synchronization is achieved at any frequency within the range of the improved 606B (50kc/s to 65Mc/s) or 608F (10Mc/s to 455Mc/s) signal generators. In addition to providing increased stability, the improved generators have new capability for both frequency and phase modulation and for low-distortion amplitude modulation. Both power output and modulation percentage remain constant with frequency change, the stabilized output frequency has high spectral purity, and the frequency/phase modulation capability is highly linear.

EE 98 759 *for further details*

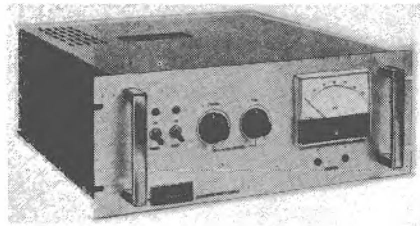
HIGH VOLTAGE POWER SUPPLIES

Brandenburg Ltd, 139 Sanderstead Road, South Croydon, Surrey

(Illustrated above right)

Latest in the new BETA range of high voltage stabilized power supplies are models 905 and 906. They both provide 400μA at up to 60kV yet are only 7in high, 17in deep and 19in wide.

Output voltage can be adjusted, by coarse and fine potentiometers, to any value between 6 to 60kV and is monitored by a front panel meter. Provision



for current metering is included on the rear panel. Model 905 has positive polarity and model 906 negative.

The extremely compact size of these power supplies is due largely to the advanced design and construction of the high voltage head. The solid state comparator amplifier is built as a printed circuit board coupled by edge connectors and the remaining larger components are supported by a robust alloy frame. Styling and construction are almost identical to models 705 and 800; the 905 and 906 are also suitable for both bench and 19in rack mounting. The performance is very similar to existing BETA models, covering all industrial and many laboratory requirements.

EE 98 760 *for further details*

VALVE VOLTMETER

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

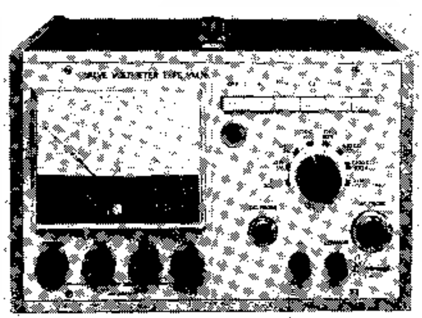
(Illustrated below)

The VM76 is the latest in the Advance range of five voltmeters and extends the range to include a.c. frequencies in excess of 1Gc/s and d.c. measurements up to 1000V with accuracies of ±2 per cent up to 100V.

Push-button selection is employed for the various functions. In d.c. operation this method can be used to select centre-zero readings and either positive or negative polarity. Resistance measurements between 0·02Ω and 500MΩ to an accuracy of ±5 per cent at midscale can be selected in a similar way. Typical input impedances are 7·5MΩ at 1kc/s and 150kΩ at 100Mc/s, both with 1·5pF shunt capacitance.

Excellent component accessibility is provided by mounting the majority of components on a single printed circuit board. Each instrument is provided with probes and connecting leads with single plug and socket connextions. Optional accessories for all voltmeter applications are available.

The VM76 has a temperature range of 0°C to +40°C and is mains operated. In line with other recently introduced



Advance instruments it is built into a robust modular case which is 2/3 standard rack width.

EE 98 761 *for further details*

ECHO SOUNDER SIGNAL GENERATORS

Elliott Instruments Co. Ltd. Dunmow, Essex

(Illustrated above)

Elliott Instruments Co. Ltd announces new versions of its echo sounder signal generators. These instruments have been specially designed for the engineer carrying out servicing work on board ship, but provide also convenient signal sources in the laboratory.

Fully transistorized, the instruments, known as model UL3, generate sine wave outputs covering the frequency range 7 to 60kc/s, and produce a signal variable from 1μV to 1V. Battery operated, they are robust and compact units, dimensions 8in × $6\frac{1}{2}$in × 2in, weight 4 lb.

Monitoring facilities are provided for gain measurements on echo sounder amplifiers, the meter in the instrument being directly connected to the amplifier output. It can indicate either 10V d.c. at 10kΩ impedance for wet paper recorders, or 500V a.c. at 200Ω impedance for dry paper recorders. The instrument can also be supplied as model UL3a without monitoring facilities.

EE 98 762 *for further details*

PACKAGED CRYSTAL OSCILLATORS

The Marconi Co. Ltd. Chelmsford, Essex

(Illustrated above right)

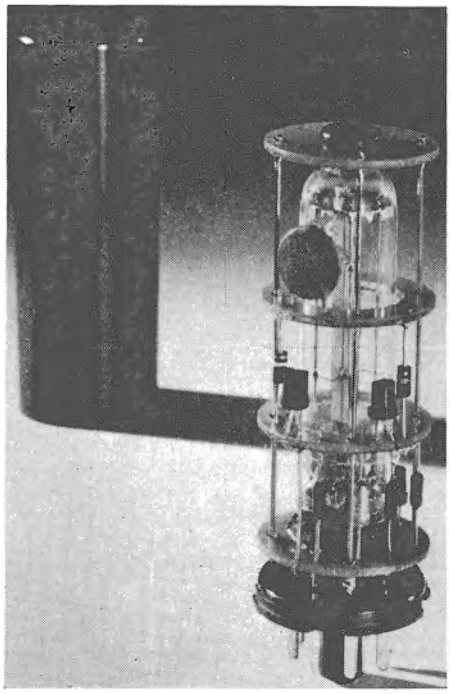
The Marconi Company has announced a new range of packaged transistorized oscillators offering design engineers a low priced, medium stability frequency source, housed in a single plug-in unit, now in full production.

A 6V, low current supply is all that is required to provide an efficient ready-to-use crystal oscillator. Mounted on International Octal or B7G valve bases, they can be provided at frequencies in the range of 1kc/s to 100Mc/s. The more compact B7G unit, type F3171, covers the range 115kc/s to 100Mc/s and has the added advantage that it can be placed in a standard Marconi F3006 crystal oven to provide increased temperature stability.

The larger Octal based unit, type F3170, covering the range 1 to 115kc/s, is contained entirely in an aluminium tubular case, 1·125in in diameter. Depending on frequency, the seated height varies from 3·125 to 5in whereas the higher frequency oscillators, with a B7G base, have a standard seated height of 3·125in. These units are also contained in a cylindrical aluminium can, but of $\frac{3}{4}$in diameter.

Stabilities better than 1 part in 10^4 are obtainable over the operating range of -20 to $+70$°C. If desired, the frequency of the units can be trimmed by means of an external variable capacitor connected between one of the base pins and earth.

Each oscillator is built around a glass encapsulated quartz crystal and has a buffer output stage which maintains a frequency stability of less than 1 part in



10^{-6} for a 10 per cent variation in the load impedance. The output impedance is 5kΩ ±10 per cent with an output voltage of 2V peak-to-peak.

EE 98 763 *for further details*

RECORDING HEAD

Gresham Lion Electronics Ltd, Magnetic Recording Head Division, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

Gresham Lion Electronics Ltd has developed a 9 track read after write magnetic recording head. This head is designed for digital applications, gives immediate verification of recorded information and is intended for use with the new IBM 1 600 bit system.

It is available with record elements on the left, for compatibility with IBM equipment, or with record elements on the right. The front face is profiled to minimize long wavelength noise and is cut back beyond the track to prevent grooving by tape edges.

The head measures $1\frac{1}{4}$in high, 1 1/32in deep and 1in wide; azimuth is held within ±·0002 radians and elevation within ·0005 radians. Gap scatter is contained within a band ·00008in wide and the gap lines are parallel to within ·0002 radians. The read tracks are 0·040in wide and the write tracks are 0·044in wide. Separation between gap lines is 0·150in. Gresham is offering a choice of read gap lengths; ·0002 or ·0001in; the latter being intended for 1 600 bit use.

Saturation current on 3M T488 tape at 300 bits to the inch and 99 per cent saturation is 66mA ±15 per cent peak-to-peak.

Electrical characteristics are as follows:

(a) Gap read ·0002in
Read: Resistance 3·7Ω; Inductance 600μH
Write: Resistance 2·3Ω; Inductance 415μH

(b) Gap read ·0001in
Read: Resistance 7·4Ω; Inductance 2·4mH
Write: Resistance 4·6Ω; Inductance 1·6mH

The write head winding is centre-tapped.

EE 98 764 *for further details*

STATIC INVERTOR

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(Illustrated below)

Industrial Instruments Ltd has added the 606/8L to its existing range of TRANSIPACK static invertors. This unit has been developed in order that the teleprinter may be included in the field of mobile communications equipment. It will supply power to operate the motor, activate the time/space relays and energize the transmitter from one 24V d.c source. This fully stabilized unit will deliver +80V and −80V (other signal voltages available) at a load of 1A maximum and it is provided with complete overload protection. It is housed in a free standing stove enamelled cabinet with easy access to the batteries and includes a built in charger.

With the use of this equipment, the teleprinter can be made truly mobile

and, therefore, used in a great many more applications, such as building, oil fields, military applications, police control of vehicles, etc.

Industrial Instruments Ltd manufacture a wide range of static invertors, which produce simulated mains of 230V a.c. at a frequency of 50c/s from d.c. sources. Typical of these is the TRANSIPACK type 606/ST/300, shown here in a Service enclosure. This unit accepts 24V from a vehicle, or equipment battery, and produces a continuously rated power output of 300VA at 230 a.c. Frequency stability is better than ± 1 per cent and the harmonic distortion between 3 and 6 per cent.

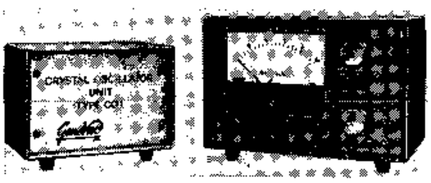
EE 98 765 *for further details*

THIN-FILM MONITOR

Genevac Ltd, Pioneer Mill, Radcliffe, Manchester

(Illustrated below)

A new quartz crystal film thickness monitor from Genevac Ltd, is designed for the measurement of the total thickness and also of the rate of deposition of vacuum-deposited thin films. To achieve this a portion of the evaporant stream is collected on a quartz crystal mounted in a suitable position within the evaporator. Using the principle of linear change of the resonant frequency with increasing mass of the monitor crystal, the device indicates the change as a displayed d.c. voltage and also the



rate of change of this voltage. Feedback from the instrument may be used to control the heat input to the evaporating source and thus the rate of disposition; by pre-setting the end point of the operation the unit can be operated as an electromechanical shutter.

The instrument can control laboratory and production depositions with a high degree of repeatability, and indicates the deposited mass over four ranges of 1, 5, 10 and 50kc/s full scale deflexion. Provision is made for backing off to zero between consecutive depositions up to a total of 60kc/s shift, when the crystal requires cleaning. The equipment is housed in two cases both 9in wide, 6in high, 6in deep; one case contains all the necessary power supplies and displays the mass deposited, the second case displays the rate of deposition and can be purchased separately. Connected to the monitor's quartz crystal through a coaxial vacuum seal is an oscillator placed close to the evaporator.

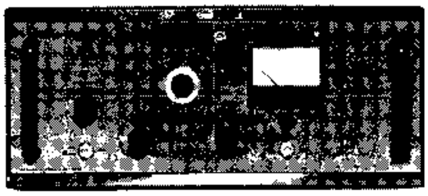
EE 98 766 *for further details*

LOW-NOISE AMPLIFIER

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustrated above right)

Brookdeal Electronics has introduced a new all-silicon transistor amplifier LA350. This is a high-performance



laboratory amplifier with an exceptional specification which is believed to be in advance of any hitherto available.

Conventionally, a good noise figure is only achieved from a discrete source impedance over a narrow bandwidth. The LA350 however has a noise figure of lower than 2dB with source impedances ranging from 800Ω to 250kΩ at frequencies above 500c/s, so that the limiting factor in most measurements is the thermal noise in the source. The novel circuits employed also extend its capabilities beyond the normal restriction of conventional low-noise amplifiers and give it the properties of a high quality general purpose amplifier: good linearity, wide frequency range (with acceptable square wave response) and fast overload recovery.

The instrument features distributed feedback attenuators which operate over successive stages so as to prevent amplifier overload and attenuate the noise by the same factor as the signal. These are operated by a single control which gives 5dB steps. The maximum gain is 100dB. The overall frequency range is 3c/s to 300kc/s; it can be reduced by high and low-pass filters which alter the 3dB points by successive factors of 3.

Other features include: an output meter which enables the instrument to be used as an a.c. microvoltmeter, switchable earthing, and the option of rack or bench mounting.

EE 98 767 *for further details*

10Mc/s PULSE GENERATOR

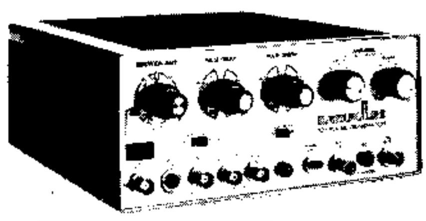
Distributed by: Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated below)

The Datapulse model 101, marketed by Dynamco Instruments in the U.K., is an outstanding book-size pulse generator, being only 3½in high, 8½in wide and 11in deep and weighing 8 lb. It offers repetition rates to 10Mc/s, a rise time of 5nsec and outputs to ± 10V.

The 101 has features normally only found on the more expensive type of pulse generator. These include single or double pulse operation, variable duration and delay to 10msec, flexible control of external triggering and gating, and 1 000 : 1 timing control coverage.

External gating capabilities include



synchronous gating for stable oscilloscope display of pulse bursts, asynchronous gating for advanced trigger outputs as an uninterrupted system clock, and coincidence gating for pulse coincident with gate input and an external trigger source.

Oscilloscope type triggering with level and slope polarity controls provides sensitivity of 250mV. The unit may be effectively triggered from any point on the positive or negative slope of the a.c. line voltage waveform. Triggering versatility also allows synchronous countdown of external trigger (rep. rate countdown) for applications requiring frequency division.

EE 98 768 *for further details*

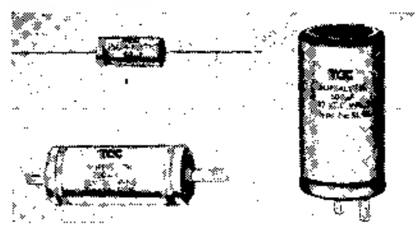
ELECTROLYTIC CAPACITORS

The Plessey Co. Ltd, Ilford, Essex

(Illustrated below)

Two new capacitor ranges from the TCC Division of the Plessey Components Group are designed to meet the demands of designers of specialist electronic equipment.

The 'Elkomold' miniature electrolytic capacitors are intended for use with transistor circuits and printed board mounting. Light in weight and small in size, the plastic cases eliminate the need for insulating sleeving. Four case sizes provide for a wide range of ratings.



The cases are suitable for vertical mounting and are coded for polarity. Construction and performance are similar to those of the TCC standard miniature electrolysis in metal cases. The plastic case is moulded in durable polypropylene and one end closed by a rubber faced phenolic disk.

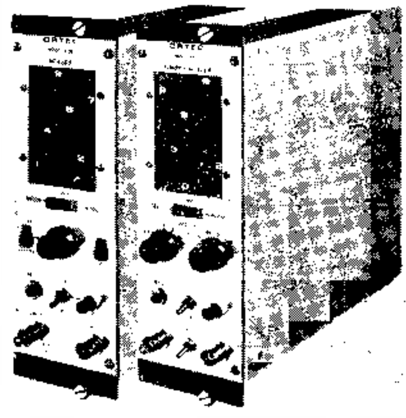
Capacitances between 1 and 1 000μF are available, with d.c. working voltages up to 350V for lower values. Capacitors can be operated continuously without derating over the temperature range $-25°$C to $+70°$C.

TCC 'Superlytic' capacitors are specially designed for long life service in professional applications. Rigorous process control and specially selected raw materials ensure a service life of up to ten years.

The operating temperature range is between $-25°$C and $+85°$C and the leakage current will not exceed 0·01CV μA or 5μA whichever is the greater. 'Superlytic' capacitors are available in aluminium cans with p.v.c. insulating sleeves.

Capacitances available are 1 to 3000μF with working voltages between 3V and 250V d.c. Can sizes vary between 3/4in × 3/8in diameter and 4½in × 2½in diameter.

EE 98 769 *for further details*



SCALER-TIMER

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

(Illustrated above)

Now available from High Volt Linear Ltd are the new ORTEC model 430 scaler and model 431 scaler-timer.

