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Commentary

IT is just ten years ago that the phenomenon of stimulated emission was introduced by C. H. Townes at Columbia University who together with J. P. Gordon and H. J. Zeiger produced the first device for stimulated emission.

This device, which subsequently became known as a Maser (an acronym for *M*icrowave *A*mplification by the *S*timulated *E*mission of *R*adiation) used an ammonia gas beam and was an amplifier of remarkably low noise level, and although the power level was very limited there appeared to be significant prospects for its use in microwave communication.

Considerable attention the world over was being focused on the maser and one aspect which was being very closely studied was the possibility of extending this principle of stimulated emission into the optical region of the spectrum.

This was achieved in 1960 by T. H. Maiman at the Hughes Research Laboratories in America who succeeded in producing amplification in the red part of the optical spectrum using a ruby crystal and the device he produced became known as an optical maser.

At the time it was regarded as not more than a scientific novelty but it soon became apparent that it had the most intriguing possibilities for the future. In the period which followed a vast amount of research was concentrated on the optical maser—particularly in America and the Soviet Union—and today the optical maser, or laser as it is now known, shows every prospect of becoming what can be described as a revolutionary breakthrough in the world of electronics.

Research for new materials as alternatives to the ruby began at a very high pace and it was actually at the end of 1960 that A. Javan at the Bell Telephone Laboratories developed the helium-neon laser.

Today we have not only gas lasers but solid-state lasers as well extending not only into the optical region of the spectrum but also into the far infra-red region.

The laser of course is awaiting a much fuller understanding before it can be more widely exploited, and in this respect the present day position is similar to that of the transistor in its early stages of development, but there is no doubt that its future prospects are very bright.

It is therefore appropriate that a three-day conference on lasers and their applications should have been held in London and which was attended by about 460 delegates including some 95 overseas visitors.

At this conference, which was sponsored by the Electronics and Science Divisions of the Institution of Electrical Engineers, the Institute of Electrical and Electronic Engineers (U.K. and Eire Section) and the Institution of Electronic and Radio Engineers, some fifty-four papers were presented of which no more than four were from overseas.

The conference may therefore be regarded as a survey of British progress in the research and development of the laser and the conclusion one reaches is that while we are not attacking the problem on such a broad front as America we have nevertheless made some important contributions and the results so far are encouraging.

Much of the laser research in this country is being supported by the Ministry of Aviation who are ensuring proper co-ordination of effort in the various Government Laboratories, and considerable fundamental research into the new types of lasers and their applications is being carried out in such places as the Royal Radar Establishment, the Services Electronics Research Laboratory and the Royal Aircraft Establishment, together with other centres such as the National Physical Laboratory and the Research Department of the General Post