The model 430 scaler is an integrated circuit six decade scaler with a self-contained discriminator and printing output. The unit is built to the new AEC standard on Nuclear Instruments TID-20893 and occupies two positions in the 12 position model 401/402 modular system bin and power supply. Other features include a 16Mc/s counting rate, internal discriminator, positive or negative inputs and gating facilities. By including one model 432 print-out control in the system, a flexible serial output data acquisition system of up to 50 scalers is possible.

The model 431 scaler-timer is an integrated circuit six decade presettable printing scaler that can also be used as an accurate timer. The 431 conforms to the same mechanical specification and standard as the 430 above. Other features include 4Mc/s scaling rate, a tuning fork option giving 0·04per cent accuracy and gating facilities. Like the model 430 up to 50 scaler-timers can be used in conjunction with the model 432 print-out control to form a flexible data acquisition system.

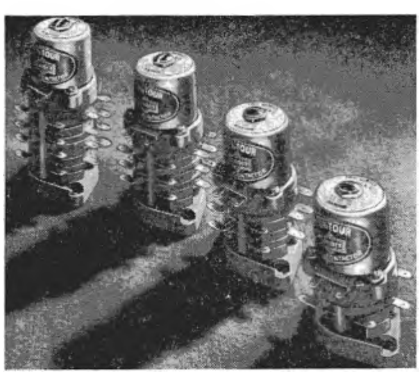
EE 98 770 *for further details*

RELAY CONTACTOR

Londex Ltd, 207 Anerley Road, London, S.E.20

(Illustrated below)

A new range of relay contactors of particular importance for machine tool and similar control applications is being manufactured by Londex Ltd, a member of the Elliott-Automation Group.



These new "Contour" type units are extremely robust, require very little panel space and comply with the requirements of BSS 775-1956, CSA and CEE specifications.

A unique feature of the design lies in the triangular construction of the fixed contact carriers which enables the same components to be arranged for either 'make' or 'break' operation merely by inverting them. The use of bridge-type contacts makes it possible to produce a single-pole changeover action by connecting one side of a 'break' contact with one side of a 'make' contact. Coils can be changed remarkably simply without affecting contact adjustment. A new coil of the same or different voltage can be fitted by removing the three screws which hold both coil cover and coil in place. The operating mechanism is self-aligning and ensures good contact seating and pressures.

Up to twelve single-pole, double-break contacts, in multiples of three, can be fitted. These can be 'make' or 'break' or a combination of both types. Typical non-inductive ratings are 15A at 250V a.c., 15A at 50V d.c.

Standard coils for 50c/s a.c. are: 6V, 12V, 25V, 115V, 380V, 415V, and 440V. Coils for other a.c. voltages and for d.c. are available.

EE 98 771 *for further details*

POTENTIOMETRIC RECORDER

A.E.I. Electronics Ltd, New Parks, Leicester

(Illustrated above right)

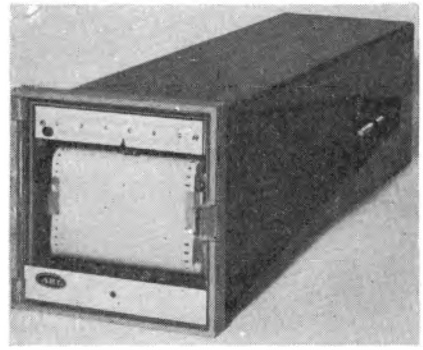
An improved version, the mark 5, of its 4in potentiometric chart recorder has been developed by AEI Electronics, for use with thermocouples, resistance thermometers, strain gauges and various signal generators.

Solid state components are used throughout to give high reliability, compactness and low power consumption. The mark 5 is available as a single continuous trace or multi-point recorder accepting signals from two, three or six channels and printing out in coloured dots. Its outstanding features include an integral amplifier/power pack, an input unit allowing rapid change of both range and function, provision for up to six blind setting alarm switches and external control of the chart speed and the printing interval.

The instrument has an absolute accuracy within 0·5 per cent and a resolution of 0·2 per cent. Its response speed is one second and the standard multi-point printing speed is one point every five seconds, although this can be reduced to two seconds, on request.

Designed for panel, or desk mounting, the recorder is approximately 21in long by 6in square and is housed in a dust-proof metal case.

All the resistors determining the type and range of signal to be measured are mounted on an interchangeable printed circuit range unit. A maximum input source impedance of 1000Ω is recommended on all ranges.



EE 98 772 *for further details*

MOISTURE METER

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, Yorkshire

(Illustrated below)

Accurate moisture testing of organic or inorganic powders and similar substances requires that test samples should always be at the same density, as the dielectric constant will otherwise vary. This is the parameter measured by capacitance type moisture meters. The new Shaw pressure-type moisture meter is believed to be the first instrument specially designed to ensure this condition.

The instrument consists of two units, viz. an electronics cabinet surmounted by a hand-operated pressure head containing sensing electrodes. The plunger actuating mechanism is designed to give maximum leverage at the highest pressure required and enables female labour to operate the unit over long periods without fatigue. According to the material tested, pressures from 0 to 120 lb can be exerted.

'Level' and range controls are 10-turn potentiometers with micrometer dials marked 0 to 1000 and used to obtain a suitable setting. Moisture content readings on, e.g., instant coffee, malt flour, tobacco and starch can be obtained with 0·1 per cent accuracy or better. With an all solid state circuit and operating at v.h.f., the new Shaw instrument requires no maintenance.

EE 98 773 *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.

Date: 3 October.
Discussion: The Mossbauer Spectrometer.
Opened by: E. L. Bush and H. Grayson.

Date: 3 October.
Discussion: Parameters Which Need to be Standardised to Specify a p.c.m. Interface.
Opened by: R. O. Carter.

Date: 5 October.
Lecture: Measurement of Arterial Blood Flow by the Pressure-gradient Method.
By: I. Gabe and S. Montgomery.
(Joint meeting with IERE Medical and Biological Group at the IERE, 8-9 Bedford Square, London, W.C.1, at 6 p.m.).

Date: 6 October.
President's Inaugural Address.
By: J. A. Ratcliffe.

Date: 10 October.
Lecture: Dielectric Properties of Polythene for Submarine Telephone Cables.
By: J. T. Barrie, K. A. Buckingham and W. Reddish.

Date: 10 October.
Colloquium: Microprogramming and Fixed Stores.
(Joint meeting with IERE Computer Group at 2.30 p.m.).

Date: 11 October.
Automation and Industry—the Technical and Economic Approach to Wider Utilisation. Control and Automation Division chairman's address by S. S. Carlisle.

Date: 13 October.
Discussion: Engineering Industry Training Board—A New Climate in Training.
Opened by: F. Metcalfe.

Date: 14 October.
Lecture: Barn Hay Drying.
By: C. W. Gould.

Date: 18 October.
Report: Dubrovnik Symposium Report—the Control of Powered and Artificial Limbs:
By: A. B. Kinnier Wilson, G. W. Lange and other speakers. (Joint meeting with IMechE automatic control group).

Date: 18 October.
Discussion: Transfer-function-measuring Instruments.
Opened by: F. Jones and J. Kirby.

Date: 19 October.
Some Highlights in the Engineering of the Independent Television Service. Electronics Division chairman's address by P. A. T. Bevan.

Date: 20 October.
Discussion: Standards of Reliability of Electrical Equipment in Sea, Air and Road Transportation.
Opened by: J. Davidson, F. D. Hoyle and another speaker.

Date: 24 October.
Lecture: A Slow-motion Videotape Recorder.
By: P. Rainger.
(Joint meeting with Television Society and IERE).

Date: 25-26 October.
(Joint Technical Conference: Euston Main-Line Electrification. Registration forms available from IMechE, 1 Birdcage Walk, London, S.W.1.

Date: 26 October.
Power Supply for 1970. Power Division chairman's address by E. S. Booth.

Date: 27 October.
Lecture: Mechanisms of Microwave Generation in Gunn Oscillators and Avalanche Diodes.
By: A. Chynoweth.

London Graduate and Student Section London Meetings

All meetings will be held at Savoy Place, commencing at 6.30 p.m. unless otherwise stated.

Date: 17 October.
Chairman's address: The Electrical Breakdown of Gases.
By: S. M. Foggo.

Outer London Meetings

Held at: Hendon College of Technology, The Burroughs, N.W.4.
Date: 26 October.
Time: 6.30 p.m.
Lecture: Colour Television Systems.
By: A. V. Lord.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Joint I.E.R.E.-I.E.E. Medical and Biological Electronics Group

Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 5 October.
Time: 6 p.m.
Lecture: Measurement of Arterial Blood Flow by the Pressure-gradient Method.
By: I. Gabe and S. Montgomery.

Joint I.E.E.-I.E.R.E. Computer Group

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 10 October.
Time: 2.30 p.m.
Colloquium: Microprogramming and Fixed Stores.

Joint I.E.E.-I.E.R.E. Television Group

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 24 October.
Time: 5.30 p.m.
Lecture: A Slow Motion Video Tape Recorder.
By: P. Rainger.

West Midland Section

Held at: Department of Electronic and Electrical Engineering, University of Birmingham.
Date: 27 October.
Time: 7.15 p.m.
Lecture: Radiophonic Effects in the BBC.
By: C. Brooker.

Held at: Wolverhampton College of Technology, Wolfruna Street.
Date: 12 October.
Time: 7.15 p.m.
Lecture: The Principles of F.D.M., T.D.M. and P.C.M., as Applied to Telephone Transmission.
By: A. C. Frost.

South Western Section

Held at: Bristol Technical College, Ashley Down, Bristol 7.
Date: 12 October.
Time: 7 p.m.
Lecture: Commercial Personal Portable Communications Schemes.
By: D. A. S. Dryborough.

Scottish Section

Held at: Department of Natural Philosophy, The University, Drummond Street, Edinburgh.
Date: 5 October.
Time: 7 p.m.
Lecture: Colour Television.
By: G. N. Patchett.

Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Date: 6 October.
Time: 7 p.m.
Lecture: Colour Television.
By: G. N. Patchett.

Southern Section

Held at: Farnborough Technical College, Farnborough, Hants.
Date: 13 October.
Time: 7 p.m.
Lecture: M.O.S. Transistors.
By: G. G. Bloodworth.

Held at: Physics Department, Lecture Theatre No. 1, The University, Southampton.
Date: 25 October.
Time: 6.30 p.m.
Lecture: Plasma Spectroscopy.
By: D. W. Braggins.

East Anglia Section

Held at: Associated Electrical Industries Research Laboratories, West Road, Temple Fields, Harlow, Essex.
Date: 19 October.
Time: 7 p.m.
Lecture: Micro-Electronics.
By: P. L. Hawkes.

Held at: Hornchurch College of Further Education, 42 Ardleigh Green Road, Hornchurch, Essex.
Date: 5 October.
Time: 7 p.m.
Lecture: The Development of Satellite Communications.
By: J. K. S. Jowett.

Yorkshire Section

Held at: Huddersfield Technical Teachers Training College.
Date: 13 October.
Time: 7 p.m.
Lecture: Education and Training of Electronic Engineers.
By: R. A. Elsdon.

Merseyside Section

Held at: Liverpool College of Technology.
Date: 19 October.
Time: 6.30 p.m.
Lecture: Numerical Control of Machine Tools.
By: D. F. Walker.

North Eastern Section

Held at: Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Date: 12 October.
Time: 6 p.m.
Lecture: The Communications Aspects of the Post Office Tower.
By: L. R. N. Mills.

Thames Valley Section

Held at: Lecture Theatre, Reading College of Technology.
Date: 26 October.
Time: 7.30 p.m.
Lecture: Reliability of Electronic Components and Microelectronic Circuits.
By: K. G. Nichols.
(Joint I.E.R.E.-British Productivity Council Meeting).

SOCIETY OF ELECTRONIC AND RADIO TECHNICIANS

London

Held at: Royal Society of Arts, John Adam Street, London, W.C.2.
Date: 12 October.
Time: 7 p.m.
Lecture: Hybrid Television Receivers.
By: H. A. Heins (Mullard Limited).

Newcastle

Held at: Charles Trevelyn College, Maple Terrace, Newcastle.
Date: 5 October.
Time: 7.15 p.m.
Lecture: Colour Television.
By: B. J. Rogers (Rank-Bush Murphy).

Swansea

Held at: College of Technology, Mount Pleasant, Swansea.
Date: 26 October.
Time: 7.15 p.m.
Lecture: Television Systems.
By: E. A. W. Spreadbury (Electrical and Electronic Trader).

Southampton

Held at: College of Technology, East Park Terrace, Southampton.
Date: 12 October.
Time: 7.30 p.m.
A.G.M. followed by: Colour Television.
By: E. J. Finn.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Formal Meetings

Formal Meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embarkment, London, W.C.2, commencing at 5 p.m.

Date: 4 October.
Lecture: Operational Research in Telecommunications Economics.
By: N. V. Knight.

Informal Meetings

Informal Meetings will be held in the Conference Room, Fleet Buildings, 70 Farringdon Street, London, E.C.4, commencing at 5 p.m.

Date: 24 October.
Lecture: Planning a PABX Satellite Scheme.
By: D. Blackwell.

NOUVELLES Réalisations

Traduction des pages 682 à 687

Une description basée sur des renseignements fournis par les fabricants de nouveaux composants, accessoires et instruments d'essai

SERRURE DE SÛRETÉ ÉLECTRONIQUE

Dynastat Electronics Ltd. 495 Green Lanes, London. N.13

(Illustration à la page 682)

Cette serrure de sûreté d'un modèle très perfectionné constitue non seulement un élément de contrôle indispensable pour coffres-forts, caisses-fortes, abris, dépôts à documents, etc., mais également pour un nombre étendu d'applications industrielles, telles que les calculatrices et machines automatiques, les dépôts de bandes magnétiques, de produits dangereux, etc. En somme, pour tout usage commercial ou industriel nécessitant un degré très élevé de protection contre toute irruption non-autorisée.

Elle permet le choix d'une combinaison aléatoire de quatre, huit, douze chiffres ou même d'avantage et comporte un élément à mémoire pour trouver et emmagasiner la combinaison voulue. L'appareil ou le matériel est ainsi protégé de manière absolue et toute tentative d'en forcer le secret sans connaître la combinaison correcte déclenche le circuit d'alarme.

La combinaison d'ouverture, dont la formule peut être partagée par deux ou plusieurs personnes, est choisie au hasard au moment du verrouillage et peut être changée à volonté. Il n'existe donc pas de combinaison permanente ou semi-permanente préréglée et de risques de sécurité.

La combinaison choisie ne peut être retrouvée par un quelconque processus de déduction, pas plus qu'une manipulation quelconque du sélecteur ne trahira à l'un des sens la proximité de la position correcte.

Notre photographie montre le dispositif en cours de démonstration et utilisant des sélecteurs à quatre chiffres. On le voit sous forme modulaire mais il peut être adapté à n'importe quelle utilisation voulue. En pratique, cependant, les sélecteurs à huit chiffres fournissent un degré supérieur de sûreté de sorte qu'un dispositif de ce genre ne donne qu'une chance sur 10^8 à toute personne non-autorisée de trouver la bonne combinaison.

Le choix d'une fausse combinaison ne laisse par seulement le dispositif complètement sûr mais met en branle le système d'alarme.

La sûreté de l'appareil ne peut être compromise par une forme quelconque d'attaque électrique ou mécanique; même à distance du matériel protégé, l'emploi de fils de coupage ou de court-circuitage ou d'une forme quelconque de signal ne peut que résulter en un déclenchement du circuit d'alarme.

Le verrouillage ou le déverrouillage s'effectuent uniquement par courant électrique et la serrure est entièrement indépendante du courant secteur durant les périodes de contrôle. Les variations de courant secteur ou les pannes n'ont donc aucun effet sur l'instrument.

EE 98 751 *pour plus amples renseignements*

LECTEUR DE BANDES DE PAPIER

Electrographic. 38 Corwell Lane. Hillingdon, Middlesex

(Illustration à la page 682)

Un nouveau lecteur photoélectrique de bandes de papier perforées a été réalisé par la société Electrographic. Cet appareil, type 800, peut lire 100 chiffres binaires par seconde et s'arrêter sur n'importe quel caractère voulu.

Il peut revevoir des bandes de n'importe laquelle des dimensions normales, de 5 à 8 trous, plus le trou d'attaque. La bande est mise en mouvement par un galet entraîné par un moteur à gradins qui assure un entraînement positif et une réponse rapide.

Grâce à un nouveau montage optique, n'importe quelle bande de papier peut être lue avec un rapport élevé d'arrêt/marche. Une cellule photoélectrique placée dans la position du trou d'attaque permet d'effectuer l'alignement précis de la bande dans la position de lecture.

Le lecteur a été conçu de manière à ce que l'optique ne puisse pas être encrassée par la poussière ou des impuretés. Le code lu par l'appareil peut être contrôlé visuellement dans la position de lecture.

Une lampe à faible tension avec un gros filament est employé dans un état d'utilisation réduite de manière à prolonger considérablement sa durée de vie.

La bande peut être déplacée dans l'une ou l'autre direction par simple inversion de l'entraînement du moteur à gradins.

Un amplificateur intermédiaire à cellule photoélectrique et une plaquette d'entraînement du moteur sont livrables montées sur châssis de façon à pouvoir les adapter à un bâti standard de circuit imprimé "Imhof".

EE 98 752 *pour plus amples renseignements*

LIGNE À RETARD DE TÉLÉVISION

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

(Illustration à la page 683)

Les principaux avantages de la nouvelle ligne à retard préréglée à ultrasons réalisée par Mullard pour les télérécepteurs PAL en couleurs sont la simplification des circuits décodeurs du récepteur et l'élimination de tout réglage sur place de la ligne à retard.

En plus des avantages particuliers de la nouvelle ligne Mullard, l'emploi d'une ligne à retard dans les récepteurs PAL peut permettre de simplifier les problèmes de construction des autres parties du récepteur et d'améliorer la qualité de l'image.

La production en série de la nouvelle ligne, type DL1, commencera cette année, des quantités "échantillons" étant déjà fournies pour les récepteurs de mise au point.

La nouvelle ligne, constituée par une pièce de verre pliée et réfléchissante est plus facile à utiliser que d'autres lignes conçues pour le même usage car elle fournit le retard désiré de 63,943 μsec à 4,4336 MHz sans nécessité de réglage sur place à l'aide de circuits supplémentaires et étrangers.

La précision de la ligne à retard est préréglée en cours de fabrication, d'abord par le meulage précis de la surface réfléchissante du bloc de verre, puis par le réglage des inducteurs incorporés variables.

L'entrée à la ligne est appliquée à un transducteur en céramique qui convertit le signal électrique en une onde acoustique transmise à travers le bloc. Lorsque cette onde est renvoyée à un point proche de l'entrée, elle est reconvertie en un signal électrique par un autre transducteur analogue. Etant donné que la ligne est du type plié et réfléchissant, les transducteurs peuvent rester en place pendant que le temps de retard est préréglé.

La perte par insertion nominale n'étant que très faible, c'est à dire de 10 dB, la DL1 peut fournit une tension suffisante aux bornes de sortie pour alimenter directement un démodulateur à diodes. Elle a été étudiée pour l'utilisation entre des sources sensiblement non-réactives de 150Ω.

Un avantage supplémentaire pour les fabricants de télérécepteurs est que la ligne DL1 est fournie complète avec des informations de circuit donnant des signaux d'entraînement de différence de couleur (*R—Y*), (*B—Y*) et (*G—Y*) d'une amplitude constante à partir d'un simple réseau, ce qui obvie aux problèmes dûs à la répartition de la perte et du gain entre les circuits.

La ligne DL1 est munie de broches de connexion qui permettent de l'insérer directement dans une plaquette de circuit imprimé.

EE 98 753 *pour plus amples renseignements*

HORLOGE NUMÉRIQUE

Alma Components Ltd, Park Road, Diss, Norfolk

(Illustration à la page 683)

Cette nouvelle horloge numérique a été principalement conçue pour servir de minuterie principale de programme. Un moteur synchrone alimenté par le courant secteur entraîne en cascade une série d'unisélecteurs à lame vibrante Alma pour fournir une indication numérique d'heures et de minutes. Le réglage initial à l'heure exacte s'effectue simplement par 4 commutateurs à boutons-poussoirs à lames vibrantes se trouvant à l'arrière de l'instrument. La conception de l'appareil est telle qu'à toutes les minutes du cycle de 24 heures, les unisélecteurs peuvent être reliés par l'utilisateur pour fermer ou ouvrir un contact interne d'une combinaison de 1440 contacts de commutation dont la puissance nominale de chacun est de 250 V, 1 A, la puissance nominale maxima de la tension étant de 60 W. Le cycle complet est répété toutes les 24 heures; en variante, on peut convertir l'horloge pour un fonctionnement de 12 heures en actionnant un commutateur. L'horloge peut être prévue pour n'importe quel cycle de temps, c'est à dire 48 heures, 60 heures, etc. ou encore pour le minutage en minutes et secondes. Grâce à la souplesse de la réalisation, des modifications spéciales peuvent être apportées a l'horloge pour la rendre propre à des applications particulières, telles que, par exemple, retarder l'action pour permettre un temps de lecture prolongé.

L'horloge normale de 24 heures peut être fournie soit comme appareil autonome dans un coffret en aluminium soit sous forme de châssis pouvant être installé dans un autre équipement.

EE 98 754 *pour plus amples renseignements*

ASSEMBLAGES DE MATRICES À NOYAUX

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

(Illustration à la page 683)

La société Mullard a mis au point une gamme de matrices à noyaux peu coûteuses et à plan unique. Ces matrices on été conçues pour les petites calculatrices électroniques de bureau. Utilisées dans les réservoirs et registres de ces calculatrices, les matrices sont plus économiques et plus petites que les systèmes à semi-conducteur et elles conservent toutes les données emmagasinées même lorsque la calculatrice est débranchée du courant secteur. La stabilité opérationnelle dans une gamme étendue de température (0 à 70°C) est assurée par l'emploi de noyaux de ferrite au lithium-nickel. Le fonctionnement à un faible courant d'entraînement (190 mA) s'effectue par l'emploi d'enroulements à deux tours X, Y et Z, afin de réduire les exigences d'entraînement et de permettre ainsi l'emploi de circuits d'entraînement et de sélection relativement simples. Le bobinage est conforme au système bien connu MIT.

Les matrices résistent parfaitement aux manipulations brutales car l'assemblage de noyaux est supporté par un cadre rigide, les noyaux étant tenus par une laque spéciale à une plaque à dos de papier laminé. Prévues pour le soudage direct sur des plaquettes de circuit imprimé, les connexions de broches des matrices s'adaptent à une plaquette de circuit de 25 mm.

Des cadres en époxyde pouvant résister à des conditions d'humidité plus élevées sont fournis moyennant un léger supplément de prix. Des assemblages de matrices allant jusqu'à quatre niveaux avec des fils d'entraînement reliés en série peuvent également être fournis. La gamme suivante constitue la série normale des assemblages:

Numéro de type	*Nombre de noyaux*	*dimensions extérieures*
AW3526	16 × 16	82 × 82 mm
AW3527	4 × 8 × 8	82 × 82 mm
AW3529	4 × 12 × 12	102 × 102 mm
AW3531	4 × 16 × 16	122 × 122 mm
AW3532	32 × 32	122 × 122 mm

EE 98 755 *pour plus amples renseignements*

PLAQUETTES À PROGRAMMES

Sealectro Ltd, Farlington, Portsmouth, Hampshire

(Illustration à la page 683)

Pour répondre à la demande de plaquettes répondant à des conditions spéciales de sécurité, la société Sealectro Ltd a réalisé un "Sealectoboard" équipé d'un panneau de recouvrement transparent qui empêche le retrait non-autorisé des broches. En même temps, le recouvrement protecteur assure une vision non-obstruée du programme et de la disposition des broches.

Le panneau à recouvrement transparent est également prévu pour les applications où la plaquette de programme est sujette à des vibrations qui pourraient autrement laisser les broches se détacher de la plaquette.

La plaquette à programme sans corde est prévue en montages spécial et standard offrant des combinaisons pratiquement illimitées de dispositions de matrices électriques à trous verticaux et horizontaux répondant à tous les besoins.

EE 98 756 *pour plus amples renseignements*

CADRANS DE PRÉCISION MULTITOURS

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

(Illustration à la page 684)

La société Kynmore Engineering Co. Ltd vient d'ajouter à la gamme des instruments Kynmore trois nouveaux cadrans de précision multitours étalonnés d'un bel aspect, fabriqués par le groupe atomique français.

Le nouveau cadran micro-montre (sur le côté gauche de notre gravure) est un cadran analogique à dix tours comportant 50 divisions marquées par tour. Son diamètre est de 22 mm et son épaisseur est de 29 mm. Le nouveau cadran analogique à 15 tours, modèle US45 (en haut) a cent divisions marquées par tour. Il mesure 45,5 mm de diamètre et 30,5 mm de profondeur. Le nouveau cadran numérique miniature, modèle Flash 22 (à droite) est un cadran à cent tours à lecture agrandie et à dix divisions numérotées par tour. Il mesure 23,4 mm de diamètre et 35,2 mm de profondeur.

Les trois cadrans sont conçus pour des axes de 6 mm de diamètre et ils sont munis de leviers à freins qui permettent de les fixer dans n'importe quelle position voulue.

EE 98 757 *pour plus amples renseignements*

OSCILLOGRAPHE À LECTURE DIRECTE

Honeywell Controls Ltd, Brentford, Middlesex

(Illustration à la page 684)

Le nouvel oscillographe 2106 "Visicorder" de la société Honeywell Controls Ltd a été conçu et mis au point aux usines de Lanarkshire afin de produire une spécification extrêmement élevée dans une structure à bas prix, offrant de nombreuses caractéristiques qui ne sont pas normalement le propre des instruments dans cette gamme de prix. Cet élément compact comporte en effet un diagramme de 15,24 cm de large pouvant effectuer des enregistrements à partir de douze canaux (plus deux canaux auxiliaires supplémentaires).

Un des principaux avantages de cet élément est que la lampe UV peut être allumée instantanément, quelle soit chaude ou froide. L'appareil peut enregistrer une multiplicité de variables sur le même diagramme et offre en outre des lignes de minutage et de grille ainsi que la télécommande.

L'appareil se caractérise en outre par le choix de vitesses de diagramme avec des boîtes de vitesses qui fournissent une gamme totale de 1 cm/min à 300 cm/sec et une plaquette de résistance individuelle à fiches pour chaque galvanomètre permettant de constituer une série de plaquettes d'entrée avec différents réseaux d'équilibrage et d'atténuation, ce qui élimine le temps de réglage toujours fastidieux et coûteux. La vitesse d'écriture est supérieure à 125 cm/ms et une commande facultative de longueur d'enregistrement empêche le gaspillage de papier. L'oscillographe numéro 2106 peut recevoir jusqu'à 45,72 mètres de papier dont la largeur peut se réduire jusqu'à un minimum de 7,62 cm.

La stabilité du rendement est maintenue par le chauffage et le contrôle thermostatique de la température du bloc galvanométrique. N'importe quel galvanomètre peut être introduit ou retiré en quelques secondes.

On prévoit un marché mondial pour cet élément de construction anglaise ainsi que pour la nouvelle gamme Honeywell de galvanomètres de la série M pouvant être utilisée avec cet oscillographe.

Le miroir de ces galvanomètres à haute fréquence n'est pas immergé dans le liquide d'amortissement. La déviation de l'image lumineuse à travers le diagramme n'est donc pas limitée par la dispersion de la lumière UV dans le fluide. Les galvanomètres allient la haute sensibilité et la stabilité à des réponses de fréquence linéaire allant du courant continu à 13 kHz.

EE 98 758 *pour plus amples renseignements*

STABILITÉ DE FRÉQUENCE

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustration à la page 684)

Une stabilité de fréquence nettement améliorée au-dessus de 50 kHz est assurée par le nouveau synchronisateur de générateur de signaux Hewlett-Packard modèle 8708A.

Spécialement conçu pour fonctionner en liaison avec les nouvelles versions de générateur de signaux HP modèle 606 et 608, le nouveau synchronisateur utilise les méthodes de verrouillage de phase pour fournir des signaux d'essai d'une stabilité de deux parties dans 10^7 pendant une période de 10 minutes. Ceci constitue une amélioration de plus de 250 fois par rapport à la stabilité précédemment fournie avec les générateurs de signaux standard fonctionnant dans cette gamme de fréquence.

La synchronisation automatique s'effectue à n'importe quelle fréquence dans la gamme des générateurs de signaux perfectionnés 606B (50 kHz à 65 MHz) ou 608F (10 MHz à 455 MHz). Les générateurs perfectionnés fournissent non seulement une stabilité accrue mais ils ont également une nouvelle capacité pour la modulation de fréquence et de phase ainsi que pour la modulation d'amplitude à faible distorsion. Le pourcentage de sortie de puissance ainsi que celui de modulation demeure constant en fonction des changement de fréquence, la fréquence de sortie stabilisée ayant une grande pureté spectrale et la modulation de phase/fréquence étant hautement linéaire.

EE 98 759 *pour plus amples renseignements*

BLOC D'ALIMENTATION À HAUTE TENSION

Brandenburg Ltd, 139 Sanderstead Road, South Croydon, Surrey

(Illustration à la page 684)

Les modèles 905 et 906 constituent les plus récents blocs d'alimentation stabilisée à haute tension dans la nouvelle gamme BETA. Ils fournissent tous deux 400 μA jusqu'à 60 kV bien que ne mesurant que 17,7 cm de haut; 43,18 cm de profondeur et 48,2 cm de largeur.

La tension de sortie peut être réglée au moyen de potentiomètres à réglage précis ou approximatif jusqu'à n'importe quelle valeur entre 6 et 60 kV et elle est contrôlée par un instrument de mesure du panneau frontal. Le panneau arrière permet en outre la mesure du courant. Le modèle 905 a une polarité positive et le modèle 906 une polarité négative.

Les dimensions extrêmement compactes de ces blocs d'alimentation sont dues en grandes partie au dessin et à la construction perfectionnée de la tête à haute tension. L'amplificateur comparateur constitué de corps solides a été conçu sous forme de plaquettes accouplées à des connecteurs et les autres grands composants sont portés par un robuste câble en alliage. La construction et l'aspect sont presque identiques à ceux des modèles 705 et 800; les modèles 905 et 906 conviennent également au montage sur bâti de 48,2 cm ou sur banc d'essai. Le rendement est très semblable à celui des modèles BETA courants couvrant toutes les exigences industrielles et ce laboratoire.

EE 98 760 *pour plus amples renseignements*

VOLTMÈTRE ÉLECTRONIQUE

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustration à la page 684)

Le VM76 est le plus récent dans la gamme des 5 voltmètres Advance et permet à la gamme d'inclure des fréquences de courant alternatif dépassant 1 GHz et des mesures de tension continue allant jusqu'à 1000 V avec des précisions de $\pm 2\%$ jusqu'à à 100 V.

La sélection par bouton-poussoir est employée pour les diverses fonctions. Dans le fonctionnement à tension continue cette méthode peut être utilisée pour choisir des lectures à zéro central et une polarité négative ou positive. La mesure de résistance entre 0,02 Ω et 500 MΩ à une précision de $\pm 5\%$ au milieu de l'échelle peut être obtenue de façon analogue. Les impédances d'entrée typiques sont de 7,5 MΩ à 1 kHz et 150 kΩ à 100 MHz, toutes deux avec une capacité de dérivation de 1,5 pF.

Une excellente accessibilité aux composants est assurée grâce au montage de la plupart des composants sur une plaquette unique à circuit imprimé. Tous les instruments sont munis de sondes ainsi que de câbles connecteurs avec des connexions à fiches et prises uniques. Des accessoires facultatifs sont prévus pour toutes les applications du voltmètre.

Le VM76 a une gamme de température de 0°C à +40°C et fonctionne sur courant secteur. De même que les autres instruments Advance récemment introduits il se trouve dans un robuste coffret modulaire à largeur de bâti du 2/3 de la norme.

EE 98 761 *pour plus amples renseignements*

GÉNÉRATEURS DE SIGNAUX POUR SONDEUR ACOUSTIQUE

Elliott Instruments Co. Ltd, Dunmow, Essex

(Illustration à la page 685)

La Elliott Instruments Co Ltd vient d'annoncer la création de nouvelles versions de ses générateurs de signaux pour sondeur accoustique. Ces instruments ont été spécialement conçus à l'intention des ingénieurs effectuant des travaux d'entretien à bord des navires; ils constituent également des sources commodes de signaux pour les laboratoires.

Entièrement transistorisés, ces instruments qui portent la référence UL3, produisent des sorties d'ondes sinusoïdales couvrant la gamme de fréquence de 7 à 60 kHz ainsi qu'un signal variant de 1 microvolt à 1 volt. Fonctionnant sur batterie, ils sont robustes et compacts mesurant 16,51 cm × 5,08 cm et pesant environ 2 kg. Ils permettent la mesure du gain sur les amplificateurs de sondeurs acoustiques, l'instrument de mesure de l'appareil étant relié directement à la sortie de l'amplificateur. Il peut indiquer soit 10 volts c.c. à 10 kΩ d'impédance pour les enregistrers à papier humide ou 500 volts c.a. à 200 kΩ d'impédance pour les enregistreurs à papier sec. L'instrument peut être fourni également sous le modèle UL3a sans possibilité de contrôle.

EE 98 762 *pour plus amples renseignements*

ENSEMBLE D'OSCILLATEUR À CRISTAL

The Marconi Co. Ltd, Chelmsford, Essex

(Illustration à la page 685)

La société Marconi a mis au point une nouvelle gamme d'oscillateurs transistorisés offrant aux ingénieurs une source de fréquence de stabilité moyenne et à bas prix, logée dans un seul coffret à fiches. Cette nouvelle gamme est maintenant en pleine production.

Une alimentation à faible courant de 6 volts est tout ce qui est nécessaire pour fournir un oscillateur à cristal efficace et prêt à l'emploi. Monté sur une base internationale Octal ou B7G, cet oscillateur peut être fourni pour des fréquences de 1 kHz à 100 MHz. Le modèle plus compact B7G, type F3171, couvre la gamme de 115 kHz à 100

MHz et offre en outre l'avantage de pouvoir être placé dans un four à cristal ordinaire Marconi F3006 pour donner une stabilité de température accrue.

L'élément à base plus grande Octal type F3170, couvrant la gamme de 1 à 115 kHz est entièrement enfermé dans un coffret tubulaire d'aluminium d'un diamètre de 1,125 pouce de diamètre. Suivant la fréquence, la hauteur varie de 3,125 à 5 pouces tandis que les oscillateurs à plus haute fréquence, à base B7G ont une hauteur standard de 3,125 pouces. Ces éléments sont également enfermés dans un coffret en aluminium cylindrique d'un diamètre de 0,65 pouce. Des stabilités supérieures à une partie dans 10^6 peuvent être obtenues dans la gamme de fonctionnement de -20 à $+70$°C. Si nécessaire la fréquence des éléments peut être réglée au moyen d'un condensateur variable externe relié entre l'une des broches de base et la terre.

Tous les oscillateurs entourent un cristal de quartz encapsulé dans le verre et comportent un étage de sortie intermédiaire qui maintient une stabilité de fréquence inférieure à une partie dans 10^{-6} pour une variation de 10% dans l'impédance de charge. L'impédance de sortie est de 5000 Ω ±10% avec une tension de sortie de 2 volts de crête à crête.

EE 98 763 *pour plus amples renseignements*

TÊTE D'ENREGISTREMENT

Gresham Lion Electronics Ltd, Magnetic Recording Head Division, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

La société Gresham Lion Electronics Ltd a réalisé une tête d'enregistrement magnétique à neuf pistes de lecture après écriture. Cette tête est prévue pour les applications numériques; elle effectue en effet une vérification immédiate de l'information enregistrée et elle doit être utilisée en liaison avec le nouveau système IBM à 1600 chiffres binaires.

La tête est fournie avec des éléments enregistreurs à gauche pouvant s'accorder avec le matériel IBM ou avec des éléments enregistreurs à droite. Le côté frontal est profilé pour minimiser le bruit prolongé des longueurs d'onde et il est replié en arrière pour empêcher les cannelures causées par les bords des bandes magnétiques.

La tête mesure 3,17 cm de hauteur, 2,61 cm de profondeur et 2,54 cm de largeur; l'azimuth est maintenu entre ±0,0002 radians et l'élévation est maintenue à 0,0005 radians. La dispersion d'entrefer est maintenue dans une bande de 0,00008 pouce de large et les lignes d'entrefer sont parallèles à 0,0002 radians près. Les pistes de lecture ont 0,040 pouce de large et les pistes d'écriture ont 0,044 pouce de large. La séparation entre les lignes est de 0,150 pouce. La société Gresham offre un choix de longueur de lecture: 0,0002 ou 0,0001 pouce; cette dernière longueur est prévue pour le système à 1600 chiffres binaires.

Le courant de saturation sur bande 3M T488 à 300 chiffres binaires par pouce et la saturation de 99% est de 66 mA ±15% de crête à crête.

Les caractéristiques électriques sont comme suit:

(a) Lecture d'entrefer 0,0002 pouce. Lecture: Résistance 3,7 Ω; inductance 600 microhenries. Ecriture: Résistance 2,3 Ω; inductance 415 microhenries.

(b) Lecture d'entrefer 0,0001 pouce. Lecture: Résistance 7,4 Ω; inductance 2,4 millihenries. Ecriture: Résistance 4,6 Ω; inductance 1,6 millihenries.

Le bobinage de la tête d'écriture est à fiche centrale.

EE 98 764 *pour plus amples renseignements*

INVERSEUR STATIQUE

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(*Illustration à la page 685*)

La société Industrial Instruments Ltd vient d'ajouter le modèle 606/8L à sa gamme d'inverseurs statiques Transipack. Cet instrument a été mis au point afin de pouvoir inclure le téléimprimeur dans le domaine du matériel mobile de communication. Il fournit le courant pour entraîner le moteur, actionne les relais de temps/espace et excite l'émetteur de l'une des sources de 24 Vc.c. Cet instrument pleinement stabilisé fournit une tension de +80 V et −80 V (d'autres tensions de signal peuvent être fournies) à une charge de 1 A au maximum et il est muni d'une protection complète contre les surcharges. Il est logé dans un coffret émaillé indépendant offrant un accès facile aux batteries et il comprend un chargeur incorporé.

Grâce à cet instrument, le téléimprimeur peut être rendu réellement mobile et, par conséquent, utilisé dans un beaucoup plus grand nombre d'applications, telles que la construction, les champs pétrolifères, les utilisations militaires, le contrôle des véhicules par la police, etc.

La société Industrial Instruments Ltd fabrique une gamme étendue d'inverseurs statiques qui produisent des courants secteur simulés de 230 Vc.a. à une fréquence de 50 Hz à partir de sources de tension continue. Cette gamme comprend notament le TRANSIPACK, type 606/ST/300 que l'on voit dans notre gravure. Cet appareil peut recevoir 24 volts d'un véhicule ou d'une batterie et produit une sortie continue nominale de 300 VA à 230 c.a. La stabilité de la fréquence est supérieure à ±1% et la distorsion harmonique varie entre 3 et 6%.

EE 98 765 *pour plus amples renseignements*

CONTRÔLEUR DE FILMS MINCES

Genevac Ltd, Pioneer Mill, Radcliffe, Manchester

(*Illustration à la page 686*)

Le nouveau contrôleur de films à cristal de quartz réalisé par la société Genevac Ltd, Radcliffe a été conçu pour mesurer l'éspaisseur totale ainsi que la vitesse de dépôt de films minces déposés sous vide. Cette opération s'effectue en receuillant une partie du courant évaporant sur un cristal de quartz fixé dans une position appropriée à l'intérieur de l'évaporateur. Utilisant le principe du changement linéaire de la fréquence résonante en fonction de la masse accroissante du cristal du contrôleur, le dispositif indique le changement sous forme de tension continue affichée ainsi que la vitesse de changement de cette tension. La réaction de l'instrument peut être utilisée pour contrôler l'entrée de chaleur dans la source évaporante ainsi que la vitesse de dépôt; en préréglant la pointe d'extrémité de l'opération on peut utiliser l'appareil comme obturateur électromécanique.

L'instrument peut contrôler des dépôts de laboratoire et de production avec un degré élevé de répétition et indiquer la masse déposée dans quatre gammes de 1, 5, 10 et 50 kHz de déviation sur la totalité de l'échelle. L'appareil permet également l'équilibrage à zéro entre dépôts consécutifs, jusqu'à un total de 60 kHz de déphasage. A ce moment-là le cristal doit être nettoyé. L'équipement est logé dans deux coffrets mesurant 22,86 cm de largeur, 15,24 cm de hauteur et 15,24 cm de profondeur; l'un des coffrets contient toutes les alimentations nécessaires et affiche la masse déposée tandis que l'autre affiche la vitesse de dépôt et peut être acquis séparément. Un oscillateur placé à côté de l'évaporateur est relié au cristal de quartz du contrôleur par un scellement sous vide coaxial.

EE 98 766 *pour plus amples renseignements*

AMPLIFICATEUR À FAIBLE BRUIT

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(*Illustration à la page 686*)

La société Brookdeal Electronics a réalisé un nouvel amplificateur à transistors au silicium, modèle LA350. Il s'agit d'un amplificateur de laboratoire à haute performance avec une spécification exceptionnelle qui serait en l'avance sur tout ce qui a été produit à ce jour. Ordinairement, un bon indice de bruit n'est réalisé qu'à partir d'une impédance de source distincte dans une largeur de bande étroite. Le LA350 cependant a un indice de bruit inférieur à 2 dB avec des impédances de source allant de 800 Ω à 250 kΩ à des fréquences dépassant 500 Hz de sorte que le facteur limitatif dans la plupart des mesures est constitué par le bruit thermique dans la source. Les nouveaux circuits utilisés portent également ces capacités au delà des restrictions normales des amplificateurs conventionnels à faible bruit et lui donnent les propriétés d'un amplificateur universel de haute qualité: une bonne linéarité, une gamme étendue de fréquence (avec une réponse d'onde carrée acceptable) et un rétablissement rapide de surcharge.

L'instrument comporte des atténuateurs de réaction répartis qui fonctionnent à travers des étages successifs afin d'empêcher la surcharge d'amplification et d'atténuer le bruit par le même facteur que le signal. Ces atténuateurs sont mis en oeuvre par une commande unique par plots de 5dB. Le gain maximum est de 100 dB. La gamme totale de fréquence va de 3 Hz à 300 kHz; elle peut être réduite par des filtres passe-haut et passe-bas qui modifient les plots de 3 dB par des facteurs successifs de 3. D'autres particularités enfin comprennent: un instrument de mesure de sortie qui permet d'utiliser l'instrument comme microvoltmètre de courant alternatif, et le choix entre le montage sur bâti ou bande d'essai.

EE 98 767 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS DE 10 MHz

Distributeurs: Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustration à la page 686)

Le Datapulse, modèle 101, vendu dans le Royaume-Uni par la Dynamco Instruments Ltd, est un générateur d'impulsions de la grandeur d'un livre dont les performances sont absolument remarquables. Ne mesurant que 8,89 cm de hauteur, 21,59 cm de largeur, 27,9 cm de profondeur et pesant environ 3,5 kg, il offre, en effet, des taux de répétition allant jusqu'à 10 MHz, un temps de montée de 5 nanosecondes et des sorties de ± 10 V.

Les caractéristiques du 101 sont celles que l'on ne trouve normalement que dans les modèles coûteux de générateurs d'impulsions. Elles comprennent le fonctionnement à impulsion double ou simple, la durée variable et le retard jusqu'à 10 millisecondes, le contrôle en souplesse du déclenchement et de l'interruption extérieurs et un contrôle de minutage de 1000:1.

Le déclenchement d'impulsions sélectrices comprend le déclenchement synchrone pour l'affichage oscilloscopique stable d'éclats d'impulsions, le déclenchement asynchrone pour les sorties avancées sous forme de minuterie à système ininterrompu et le déclenchement par coïncidence.

Le déclenchement de type oscilloscopique avec commandes de niveau et de polarité de bande fournit une sensibilité de 250 mV. L'appareil peut être efficacement déclenché à partir de n'importe quel point sur la pente positive ou négative de la forme d'onde de tension de ligne de courant alternatif. La souplesse du déclenchement permet également le compte à rebours synchrone du déclencheur exterieur (compte à rebours du taux) pour les applications exigeant la division de fréquence.

EE 98 768 *pour plus amples renseignements*

CONDENSATEURS ÉLECTROLYTIQUES

The Plessey Co. Ltd, Ilford, Essex

(Illustration à la page 686)

Deux nouvelles gammes de condensateurs mises au point par la division TCC du groupe des composants Plessey ont été réalisées pour répondre à la demande des ingénieurs de matériel électronique spécial.

Les condensateurs électrolytiques miniature "Elkomold" sont prévus pour l'emploi avec les circuits transistorisés et le montage sur plaquette de circuit imprimé. Légers et de dimensions réduites, les coffrets en matière plastique rendent inutile l'emploi d'un enrobement isolant. Quatre dimensions de coffrets sont prévues pour une gamme étendue d'utilisations.

Les coffrets ont été étudiés pour le montage vertical et ils ont été codés pour la polarité. La construction et la performance sont semblables à celles des condensateurs électrolytiques miniature standard TCC en coffret métallique. Le coffret en matière plastique est moulé en polypropylène durable et une des extrémités est fermée par un disque phénolique à face de caoutchouc. Des "Elkomolds" réversibles peuvent être fournis sur demande spéciale.

Il existe des capacitances entre 1 et 1000 μF, avec des tensions continues de régime allant jusqu'à 350 V pour les valeurs inférieures. Les condensateurs peuvent être utilisés de manière continue sans qu'ils perdent leur valeur dans la gamme de température de -25°C à $+70$°C.

Les condensateurs "Superlytic" TCC ont été spécialement conçus pour une longue durée de vie Un contrôle de processus rigoureux et des matières premières spécialement choisies leur assurent une durée de vie allant jusqu'à 10 ans.

La gamme de température de fonctionnement va de -25°C à $+85$°C et le courant de fuite ne dépasse pas 0,01 CV μA ou 5 μA, suivant celles de ces deux valeurs qui est la plus élevée. Les condensateurs "Superlytic" sont fournis en boîtier d'aluminium avec manchon isolant en chlorure de polyvinyle.

Les capacités prévues sont de 1 à 3000 μF avec des tensions de régime de 3 V à 250 V c.c. Les dimensions des boîtiers varient entre 1,90 cm $\times$ 0,95 cm de diamètre et 11,43 cm $\times$ 6,35 cm. de diamètre.

EE 98 768 *pour plus amples renseignements*

INTÉGRATEUR-MINUTERIE

Distributeurs: High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Illustration à la page 687)

La société High Volt Linear Ltd vient de mettre au point le nouvel intégrateur ORTEC modèle 430 et le nouvel intégrateur-minuterie modèle 431.

L'intégrateur modèle 430 est un appareil de comptage à 6 décades et à circuit intégré comprenant un discriminateur autonome et une sortie d'impression. Il a été conçu suivant la nouvelle norme AEC pour instruments nucléaires TID-20893 et occupe deux positions dans le nouveau système modulaire à douze positions, modèle 401/402. Les autres particularités de l'appareil comprenent un taux de comptage de 16 MHz, un discriminateur interne, des entrées positives ou négatives et des facilités de porte. En ajoutant au système l'une des commandes d'impression modèle 432 on peut obtenir un système d'obtention de données de sortie très souple allant jusqu'à 50 intégrateurs.

L'intégrateur-minuterie modèle 431 est un dispositif de comptage et d'impression préréglable à six décades et à circuit imprimé pouvant être utilisé aussi comme minuterie de précision. Le modèle 431 se conforme à la même norme que le modèle 430. D'autres particularités comprennent le taux de comptage de 4 MHz une possibilité d'accord donnant une précision de 0,04% et des facilités de porte. Comme le modèle 430, jusqu'à 50 intégrateurs-minuteries peuvent être utilisés en liaison avec la commande d'impression modèle 432 pour former un système fort souple d'acquisition de données.

EE 98 769 *pour plus amples renseignements*

CONTACTEUR DE RELAIS

Londex Ltd, 207 Anerley Road, London, S.E.20

(Illustration à la page 687)

Une nouvelle gamme de contacteurs de relais d'une importance particulière pour les machines-outils et les applications de contrôle similaire est en cours de production à la société Londex Ltd, membre du groupe Elliott Automation.

Ces nouveaux contacteurs type "Contour" sont extrêmement robustes n'exigent qu'un espace de panneau très restreint et sont conformes aux normes BSS 775-1956, CSA et CEE.

Ils se caractérisent en particulier par la construction triangulaire des porteurs de contacts fixes qui permet de disposer les mêmes composants soit pour le fonctionnement de rupture soit pour celui d'ouverture par simple inversion. L'emploi de contacts du type "à pont" permet de produire un effet de permutation unipolaire en reliant l'un des côtés du contact de rupture avec l'autre côté du contact d'ouverture. Les bobines peuvent être changées avec une simplicité remarquable sans affecter le réglage des contacts. Une nouvelle bobine de même tension ou d'une tension différente peut être fixée en retirant les trois vis qui tiennent la bobine en place ainsi que l'enveloppe de cette dernière. Le mécanisme de fonctionnement est à alignement automatique et assure un bon contact ainsi qu'une pression appropriée.

On peut fixer jusqu'à douze contacts unipolaires à double fermeture par multiple de trois. Ces contacts peuvent ouvrir ou fermer une combinaison des deux types. Les valeurs nominales non inductives typiques sont de 15 A à 250 Vc.a., 15 A à 50 Vc.c.

Les bobines normales pour 50 Hz c.a. sont de: 6 V, 12 V, 25 V, 115 V, 380 V, 415 V et 440 V. Il existe également des bobines pour d'autres tensions alternatives ainsi que pour la tension continue.

EE 98 770 *pour plus amples renseignements*

ENREGISTREUR POTENTIOMÉTRIQUE

A.E.I. Electronics Ltd, New Parks, Leicester

(Illustration à la page 687)

La société AEI Electronics Harlow vient de mettre au point une version perfectionnée—le modèle Mark 5—de son enregistreur à diagramme potentiométrique de 10 cm prévu pour l'utilisation avec les couples thermoélectriques, les thermomètres à résistance, les extensomètres et divers générateurs de signaux.

L'enregistreur Mark 5 est entièrement muni de composants constitués de corps solides afin de lui assurer une grande fiabilité, le maximum de compacité et une consommation réduite de courant Le Mark 5 peut être fourni sous forme d'enregistreur à trace continue unique ou à plusieurs points et capable de recevoir des signaux de 2, 3 ou 6 canaux et d'effectuer une impression par points en couleur. Ses principales caractéristiques comprennent un bloc d'alimentation/amplificateur faisant partie intégrante de l'appareil, un élément d'entrée permettant un changement rapide de gamme et de fonction, la possibilité de lui ajouter jusqu'à 6 commutateurs d'alarme de réglage aveugle et la commande externe de la vitesse du diagramme et l'intervalle d'impression.

L'instrument a une précision absolue de 0,5% et une résolution de 0,2%. Sa vitesse de réponse est de 1 seconde et la vitesse d'impression normale à plusieurs points est de un point par 5 secondes, bien que cette vitesse puisse être réduite à 2 secondes si nécessaire.

Conçu pour montage sur panneau ou sur pupitre, l'enregistreur mesure environ 53,34 cm de longeur × 15,24 cm de côté et il est logé dans un coffret métallique à l'épreuve des poussières.

Toutes les résistances qui déterminent le type et la gamme de signaux devant être mesurés sont montés sur un élément de gamme à circuit imprimé interchangeable. Une impédance de source à entrée maxima de 1000 Ω est recommandée pour toutes les gammes.

EE 98 772 *pour plus amples renseignements*

INSTRUMENT DE MESURE DE L'HUMIDITÉ

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, Yorkshire

(Illustration à la page 687)

Le contrôle précis de l'humidité de poudres organiques ou inorganiques ainsi que de substances semblables exige des échantillons d'essai qu'ils soient toujours à la même densité car la constante diélectrique pourrait varier autrement. C'est là le paramètre mesuré par les instruments de mesure de l'humidité du type à capacité. Le nouvel instrument de mesure de l'humidité du type à pression Shaw serait le premier instrument du genre spécialement conçu pour répondre à cette condition.

L'instrument se compose de deux éléments, à savoir un coffret éléctronique surmonté d'une tête à pression actionnée à la main et contenant les électrodes sensibles. Le mécanisme à plongeur a été conçu pour assurer le maximum d'abattage à la plus haute pression requise et permet aux ouvrières d'actionner l'instrument pendant de longues périodes et sans fatigue aucune. Suivant le matériau soumis à l'essai, des pressions de 0 à 55 kilos peuvent être exercées.

Les commandes de "niveau" et de gamme sont des potentiomètres à dix tours avec de cadrans micrométriques marqués de 0 à 1000 et utilisés pour obtenir un réglage approprié. Des indications sur la teneur en humidité de produits tels que le café instantané, le malt, la farine le tabac et l'amidon peuvent être obtenues avec une précision de 0,1% ou même supérieure. Avec un circuit entièrement constitué de corps solides et fonctionnant à très haute fréquence, le nouvel instrument Shaw n'exige aucun entretien.

EE 98 773 *pour plus amples renseignements*

Résumés des Principaux Articles

Une minuterie multivoies à usages multiples par A. I. Nizan

Résumé de l'article aux pages 634 à 638

La minuterie électronique dont il est question dans cet article comporte six voies de minutage indépendantes. La gamme de chaque voie va de 0,1 à + 99,8 sec. Le minutage à l'intérieur de cette gamme peut être réglé à intervalles de 0,1 sec. Le degré de précision est pratiquement égal à celui de la fréquence secteur mais il peut être augmenté si l'on utilise un oscillateur à minutage externe de précision. Cet oscillateur peut, en même temps, affecter l'échelle de temps et, par consequent, étendre la gamme de l'instrument. En outre, n'importe laquelle des six voies peut être utilisée comme porte à triple coïncidence. La minuterie fonctionne suivant le mode simple ou cyclique. Cet instrument s'est avéré très utile pour le calcul analogique ou hybride.

Convertisseurs d'impédance utilisant des amplificateurs à réaction à large bande par F. Butler

Résumé de l'article aux pages 639 à 641

Il s'agit ici d'une analyse approfondie de travaux récents dont un bref compte rendu a été donné par R. L. Ford et F. E. J. Girling concernant l'emploi d'amplificateurs linéaires utilisés en liaison avec des éléments à réaction passive pour simuler les inductances à faible perte. Cette étude décrit également quelques circuits pour la simulation de la capacité et aborde, d'autre part, la théorie des deux types en question.

Multivibrateur monostable suivi par un émetteur par R. Kitai

Résumé de l'article aux pages 642 à 646

Ce type de multivibrateur est largement utilisé d'une façon particulière suivant laquelle la durée des impulsions de sortie est en fonction quasi linéaire de la tension à la base de l'un des transistors. Toutefois, cette zone de mode linéaire comporte des limites supérieures et inférieures. Il est démontré qu'un étage peut être conçu à l'aide d'une famille universelle de courbes, à condition que le rapport de résistance soit égal au rapport de tension continue. Cette restriction n'entraine qu'une tres faible perte de rendement. La zone de fonctionnement la plus linéaire est établie, compte tenu des tensions aux bornes des transistors. Cette théorie a pu être confirmée par la pratique.

La pentode pilotée par anode en tant qu'électromètre par J. H. B. Gould

Résumé de l'article aux pages 647 à 649

L'auteur de cet article démontre qu'un tube ordinaire du type "récepteur" peut être utilisé avec succès dans l'étage d'entrée d'un électromètre. L'anode du circuit spécial inversé fait fonction d'électrode de commande cependant que la sortie est prise d'une résistance de charge dans le circuit de grille.

Mesure de bruit à faible fréquence dans la bande p M.O.S.T. par P. W. Fry

Résumé de l'article aux pages 650 à 653

Le niveau de bruit de la bande p M.O.S.T. MTO-1, utilisée en tant qu'amplificateur, a été étudié dans la gamme de 12Hz à 20kHz.

Le problème de la mesure de la tension de bruit aux fréquences réduites a été surmonté par des circuits permettant de compter la proportion de cycles dans une amplitude donnée. Cette proportion est rattachée par une simple équation à la tension de bruit efficace.

Les résultats indiquent une tension de bruit d'entrée équivalente suivant une loi 1/f, ainsi qu'une entrée de courant de bruit dont on peut ne pas tenir compte sauf aux basses fréquences et aux impédances de source supérieures à 100MΩ.

Un système d'appel à impédance transistorisé par H. N. Wroe

Résumé de l'article aux pages 654 à 655

Les lignes téléphoniques desservant les centrales H.T. doivent être fréquemment protégées par l'emploi de transformateurs d'isolement ce qui exclut les méthodes normales d'appel par courant continu L'auteur décrit un systeme d'appel utilisant des semi-conducteurs qui constituent une solution pratqiue et économique au problème. Il est basé sur le principe qu'en incorporant l'impédance de ligne au central téléphonique dans le circuit d'un oscillateur basse fréquence, les changements dans l'impedance de ligne de la centrale H.T. peuvent permettre ou empêcher les oscillations. L'oscillateur entraîne un détecteur de correction des impulsions et la combinaison de ces deux appareils traduit les changements d'impédance de ligne en signaux normaux de circuit de courant continu.

Un modulateur de puissance à très haute fréquence par M. B. Clark

Résumé de l'article aux pages 656 à 658

Le circuit modulateur dont il est question dans cet article présente les caractéristiques suivantes:

Signal d'entrée: 19—43 MHz. 50mW. 50Ω.

Fréquence porteuse: 109 MHz ou 140 MHz.

Signal de sortie: 66—184 MHz. 50Ω suivant les bandes latérales et la fréquence porteuse utilisée.

Ces largeurs de bande sont réalisées avec une sortie de niveau d'environ ±1 dB et une perte par insertion typique de 8dB.

Le modulateur annulaire conventionnel n'ayant pu fonctionner au niveau de puissance voulue, un circuit modifié fut donc utilisé. Le niveau minimum de la porteuse est obtenu pour un niveau de signal donné, soit $V_c = V_s$ et $I_c = 2I_s$. Ces données s'avèrent exactes dans la pratique. Pour le signal d'entrée de 25 mW la puissance porteuse minima est de 50mW pour la modulation correcte.

Les valeurs nominales des diodes sont fonction des niveaux du signal et de la porteuse et les niveaux de signal maximum sont calculés pour deux types de diodes. Des détails pratiques des transformateurs de modulateurs et des diodes sont examinés.

L'étude d'amplificateurs différentiels à courant continu et à faible dérive par S. K. Bhola et R. H. Murphy

Résumé de l'article aux pages 659 à 661

Le rendement des amplificateurs à courant continu est limité par leur dérive en fonction de la température. Une des méthodes pour réduire cette dérive réside dans la stabilisation par interrupteur ainsi que par l'emploi d'amplificateurs différentiels accouplés à l'émetteur. La stabilisation par sélecteur comporte des limites en ce qui concerne la tension décalée et le coefficient de température de la tension décalée ainsi que des restrictions concernant le bruit et la largeur de bande, tandis que l'emploi d'amplificateurs différentiels accouplés à émetteur exige une attention particulière aux problèmes d'équilibrage. Une connaissance plus approfondie des facteurs causant la dérive a donné lieu à la mise au point de dispositifs et de méthodes de circuit pour les amplificateurs différentiels qui réduisent la dérive à moins de 1μV/°C et présentent l'avantage du fonctionnement sans bruit et de dimensions réduites. On a pu démontrer la capacité de dérive réduite de ces amplificateurs par l'emploi de nouveaux dispositifs et de méthodes inédites.

Un convertisseur binaire/décimal à circuits intégrés par K. J. Dean

Résumé de l'article aux pages 662 à 664

Cet article décrit le dessin d'un convertisseur binaire décimal basé sur le contrôle logique d'un régistre de décalage. Ce convertisseur comporte des circuits intégrés dont la complexité n'augmente linéairement que suivant le nombre de chiffres binaires devant être décodés.

Analyse du suiveur d'émetteur à paire complémentaire par P. Mars

Résumé de l'article aux pages 665 à 667

Cet article décrit l'analyse matricielle d'un suiveur d'émetteur à paire complémentaire et des paramètres de circuit généraux qui en découlent. Le circuit témoigne de meilleures caractéristiques que le suiveur d'émetteur classique à un seul transistor et il possède une bonne stabilité de fonctionnement au courant continu.

Etage de base simple à la masse pour l'emploi avec les cables coaxiaux par B. L. Hart

Résumé de l'article aux pages 668 à 670

Tant dans les circuits analogiques à haute fréquence que dans les circuits numériques il est parfois nécessaire d'injecter un signal à un étage de transistor à travers un câble coaxial. Cet article décrit un circuit approprié utilisant un double transistor. La réalisation et le rendement de ce circuit sont également étudiés.

Instrument de mesure de la fonction de répartition de la densité d'amplitude par D. W. H. Hampshire

Résumé de l'article aux pages 671 à 673

Cet article décrit une méthode très simple pour mesurer la fonction de répartition de la densité d'amplitude de n'importe quelle forme d'onde utilisant un minimum d'équipement. L'article donne des résultats de mesure indiquant l'efficacité de cette méthode dans l'estimation de distorsions très réduites.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 682 bis 687

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Elektronisches Kombinationsschloss

Dynastat Electronics Ltd, 495 Green Lanes, London, N.13

(*Abbildung Seite 682*)

Dieses neue Schloss stellt einen wesentlichen Fortschritt in der Erzielung vollständiger Sicherung dar und ist nicht nur ein Kontrollmittel für die offensichtlicheren Anwendungen wie z.B. Tresore, Stahlkammern, Belegablagen usw., sondern auch für breite industrielle Gebiete. Das Instrument kann z.B. die Hauptsicherung von Datenverarbeitungsanlagen oder Einstellung von automatischen Maschinen, Magnetbandspeichern und gefährlichen Verfahren übernehmen. Es kann tatsächlich überall dort eingesetzt werden, wo industrielle oder kommerzielle Ausrüstungen gegen unbefugte Eingriffe stark gesichert werden müssen.

In dem Instrument kann man eine regellose Kombination von vier, acht oder mehr Ziffern wählen, die in einen Speicher eingegeben und dort als gewählte Kombination festgehalten werden. Danach bleibt der Zustand der durch das Instrument geschützten Ausrüstung sicher, und jeder Versuch, sie mit einer anderen als der korrekten Kombination freizugeben, löst die Alarmschaltung aus.

Die Freigabekombination, die zwischen zwei oder mehr Personen aufgeteilt werden kann, wird jedesmal neu und regellos eingestellt, wenn das Instrument die Ausrüstung abschliesst. Das permanenten oder halbpermanenten, vorgewählten Kombinationen innewohnende Sicherheitsrisiko besteht daher nicht.

Die gewählte Kombination kann nicht durch irgendwelche Herleitungsverfahren gefunden werden; keinerlei Handhabung des Wählers verrät in irgendwelcher Weise die Annäherung zur korrekten Position.

Die Abbildung zeigt ein für Vorführzwecke hergestelltes Schloss mit vier Ziffernwählern. Das Instrument kann aus Standardmoduln oder entsprechend Anwenderwünschen zusammengebaut werden. In der Praxis geben jedoch achtstellige Wähler einen zufriedenstellenderen Schutz, und ein Eindringling hat nur 1×10^{-8} Chancen, die richtige Kombination zu wählen.

Bei Wahl jeder anderen Kombination bleibt die geschützte Ausrüstung nicht nur weiterhin gesichert, sondern es wird auch ein Alarmsystem betätigt.

Die Sicherheit dieses Instrumentes kann durch keinen elektrischen oder mechanischen Angriff gebrochen werden; selbst wo es von der geschützten Ausrüstung entfernt installiert wird, kann weder Unterbrechen noch Kurzschliessen der Verbindungsleitungen oder Einspeisen von Signalen irgendwelcher Form etwas anderes bewirken, als den Alarm zu betätigen.

Das elektronische Schloss benötigt nur während des Ab- und Aufschliessens Energie und ist während der Sicherungsperiode völlig unabhängig vom Netz. Netzschwankungen oder Netzausfall haben daher auf die Sicherung keinen Einfluss.

EE 98 751 *für weitere Einzelheiten*

Lochstreifenleser

Electrographic, 38 Corwell Lane, Hillingdon, Middlesex

(*Abbildung Seite 682*)

Ein neuer fotoelektrischer Lochstreifenleser wurde von Electrographic angekündigt. Das Gerät hat die Typennummer 800, kann pro Sekunde 100 Zeichen von je 8 Bit lesen und an jedem gewünschten Zeichen stoppen.

Der Leser kann Lochstreifen jeder genormten Breite von 5 bis 8 Spuren plus Vorschubspur verarbeiten. Der Lochstreifen wird durch ein von einem Schrittmotor getriebenes Kettenzahnrad vorwärts bewegt, was zwangsläufigen Antrieb und schnelles Ansprechen gewährleistet.

Eine neue optische Anordnung gewährleistet, dass jeder Lochstreifen mit hohem Ein-Ausverhältnis gelesen werden kann. Eine Fotozelle in der Position des Zuführungsloches garantiert genaues Ausrichten der Leseposition.

Die Konstruktion verhindert Verschmutzung der Optik mit Staub und dergleichen. Der zu lesende Code kann in der Leseposition visuell gesehen werden.

Die Niederspannungslampe hat einen robusten Heizfaden und wird zur Verlängerung der Lebensdauer unterbelastet.

Den Lochstreifen kann man durch einfache Umkehr des Schrittmotorbetriebs in beliebiger Richtung zuführen.

Ein Trennverstärker für die Fotozelle und eine Motortriebplatte sind auch auf einem Chassis zu haben, das in das "Imhof"-Standardgestell für gedruckte Schaltungen passt.

EE 98 752 *für weitere Einzelheiten*

FS-Laufzeitleitung

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

(*Abbildung Seite 683*)

Vorteile einer von Mullard eingeführten neuen voreingestellten Ultraschall-Laufzeitleitung für PAL-Farbempfänger sind Vereinfachung der Dekoderschaltungen und Entfallen der Einregelung der Laufzeit nach Einbau.

Ausser diesen besonderen Vorteilen der neuen Mullard-Leitung kann ihre Anwendung in PAL-Empfängern Entwurfsprobleme in anderen Teilen des Empfängers vereinfachen und die Bildqualität verbessern.

Die ersten der neuen Laufzeitleitungen DL1 aus der Fertigung werden im Laufe dieses Jahres zur Auslieferung kommen; Mustermengen für Verwendung in Entwicklungsempfängern sind aber bereits greifbar.

Die als gefaltete, reflektierende Glastype ausgeführte Leitung ist leichter anzuwenden als andere für diesen Zweck entwickelte Leitungen, da die erforderliche Laufzeit von 63,943 μs bei 4,4336 MHz durch zusätzliche Aussenschaltungen erreicht wird, ohne dass man auf Abgleich nach Einbau zurückfallen muss.

Die Genauigkeit der Laufzeit wird während der Fertigung voreingestellt, und zwar erst durch Präzisionsschleifen der reflektierenden Flächen des Glasblocks und abschliessend durch Trimmen der eingebauten regelbaren Induktoren.

Das Eingangssignal zur Leitung wird an einen keramischen Schwinger gelegt, der das elektrische Signal in eine akustische Welle für Übertragung durch den Block umwandelt. Die in einen Punkt nahe am Eingang zurückreflektierte Welle wird durch einen anderen, ähnlichen Wandler wieder in ein elektrisches Signal umgewandelt. Da es sich um einen gefalteten reflektierenden Typ handelt, können die Wandler an Ort und Stelle eingebaut bleiben, während die Laufzeit justiert wird.

Wegen ihrer sehr niedrigen nominellen Einfügungsdämpfung von nur 10 dB gibt die DL1 an ihren Ausgangsklemmen ausreichend Spannung für die direkte Speisung eines Dioden-Demodulators ab Die Leitung ist für Verwendung zwischen merklich rückwirkungsfreien 150-Ω-Quellen ausgelegt

Für Empfängerhersteller ist von Bedeutung, dass die DL1 komplett mit Schaltungsinformation zur Erreichung von Farbdifferenzsignalen (R-Y), (B-Y) und (G-Y) mit konstanter Amplitude von einem einfachen Netzwerk angeboten wird, so dass die mit der Verteilung von Verstärkung und Dämpfung zwischen den Schaltungen zusammenhängenden Probleme nicht auftreten.

Die DL1 ist mit Anschlussstiften ausgerüstet, die direktes Einstecken in Druckschaltungsplatten ermöglichen.

EE 98 753 *für weitere Einzelheiten*

Digitaluhr

Alma Components Ltd, Park Road, Diss, Norfolk

(*Abbildung Seite 683*)

Diese neue Digitaluhr ist in erster Linie als Hauptprogrammschaltuhr gedacht. Ein netzgespeister Synchronmotor treibt eine Reihe von Alma-Schutzkontakt-Drehwählern in Kaskadenschaltung, wodurch eine Digitalanzeige in Stunden und Minuten entsteht. Die Einstellung der korrekten Zeit erfolgt einfach durch Betätigung von vier Drucktastenschaltern auf der Rückseite des Gerätes. Ein besonderes Merkmal dieser Konstruktion ist, dass der Anwender jede einzelne Minute im 24-Stunden-Zyklus der Drehwähler zum Schliessen oder Öffnen interner Kontakte benutzen kann und ihm somit eine Kombination von 1 440 Schalterkontakten zur Verfügung steht. Jeder der Kontakte ist für 1 A, 250 V mit einer Höchstnennleistung von 60 W bemessen. Der komplette Zyklus wird alle 24 Stunden wiederholt, aber man kann die Uhr auch durch Umlegen eines Schalters auf 12-Stundenbetrieb umstellen. Sie kann auch für jeden gewünschten Zeitzyklus, beispielsweise 48 Stunden, 60 Stunden usw., oder auch für das Zählen von Minuten und Sekunden geliefert werden. Die anpassungsfähige Konstruktion erlaubt Einbau von zusätzlichen Einrichtungen, um Sonderforderungen zu genügen, wie zum Beispiel Verzögerung der Aktion um die zum Ausdrucken erforderliche Zeit.

Die 24-Stunden-Uhr ist in Standardausführung entweder als in sich geschlossenes Gerät in Aluminiumgehäuse oder als Einbauchassis für andere Ausrüstungen lieferbar.

EE 98 754 *für weitere Einzelheiten*

Kern-Matrizenzusammenbau

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

(*Abbildung Seite 683*)

Mullard kündigt eine Serie preiswerter Einzelebenen-Kernmatrizen an, die speziell für kleine Tischelektronenrechner entwickelt wurden. In Speichern und Registern solcher Maschinen sind diese Matrizen billiger und kleiner als Halbleitersysteme; selbst bei Abschalten des Rechners vom Netz bleiben die Daten gespeichert.

Verwendung von Lithiumferritkernen gewährleistet betriebsmässige Konstanz über den breiten Temperaturbereich von 0...70°C. Betrieb mit sehr niedrigem Treiberstrom (190 mA) wird durch X-, Y- und Z-Wicklungen mit zwei Windungen erreicht, die den Treiberaufwand reduzieren und verhältnismässig einfache Treiber- und Ansteuerungsschaltungen gestatten. Verdrahtung erfolgt nach dem bekannten MIT-System.

Da der Kernzusammenbau in einen steifen Rahmen eingebaut und die Kerne mit Speziallack auf einer Papierschichtstoffunterlage befestigt sind, vertragen die Matrizen rauhe Behandlung. Sie sind für direktes Einlöten in gedruckte Leiterplatten bestimmt, und die Anschlussstifte sind nach dem 2,54-mm-Rastergrundmass ausgelegt.

Gegen geringen Aufschlag sind die Matrizen mit Epoxydrahmen, die feuchtere Zustände aushalten können, zu haben. Matrizenstapel mit bis zu vier Ebenen und seriengeschalteten Treiberdrähten sind auch lieferbar. Die folgenden Matrizen sind als Standardtypen erhältlich:

Typ Nr.	*Anzahl der Kerne*	*Aussenabmessungen*
AW 3526	16 × 16	82 × 82 mm
AW 3527	4 × 8 × 8	82 × 82 mm
AW 3529	4 × 12 × 12	102 × 102 mm
AW 3531	4 × 16 × 16	122 × 122 mm
AW 3532	32 × 32	122 × 122 mm

EE 98 755 *für weitere Einzelheiten*

Programmfeld

Sealectro Ltd, Farlington, Portsmouth, Hampshire

(*Abbildung Seite 683*)

Sealectro Ltd hat ein "Sealectroboard" mit durchsichtiger Deckplatte herausgebracht, das unbefugte Entfernung von Stiften verhindert und damit besonderen Sicherheitsanforderungen genügt. Die durchsichtige Deckplatte gestattet jedoch gleichzeitig ungehinderte Besichtigung des Programms und der Stiftanordnung.

Ausserdem wird die durchsichtige Deckplatte für Anwendungen geliefert, bei denen das Programmfeld Vibrationen ausgesetzt ist, die Stifte losschütteln könnten.

Das schnurlose Programmfeld ist in Standard- und Sonderzweckausführungen lieferbar und gibt mit seiner praktisch unbegrenzten Anzahl von Kombinationen horizontaler und vertikaler Löcher elektrische Matrixanordnungen, die allen Anforderungen genügen.

EE 98 756 *für weitere Einzelheiten*

Mehrumläufige Präzisionsskalen

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

(*Abbildung Seite 684*)

Kynmore Engineering Co Ltd hat ihr Sortiment durch drei von der französischen Atoms-Gruppe hergestellte mehrumläufige, geeichte Präzisionsskalen in geschmackvoller Ausführung ergänzt.

Die neue Mikro-Ziffernblattskala (links auf der Abbildung) ist eine Analogskala für 10 Umdrehungen mit 50 markierten Teilungen je Umdrehung. Sie hat einen Durchmesser von 22 mm und ist 29 mm tief. Die neue Analogskala US45 (oben) für 15 Umdrehungen hat 100 markierte Teilungen je Umdrehung. Sie hat 45,5 mm Durchmesser und ist 30,5 mm tief. Die neue Miniatur-Digitalskala Modell Flash 22 (rechts) ist für 100 Umdrehungen ausgelegt und hat eine vergrösserte Anzeige mit zehn numerierten Teilungen je Umdrehung. Sie hat einen Durchmesser von 23,4 mm und ist 35,2 mm tief.

Alle drei Skalen sind für Achsendurchmesser von 6 mm ausgelegt und mit Bremshebeln ausgerüstet, so dass sie in jeder Position festgestellt werden können.

EE 98 757 *für weitere Einzelheiten*

Direktschreiber

Honeywell Controls Ltd, Brentford, Middlesex

(*Abbildung Seite 684*)

Der neue Honeywell-Visicorder 2106 wurde im Lanarkshire Werk der Firma nach einem äusserst strengen Pflichtenblatt als preisgünstiger Direktschreiber mit vielen Einrichtungen, die man normalerweise in Instrumenten dieser Preisklasse nicht erwartet, entworfen und konstruiert. Das kompakte Gerät hat einen 152 mm breiten Registrierstreifen für die Ausgänge von 12 Kanälen und zwei weiteren Hilfskanälen.

Ein unvergleichlicher Vorteil dieses Gerätes ist die Möglichkeit, die UV-Lampe—einerlei ob warm oder kalt—sofort zu starten. Der 2106 kann nicht nur eine Mehrzahl von Variablen auf demselben Streifen registrieren, sondern gibt auch durch Standardeinrichtungen Raster- und Zeitlinien sowie Fernsteuerung.

Weitere Konstruktionsmerkmale sind u.a. eine Wahl von Streifenvorschüben im Gesamtbereich von 1 cm/min bis zu 300 cm/min durch Getriebe und für jedes Galvanometer getrennt einsteckbare Widerstandsplatten, mit deren Hilfe sich zur Vermeidung zeitraubender Rüstzeit Bibliotheken von Dämpfungs- und Anpassungsnetzwerken aufbauen lassen. Die Schreibgeschwindigkeit ist höher als 125 cm/ms, und eine im Sonderauftrag lieferbare Längenbegrenzung der Registrierung vermeidet Papierverschwendung. Der 2106 kann bis zu 45 m Registrierstreifen zwischen 76 und 152 mm breit aufnehmen.

Durch thermostatisch geregelte Beheizung des Galvanometerblocks wird Konstanz der Leistungen erzielt. Jedes der Galvanometer lässt sich in Sekunden einsetzen oder ausbauen.

Für dieses in Grossbritannien gebaute Instrument und für die neuen Galvanometer der Honeywell-Baureihe M, die darin Verwendung finden, erwartet man weltweiten Absatz.

Der Spiegel dieses hochfrequenten, flüssigkeitsgedämpften Galvanometers taucht nicht in das Dämpfungsmittel ein. Die Ablenkung des Lichtbildes über den Registrierstreifen wird daher nicht durch die Streuung des ultravioletten Lichtes in der Flüssigkeit beeinflusst. Die Galvanometer kombinieren hohe Empfindlichkeit und Stabilität mit flachem Frequenzgang von 0...13 kHz.

EE 98 758 *für weitere Einzelheiten*

Frequenzkonstanz

Hewlett-Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Abbildung Seite 684)

Über 50 kHz kann mit dem neuen Hewlett-Packard-Synchronisierungsgerät Modell 8708A für Messender eine wesentlich verbesserte Frequenzkonstanz erreicht werden.

In dem besonders für die neuen Ausführungen der HP-Messender 606 und 608 entwickelten Gerät wird zur Erzeugung von Testsignalen mit einer Konstanz innerhalb 1×10^{-7} über 10 Minuten eine Phasenmitnahmetechnik angewendet. Damit wird die früher erreichbare Konstanz gegenüber derjenigen der in diesem Frequenzbereich arbeitenden Standardmessender um mehr als 250× verbessert.

Automatische Synchronisierung wird für jede Frequenz im Bereich der verbesserten Messender 606B (50 ... 65 MHz) oder 608F (10 ... 455 MHz) erreicht. Die verbesserten Messender bieten ausser der erhöhten Konstanz auch die Möglichkeit für Frequenz- wie auch Phasenmodulation und für Amplitudenmodulation mit geringem Klirrfaktor. Sowohl Ausgangsleistung wie Modulationsgrad bleiben bei Frequenzänderung konstant, die stabilisierte Ausgangsfrequenz besitzt eine hohe Spektralreinheit, und die Fähigkeit für Frequenz- und Phasenmodulation ist äusserst linear.

EE 98 759 *für weitere Einzelheiten*

Hochspannungs-Stromversorgungen

Brandenburg Ltd, 139 Sanderstead Road, South Croydon, Surrey

(Abbildung Seite 684)

Die neusten Modelle aus dem BETA-Fertigungsprogramm sind die Hochspannungs - Konstantstromversorgungen 905 und 906. Sie geben beide 400 µA bei bis zu 60 kV ab, sind jedoch nur 178 mm hoch, 432 mm tief und 483 mm (19″) breit.

Die Ausgangsspannung kann durch Grob- und Feinregelung auf jeden Wert zwischen 6 und 60 kV eingestellt und mittels eines Messgerätes auf der Frontplatte überwacht werden. Anschlüsse für Strommessungen sind auf der Rückseite vorhanden. Modell 905 hat positive, Modell 906 negative Polarität.

Die sehr kompakte Konstruktion dieser Stromversorgungen ist hauptsächlich auf den Entwurf und die moderne Konstruktion des Hochspannungsteiles zurückzuführen. Der Festkörper-Komparatorverstärker ist als Druckschaltungsplatte ausgeführt und durch Steckverbindungen gekoppelt; die übrigen grossen Bauelemente werden durch einen robusten Legierungsrahmen getragen. Formgebung und Konstruktion sind fast mit denen der Modelle 705 und 800 identisch. Modelle 905 und 906 sind sowohl für Tischgehäuse als auch 19″-Gestelle geeignet. Die Leistung ist der bestehender BETA-Modelle sehr ähnlich, die allen industriellen sowie vielen Laboranforderungen genügen.

EE 98 760 *für weitere Einzelheiten*

Röhrenvoltmeter

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Abbildung Seite 684)

Modell VM76 ist das fünfte und neuste Voltmeter im Advance-Programm und ergänzt das Sortiment durch Erweiterung auf Wechselstrommessungen bei Frequenzen bis über 1 GHz und Gleichstrommessungen bis zu 1 000 V mit ±2 Prozent Unsicherheit bis zu 100 V.

Die verschiedenen Funktionen werden durch Drucktasten eingestellt. Bei Gleichstrombetrieb lässt sich auf diese Weise Anzeige mit Nullpunkt in der Mitte und entweder positive oder negative Polarität wählen. Widerstandsmessungen zwischen 0,02 Ω und 5 MΩ mit ±5 Prozent Unsicherheit in der Skalenmitte werden auf ähnliche Weise eingestellt. Typische Eingangsimpedanzen sind 7,5 MΩ bei 1 kHz und 150 kΩ bei 100 MHz, beide mit 1,5 pF Parallelkapazität.

Die meisten der Bauelemente sind auf nur einer Druckschaltungsplatte untergebracht, was ausgezeichneten Zugang zu den Bauelementen gewährleistet. Jedes Gerät wird mit Messspitzen und Anschlussschnüren mit Einzelstecker-Buchsenverbindung geliefert. Für alle Anwendungsmöglichkeiten als Voltmeter geeignetes Zubehör ist lieferbar.

Modell VM76 hat einen Temperaturbereich von 0...+40° C und ist netzbetrieben. Wie andere neu eingeführte Advance-Instrumente ist auch dieses in ein robustes Modulargehäuse von 2/3 Standardgestellbreite eingebaut.

EE 98 761 *für weitere Einzelheiten*

Echolot-Messender

Elliott Instruments Co. Ltd, Dunmow, Essex

(Abbildung Seite 685)

Elliott Instruments Company Ltd kündigt neue Ausführungen ihrer Echolot-Messender an. Diese Geräte wurden besonders für den Ingenieur entwickelt, der Service an Bord ausführt, stellen jedoch auch in Labors eine zweckdienliche Signalquelle dar.

Die volltransistorisierten Geräte des Types UL3 erzeugen sinuswellenförmige Ausgänge im Bereich 7 ... 60 kHz und ein von 1 µV ... 1 V variables Signal. Das Instrument ist batteriebetrieben, robust und kompakt mit 203 × 165 × 51 mm Aussenabmessungen konstruiert und wiegt etwa 1,8 kg.

Mit vorgesehenen Überwachungseinrichtungen kann die Verstärkung der Echolotverstärker gemessen werden, wobei das Messinstrument des Gerätes direkt an den Verstärkerausgang gelegt wird. Es kann entweder für Nasspapierschreiber 10 V— bei 10 kΩ oder für Trockenpapierschreiber 500 V∼ anzeigen. Das Gerät ist auch ohne die Überwachungseinrichtungen als Typ UL3a lieferbar.

EE 98 762 *für weitere Einzelheiten*

Einbaufertige Kristalloszillatoren

The Marconi Co. Ltd, Chelmsford, Essex

(Abbildung Seite 685)

Die Marconi Company kündigt eine neue Serie einbaufertiger transistorierter Oszillatoren an, die in nur einem Einsteckbaustein untergebracht und bereits in laufender Produktion sind und Konstrukteuren eine preisgünstige Frequenzquelle mittlerer Konstanz bieten.

Man braucht nur eine 6-Volt-Stromversorgung für niedrige Stromentnahme, um einen leistungsfähigen betriebsfertigen Kristalloszillator zu haben. Die mit internationalem Oktal- oder B7G-Sockel ausgerüsteten Bausteine haben Frequenzen im Bereich 1 kHz bis zu 100 MHz. Die kompakteren B7G-Bausteine F3171 überstreichen den Bereich 115 kHz bis zu 100 MHz und haben den Vorteil, dass sie für höhere Temperaturkonstanz im Marconi-Thermostaten F3006 Verwendung finden können.

Der grössere Baustein F3170 mit Oktalsockel überstreicht den Bereich 1 bis 115 kHz und ist vollkommen in ein Rohrgehäuse aus Aluminium von 29 mm Durchmesser eingebaut. Die Einbauhöhe schwankt mit der Frequenz zwischen 79 und 127 mm, und die Oszillatoren für höhere Frequenzen mit B7G-Sockel haben eine Standard-Einbauhöhe von 79 mm. Diese Bausteine sind auch in einem Aluminium-Zylinder untergebracht, der jedoch 19 mm Durchmesser hat.

Die über einen Betriebsbereich von $-20° \ldots +70°$ C erreichbare Konstanz ist besser als 1×10^{-4}. Auf Wunsch kann die Frequenz des Bausteins durch einen zwischen einen Sockelstift und Masse geschalteten externen Drehkondensator abgeglichen werden.

Jeder Oszillator ist um ein Quarzkristall in einem Glasgefäss aufgebaut und hat eine Ausgangstrennstufe, die die Frequenzkonstanz für eine 10prozentige Änderung der Belastungsimpedanz unter 1×10^{6} hält. Die Ausgangsimpedanz ist 5000 Ω ± 10 % bei 10 V_{SS} Ausgangsspannung.

EE 98 763 *für weitere Einzelheiten*

Magnetkopf

Gresham Lion Electronics Ltd, Magnetic Recording Head Division, Lion Works, Hanworth Trading Estate, Feltham, Middlesex

Gresham Lion Electronics Ltd hat einen 9-Spur-Magnetkopf für Lesen nach Aufnahme entwickelt. Der für digitale Anwendung bestimmte Kopf gestattet sofortiges Prüfen der eingeschriebenen Information und ist speziell für Einsatz mit dem IBM-1600-Bit-System gedacht.

Für Kompatibilität mit der IBM-Ausrüstung kann der Kopf mit dem Schreibelement links, sonst aber auch rechts geliefert werden. Der Kopfspiegel wird zur Minderung des Langwellenrauschens profiliert und so angeschliffen, dass die Bandkanten nicht einschneiden können.

Der Kopf ist etwa 32 mm hoch, 26 mm tief und 25 mm breit. Die Azimuttoleranz ist ±0,0002 Radian und die der Erhebung innerhalb 0,0005 Radian. Die Spaltstreuung liegt innerhalb eines 0,002 mm breiten Bandes, und die Spaltlinien sind innerhalb 0,0002 Radian parallel. Die Lesespuren sind 1,02 mm, die Schreibspuren 1,12 mm breit. Die Trennung zwischen den Spaltlinien beträgt 3,8 mm. Für die Lesespaltbreite bietet Gresham die Wahl zwischen 5 μm und 2,5 μm, wobei die letztere für das 1600-Bit-System bestimmt ist.

Der Sättigungsstrom ist für 3M-Band T488 bei 300 Bits je Zoll und 99 Prozent Sättigung 66 mA_{SS}.

Weitere elektrische Eigenschaften sind:
(a) Lesespalt 5 μm. Lesen: Widerstand 3,7 Ω, Induktivität 600 μH. Schreiben: Widerstand 2,3 Ω, Induktivität 415 μH.
(b) Lesespalt 2,5 μm. Lesen: Widerstand 7,4 Ω, Induktivität 2,4 mH. Schreiben: Widerstand 4,6 Ω, Induktivität 1,6 mH.

Die Schreibkopfwicklung ist mittelabgezapft.

EE 98 764 *für weitere Einzelheiten*

Statischer Wechselrichter

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(Abbildung Seite 685)

Industrial Instruments Ltd hat ihr bestehendes Programm für statische TRANSIPACK-Wechselrichter durch das Modell 606/8L ergänzt. Dieses Gerät ermöglicht, Fernschreiber in bewegliche Nachrichtenverkehrsausrüstungen einzubegreifen, und liefert die für Betrieb des Motors, Betätigung der Zeitabschnittrelais und Erregung des Senders erforderliche Energie nach Anschluss an eine 24-V-Gleichstromquelle. Das vollstabilisierte Gerät gibt bei einer Höchstbelastung von 1 A +80 V und −80 V ab, ist jedoch auch für andere Signalspannungen lieferbar und mit komplettem Überlastungsschutz ausgerüstet. Es ist in einem freistehenden, brennemaillierten Gehäuse mit bequemem Zugang zu Batterien und mit einem Batterielader untergebracht.

Mit dieser Ausrüstung wird der Fernschreiber wirklich beweglich und hat daher viel mehr Anwendungsmöglichkeiten, beispielsweise im Baugewerbe, Ölfeldern, militärischem Einsatz, für Polizeikontrolle von Fahrzeugen usw.

Industrial Instruments Ltd stellt die verschiedensten statischen Wechselrichter her, die Netze mit 230 V∼ und 50 Hz von Gleichstromquellen simulieren. Typisch für diese ist der hier in einem Gehäuse für die Streitkräfte abgebildete TRANSIPACK-Typ 606/ST/300. Dieses Modell entnimmt einem Fahrzeug oder einer Ausrüstungsbatterie 24 V und gibt bei Dauerbelastung 300 VA bei 230 V∼ ab. Die Frequenzkonstanz ist besser als ±1 % und der Klirrfaktor zwischen 3 und 6 Prozent.

EE 98 765 *für weitere Einzelheiten*

Dünnfilmmonitor

Genevac Ltd, Pioneer Mill, Radcliffe, Manchester

(Abbildung Seite 686)

Ein neuer Quarzkristall-Filmdickenmonitor von Genevac Ltd in Radcliffe ist für das Messen der Gesamtdicke sowie der Aufdampfrate von vakuumbedampften Dünnfilmen ausgelegt. Für die Messung wird in geeigneter Position in der Abdampfvorrichtung ein Quarzkristall eingebaut, auf dem ein Teil des aus verdampftem Material bestehenden Stroms gesammelt wird. Durch Ausnutzung des Prinzips der linearen Änderung der Resonanzfrequenz mit zunehmender Masse des Monitorkristalls zeigt die Messeinrichtung die Änderung als Gleichspannung und ausserdem die Änderungsrate dieser Gleichspannung an. Eine Rückführung vom Instrument kann zur Regelung des Wärmeeingangs der Abdampfquelle und damit der Aufdampfrate Verwendung finden. Bei Voreinstellung des Endarbeitspunktes des Gerätes lässt es sich als elektromagnetischer Verschluss betreiben.

Die Ausrüstung kann mit sehr guter Reproduzierbarkeit zur Kontrolle und Regelung von Aufdampfungen in Labor und Fertigung eingesetzt werden und zeigt die aufgedampfte Masse in vier Bereichen mit Skalenendwerten von 1, 5, 10 und 50 kHz an. Zwischen aufeinanderfolgenden Aufdampfungen kann man durch Zurückdrehen wieder auf Null einstellen, und Reinigung des Kristalls ist erst erforderlich, wenn die Frequenzverschiebung insgesamt 60 kHz erreicht. Die Ausrüstung ist in zwei Gehäusen von je 229 mm Breite, 152 mm Höhe und 152 mm Tiefe untergebracht. Ein Gehäuse enthält alle erforderlichen Stromversorgungen und die Anzeige der aufgedampften Masse, das zweite mit der Anzeige der Aufdampfungsrate kann man getrennt bestellen. Ein Oszillator, der in der Nähe der Aufdampfvorrichtung aufgebaut wird, ist mit dem Quarzkristall durch eine koaxiale Vakuumdichtung verbunden.

EE 98 766 *für weitere Einzelheiten*

Rauscharmer Verstärker

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Abbildung Seite 686)

Ein neuer, durchweg mit Silizium-Transistoren bestückter Verstärker LA350 wurde von Brookdeal Electronics angekündigt. Es handelt sich um einen Laborverstärker hoher Leistung mit ungewöhnlicher Spezifikation, die allen bisher kommerziell bekannten voraus sein soll.

Herkömmlicherweise erreicht man eine gute Rauschzahl nur von einer diskreten Quellenimpedanz über eine schmale Bandbreite. Die Rauschzahl des LA350 liegt jedoch bei Quellenimpedanzen zwischen 800 Ω und 250 kΩ bei Frequenzen über 500 Hz unter 2 dB, so dass für die meisten Messungen das thermische Rauschen in der Quelle der begrenzende Faktor ist. Die angewendeten neuen Schaltungen erweitern ausserdem seine Fähigkeiten über die normalen Grenzen herkömmlicher rauscharmer Verstärker hinaus und geben diesem Gerät die Eigenschaften eines

Mehrzweckverstärkers hoher Qualität mit guter Linearität, breitem Frequenzumfang (mit zweckmässigem Rechteckfrequenzgang) und schneller Überlastungserholung.

Das Gerät hat verteilte Abschwächer in der Gegenkopplung, die über aufeinanderfolgende Stufen arbeiten, wodurch Überlastung des Verstärkers verhindert und das Rauschen um denselben Faktor wie das Signal geschwächt wird. Die Abschwächer werden durch einen Einzelknopf bedient, der 5-dB-Stufen ergibt. Die Höchstverstärkung ist 100 dB. Der Gesamtfrequenzbereich ist 3 Hz ... 300 kHz und kann durch Hoch- und Tiefpässe, die die 3-dB-Abfallpunkte durch aufeinanderfolgende Faktoren von 3 ändern, reduziert werden.

Weitere Merkmale sind: Ausgangsleistungsmesser, mit dessen Hilfe das Gerät als Wechselstrom-Mikrovoltmeter eingesetzt werden kann, schaltbare Erdung und wahlweise Ausführung für den Tisch oder Gestelleinbau.

EE 98 767 *für weitere Einzelheiten*

10-MHz-Impulsgeber

Vertrieb: Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Abbildung Seite 686)

Das in Grossbritannien durch Dynamco Instruments vertriebene Datapulse-Modell 101 ist ein wirklich hervorstechender Impulsgeber in Buchgrösse mit Aussenabmessungen von nur 89 mm Höhe, 216 mm Breite und 279 mm Tiefe und wiegt etwa 3,6 kg. Er bietet Impulsfrequenzen bis zu 10 MHz, eine Anstiegszeit von 5 ns und Ausgänge bis zu ±10 V.

Das Modell 101 weist Eigenschaften auf, die man normalerweise nur bei teureren Impulsgebern findet, u.a. Einzel- oder Doppelimpulsbetrieb, veränderliche Breite und Verzögerung bis zu 10 ms, vielseitige Steuerung der externen Steuerung und Torschaltungen, sowie einen Zeitsteuerungsbereich von 1000 : 1.

Zu den externen Toreinrichtungen gehören u.a. synchrones Auftasten für die standfeste Bildschirmdarstellung von Impulsstössen, asynchrones Auftasten voreilender Triggerausgänge als ununterbrochene Systemuhr, Koinzidenzauftasten für Impulskoinzidenz mit Öffnungseingang und einer externen Triggerquelle.

Durch Pegel und Flankenpolarität gesteuerte Triggerung nach der Oszillografenart ergibt eine Empfindlichkeit von 250 mV. Das Gerät kann praktisch von jedem Punkt auf der negativen oder positiven Flanke der Spannungswellenform des Wechselstromnetzes getriggert werden. Die Vielseitigkeit der Triggerung gestattet auch synchrone Vorabzählung externer Triggerimpulse (Vorabzählung der Folgefrequenz) für Anwendungen, bei denen Frequenzteilung erforderlich ist.

EE 98 768 *für weitere Einzelheiten*

Elektrolytkondensatoren

The Plessey Co. Ltd. Ilford, Essex

(Abbildung Seite 686)

Der Geschäftsbereich TCC der Plessey-Bauelementegruppe hat zur Befriedigung der Anforderungen von Konstrukteuren elektronischer Sonderzweckausrüstungen zwei neue Kondensatoren entwickelt.

Die Miniatur-Elektrolytkondensatoren "Elkomold" sind für Verwendung in Transistorschaltungen und Montage auf gedruckten Schaltungen bestimmt. Sie sind leicht, haben kleine Abmessungen, und das Kunststoffgehäuse macht Isolierhüllen überflüssig. Vier Gehäusegrössen nehmen eine breite Nennwertauswahl auf.

Die Gehäuse sind für vertikalen Einbau geeignet und mit der Polarität markiert. Konstruktion und Eigenschaften entsprechen denen der TCC-Standard-Miniaturelektrolytkondensatoren in Metallgehäusen. Die Kunststoffgehäuse sind aus dauerhaftem Polypropylen gepresst und ein Ende durch eine mit Gummi beschichtete Phenolharzscheibe geschlossen. Umkehrbare "Elkomolds" sind im Sonderauftrag lieferbar.

Kapazitätswerte sind zwischen 1 und 1000 μF, für die niedrigeren Werte mit Betriebsgleichspannungen bis zu 350 V lieferbar. Kondensatoren können ohne Leistungsverlust kontinuierlich im Temperaturbereich −25° C ... +70° C betrieben werden.

Die TCC-Elektrolytkondensatoren "Superlytic" wurden besonders für lange Lebensdauer in professionellen Anwendungen entwickelt. Rigorose Fertigungskontrolle und speziell ausgewählte Werkstoffe garantieren eine Betriebslebensdauer bis zu 10 Jahren.

Der Betriebstemperaturbereich ist −25° ... +85°C, und der Reststrom überschreitet nicht 0,01CV μA oder 5 μA, von denen jeweils der höhere Wert gilt. "Superlytic"-Kondensatoren werden in Aluminiumgehäusen mit PVC-Isolierhüllen geliefert.

Kapazitäten sind von 1 ... 3 000 μF mit Betriebsspannungen zwischen 3 und 250 V— lieferbar. Die Gehäuseabmessungen liegen zwischen 19 mm × 9,5 mm Durchmesser und 115 × 64 mm Durchmesser.

EE 98 769 *für weitere Einzelheiten*

Untersetzer-Zeitmesser

Vertrieb: High Volt Linear Ltd, 67 Dudley Street, Luton, Bedfordshire

(Abbildung Seite 687)

High Volt Linear Ltd kann nunmehr den neuen Untersetzer 430 und den Untersetzer-Zeitmesser 431 von ORTEC liefern.

Der Untersetzer 430 ist ein Sechsdekaden-Untersetzer in Form einer integrierten Schaltung mit in sich geschlossenem Diskriminator und Druckausgang. Das Gerät ist nach der neuen AEC-Norm für kerntechnische Instrumente TID-20893 konstruiert und nimmt in dem Modularsystem für Einschübe und Netzteile Modell 401/402 zwei der zwölf Positionen ein. Andere Merkmale sind eine 16-MHz-Zählfrequenz, interner Diskriminator, positive und negative Eingänge und Messzeiteinrichtungen. Bei Einbau einer Druckwerksteuerung 432 in das System wird Aufbau eines anpassungsfähigen Datenerfassungssystems für bis zu 50 Untersetzer mit Serienausdruck möglich.

Der Untersetzer-Zeitmesser 431 ist ein voreinstellbarer, druckender Sechsdekaden-Untersetzer in integrierter Schaltungstechnik, den man auch als genauen Zeitmesser einsetzen kann. Der 431 entspricht derselben mechanischen Spezifikation und Norm wie der oben angeführte 430. Zu den anderen Eigenschaften gehören eine Untersetzungsfrequenz von 4 MHz, eine wahlweise Stimmgabel, mit der 0,04 Prozent Genauigkeit erreicht werden kann, sowie Messzeiteinrichtungen. Wie beim Modell 430 können auch hier bis zu 50 Untersetzer-Zeitmesser mit der Druckwerksteuerung 432 ein anpassungsfähiges Datenerfassungssystem bilden.

EE 98 770 *für weitere Einzelheiten*

Relaisschütz

Londex Ltd, 207 Anerley Road, London, S.E.20

(Abbildung Seite 687)

Von besonderer Bedeutung für Werkzeugmaschinen und ähnliche Steuerzwecke ist ein von Londex Ltd in der Elliott-Automation-Gruppe hergestelltes neues Relaisschützprogramm.

Die neuen "Contour"-Schützen sind äusserst robust, beanspruchen wenig Raum im Schaltfeld und entsprechen allen Anforderungen der BSS 755:1956, CSA und CEE-Pflichtenblätter.

Die dreieckige Ausführung der Träger der Festkontakte ist ein für Geräte dieser Art einmaliges Konstruktionsmerkmal, da dasselbe Bauelement durch einfache Umkehrmontage von einem Ruhe- in einen Arbeitskontakt umgewandelt werden kann. Anwendung von Brückenkontakten ermöglicht Erstellung eines einpoligen Umschalters durch Verbindung einer Seite eines Arbeitskontaktes mit einer Seite eines Ruhekontaktes. Spulen lassen sich äusserst einfach und ohne Einfluss auf Kontaktjustierung auswechseln. Eine neue Spule für dieselbe oder eine andere Spannung wird nach Entfernung von drei Schrauben, die sowohl Spule wie Spulenabdeckung festhalten, eingesetzt. Der Betriebsmechanismus richtet sich selbst aus und gewährleistet gute Kontaktlagerung und guten Kontaktdruck.

Die Bestückung kann mit bis zu zwölf einpoligen Doppelruhekontakten im Vielfachen von drei erfolgen. Sie können entweder aus Ruhe- oder Arbeitskontakten oder einer Kombination derselben bestehen. Typische induktionsfreie Nennwerte sind 15 A bei 250 V~, 15 A bei 50 V—.

Für 50-Hz-Wechselstrom sind folgende Standardspulen lieferbar: 6 V, 12 V, 25 V, 115 V, 380 V, 415 V und 440 V. Spulen sind auch für andere Wechselspannungen sowie Gleichstrom lieferbar.

EE 98 771 *für weitere Einzelheiten*

Kompensationsschreiber

A.E.I. Electronics Ltd, New Parks, Leicester

(Abbildung Seite 687)

Das Baumuster 5 ist eine verbesserte Ausführung des Kompensationsschreibers für 102 mm breite Registrierstreifen, die AEI Electronics in Harlow für Thermoelemente, Widestandsthermometer, Dehnstreifen und verschiedene Signalgeber entwickelt hat.

Es werden durchweg Festkörper-Bauelemente verwendet, was Zuverlässigkeit, kompakte Bauart und niedrige Leistungsaufnahme gewährleistet. Baumuster 5 ist entweder als 1-Linienschreiber oder als Mehrpunktschreiber mit farbigem Punktdruck der Signale von zwei, drei oder sechs Kanälen lieferbar. Besondere Konstruktionsmerkmale sind u.a. ein integraler Verstärker-Netzanschluss; ein Eingangsteil, der schnelle Änderung von Bereich und Funktion gestattet; die Möglichkeit, bis zu sechs blind einstellbare Alarmschalter einzubauen; und externe Regelung des Papiervorschubs und des Druckabstandes.

Die absolute Genauigkeit des Gerätes ist innerhalb 0,5% und die Auflösung 0,2%. Die Einstellzeit ist eine Sekunde, die Standard-Mehrpunkt-Druckgeschwindigkeit ein Punkt jede fünf Sekunden, ist jedoch auf Wunsch auf zwei Sekunden reduzierbar.

Der Schreiber ist für Einbau oder Tischverwendung ausgelegt, etwa 533 mm lang × 152 mm × 152 mm und in einem staubdichten Gehäuse untergebracht.

Alle Art und Bereich der zu messenden Signale bestimmenden Widerstände sind auf eine auswechselbare Bereich-Druckschaltungsplatte montiert. Für alle Bereiche wird eine maximale Eingangsquellenimpedanz von 1 kΩ empfohlen.

EE 98 772 *für weitere Einzelheiten*

Feuchtemesser

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, Yorkshire

(Abbildung Seite 687)

Für genaue Feuchtemessungen organischer und anorganischer Pulver und ähnlicher Substanzen ist erforderlich, dass die Proben immer dieselbe Dichte haben, da anderenfalls die Dielektrizitätskkonstante schwankt. Das ist der mittels Kapazitäts-Feuchtemesser bestimmte Parameter. Soweit bekannt ist, ist der neue Shaw-Druck-Feuchtemesser das erste Instrument, das besonders zur Erfüllung dieser Voraussetzung entwickelt wurde.

Das Instrument besteht aus zwei Teilen, nämlich einem Gehäuse mit der Elektronik und einer auf dasselbe montierten handbetätigten Druckvorrichtung mit den Elektroden. Die den Stempel betätigende Mechanik ist so konstruiert, dass das maximale Hebelmoment mit dem gewünschten Höchstdruck zusammenfällt und das Instrument langzeitig auch von Frauen ohne Ermüdung bedient werden kann. Entsprechend den getesteten Materialien können 0 bis 54,4 kp Druck ausgeübt werden.

In 0...1000 unterteilte Mikrometerskalen 10gängiger Potentiometer für “Stand” und Bereich gestatten Einstellung geeigneter Werte. Anzeigen des Feuchtigkeitsgehaltes von z.B. Sofortkaffee, Malz Mehl, Tabak und Stärke mit 0,1% Genauigkeit oder besser lassen sich erreichen. Das in Festkörperschaltungstechnik ausgeführte neue Shaw-Instrument arbeitet mit VHF und benötigt keine Wartung.

EE 98 773 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein vielseitiger Mehrkanalzeitgeber von A. I. Nizan

Zusammenfassung des Beitrages auf Seite 634-638

Ein beschriebener elektronischer Zeitgeber ist mit sechs unabhängigen Zeitkanälen ausgerüstet. Der Bereich jedes Kanals umfasst 0,1 . . . 99,8 s; innerhalb dieses Bereiches ist die Zeit in Intervallen von 0,1 s vorwählbar. Die Unsicherheit entspricht praktisch der der Netzfrequenz; die Genauigkeit kann jedoch durch Einsatz eines genauen externen Taktgenerators erhöht werden. Der Generator kann gleichzeitig den Zeitmassstab beeinflussen und den Bereich des Gerätes erweitern. Ausserdem kann jeder der sechs Kanäle als Dreifachkoinzidenztor funktionieren. Der Zeitgeber kann entweder in Einmal-Betriebsart oder in zyklischer Betriebsart arbeiten. Das Gerät hat sich besonders in Analog- und Hybridrechnung als nützlich erwiesen.

Impedanzwandler mit rückgekoppeltem Breitbandverstärker von F. Butler

Zusammenfassung des Beitrages auf Seite 639-641

Der Kurzbericht von R. L. Ford und F. E. J. Girling über Arbeiten bezüglich der Anwendung linearer Verstärker mit zugeordneten passiven Rückkopplungselementen für die Nachbildung verlustarmer Induktivitäten wird besprochen und erweitert. Einige Schaltungen verwandter Art für die Nachbildung von Kapazitäten werden auch beschrieben und die Theorie für beide Typen umrissen.

Entwurf eines emittergekoppelten monostabilen Multivibrators von R. Kitai

Zusammenfassung des Beitrages auf Seite 642-646

Multivibratoren dieser Klasse werden weitgehend in einer Spezialform angewendet, in der die Ausgangsimpulsdauer in nahezu geradliniger Beziehung zur Spannung an der Basis eines der Transistoren steht. Diese geradlinige Betriebszone hat jedoch obere und untere Grenzen. Es wird gezeigt, dass unter der Voraussetzung, dass ein Widerstandsverhältnis einem Gleichspannungsverhältnis gleich gemacht wird, mit Hilfe einer Universalkurvenschar eine Stufe entworfen werden kann. Wenn durch diese Beschränkung überhaupt ein Opfer notwendig wird, so ist es nur sehr gering. Der geradlinigste Arbeitsbereich ist festgestellt, und Klemmenspannungen der Transistoren werden in Rechnung gestellt. Die Theorie wird durch Versuche unter Beweis gestellt.

Die anodengesteuerte Pentode als Elektrometer von J. H. B. Gould

Zusammenfassung des Beitrages auf Seite 647-649

In diesem Beitrag wird gezeigt, dass eine gewöhnliche Empfangsröhre erfolgreich in der Eingangsstufe eines Elektrometers Verwendung finden kann. Eine "umgekehrte" Spezialschaltung wird angewendet, in der die Anode als Steuerelektrode wirkt, während der Ausgang an einem Belastungswiderstand im Gitterkreis abgenommen wird.

Niederfrequente Rauschmessungen am P-Kanal-MOST von P. W. Fry

Zusammenfassung des Beitrages auf Seite 650-653

Die Rauscheigenschaften eines als Verstärker betriebenen P-Kanal-MOST MTO-1 werden im Bereich 12 Hz . . . 20 kHz untersucht.

Die beim Messen von Rauschspannungen bei niedrigen Frequenzen auftretenden Probleme wurden durch eine Schaltung überwunden, die Zählen des Anteils der Perioden mit Amplituden über einen vorgegebenen Wert ermöglicht. Das wird durch eine einfache Gleichung mit der effektiven Rauschspannung in Zusammenhang gebracht.

Ergebnisse weisen auf eine äquivalente Eingangsrauschspannung, die einem 1/f-Gesetz folgt, zusammen mit einem Rauschstromeingang, der ausser bei niedrigen Frequenzen und Quellenimpedanzen über 100 MΩ vernachlässigt werden kann, hin.

Ein transistorertes Impedanzwählsystem von H. N. Wroe

Zusammenfassung des Beitrages auf Seite 654-655

Fernsprechleitungen im Verkehr mit Hochspannungskraftwerken müssen oft durch Verwendung von Trenntransformatoren geschützt werden, was Anwendung der üblichen Gleichstrom-Wähltechnik unmöglich macht. Ein mit Halbleitern bestücktes Rufsystem, das beschrieben wird, gestattet eine praktische und wirtschaftliche Lösung der Aufgabe. Durch Einschalten der Leitungsimpedanz in die Schaltung eines niederfrequenten Oszillators im Fernsprechamt können in der Abschlussimpedanz im Hochspannungskraftwerk Änderungen hervorgerufen werden, die Schwingungen zulassen oder verhindern. Der Oszillator treibt eine Impulskorrektionsschaltung, und diese Kombination setzt Änderungen der Leitungsimpedanz in herkömmliche Gleichstrom-Schleifensignale um.

Ein UKW-Leistungsmodulator von M. B. Clark

Zusammenfassung des Beitrages auf Seite 656-658

Eine entwickelte Modulatorschaltung ermöglicht Abgabe von Breitband-Ausgangssignalen von 2,5 mW an 50 Ω Impedanz im Frequenzbereich 66 . . . 184 MHz. Die Bandbreiten werden bei innerhalb ±1 bB verlaufendem Ausgangspegel und einer typischen Einfügungsdämpfung von 8 dB erzielt.

Da der herkömmliche Ringmodulator nicht bei dem erforderlichen Leistungspegel arbeiten kann, wurde die Schaltung abgewandelt. Von einem gegebenen Signalpegel wird ein Mindestpegel des Trägers abgeleitet ($V_c = V_s$ und $I_c = 2I_s$). Das wird in der Praxis bewiesen. Der Mindestträgerpegel für korrekte Modulation mit einem Eingangssignal von 25 mW ist 50 mW.

Beziehungen zwischen den Kenndaten der Dioden und den Signal- und Trägerpegeln werden aufgestellt und für zwei Diodentypen Höchstsignalpegel berechnet. Praktische Einzelheiten des Modulator-Transformators und der Dioden werden besprochen.

Entwurf von Differential-Gleichstromverstärkern mit geringer Drift von S. K. Bhola und R. H. Murphy

Zusammenfassung des Beitrages auf Seite 659-661

Die Leistung von Gleichstromverstärkern wird durch ihre Drift mit Temperatur beschränkt. Alternative Wege zur Driftherabsetzung sind Chopperstabilisierung und Anwendung emittergekoppelter Differentialschaltungen. Die Nachteile der ersteren sind die Gegenspannung und der Temperaturbeiwert der Gegenspannung zusätzlich zu den durch Rauschen und Bandbreite bestimmten Einschränkungen. Die zweite Methode erfordert grosse Sorgfalt in bezug auf Anpassungsprobleme. Ein besseres Verständnis der für die Drift verantwortlichen Faktoren hat zur Entwicklung von Elementen und Schaltungen für Differentialverstärker geführt, mit denen die Drift bei kleinen Abmessungen und rauschfreiem Betrieb auf unter 1 μV/°C reduziert wird. Es wird versucht, die geringe Drift dieser mit neuerdings greifbaren Elementen bestückten Verstärker in neuster Technik anschaulich zu machen.

Ein Binär-Dezimal-Umsetzer mit integrierten Schaltungen von K. J. Dean

Zusammenfassung des Beitrages auf Seite 662-664

Der Beitrag beschreibt den Entwurf eines Binär-Dezimal-Umsetzers, der auf der logischen Steuerung eines Schieberegisters beruht. Der Entwurf wurde mit integrierten Schaltungen ausgeführt. Ein interessantes Merkmal des Entwurfes ist, dass die Verzweigtheit der Schaltung nur geradlinig mit der Anzahl der zu dekodierenden Bits zunimmt.

Analyse des mit einem komplementären Paar bestückten Emitterfolgers von P. Mars

Zusammenfassung des Beitrages auf Seite 665-667

Eine Matrix-Analyse des Emitterfolgers mit komplementärem Paar wird durchgeführt und die Parameter der Gesamtschaltung bestimmt. Die Schaltung hat bessere Eigenschaften als der konventionelle Emitterfolger mit Einzeltransistor und eine gute Ruhestromkonstanz.

Einfache basisgeerdete Stufe für Einsatz mit Koaxialkabel von B. L. Hart

Zusammenfassung des Beitrages auf Seite 668-670

Sowohl in hochfrequenten Analog- wie Digitalschaltungen ist es manchmal nötig, ein Signal über ein koaxiales Kabel zu einer Transistorstufe zu speisen. In diesem Beitrag wird eine geeignete Schaltung mit Doppeltransistor beschrieben. Eine Analyse des Entwurfes und der Eigenschaften wird gegeben.

Messausrüstung für die Verteilungsfunktion der Amplitudendichte von D. W. H. Hampshire

Zusammenfassung des Beitrages auf Seite 671-673

Der Beitrag beschreibt ein sehr einfaches Messverfahren zur Bestimmung der Verteilungsfunktion der Amplitudendichte einer beliebigen Wellenform mit Minimum-Geräteaufwand. Angeführte Messungen beweisen die Leistungsfähigkeit dieses Verfahrens in der Schätzung sehr kleiner Verzerrungen.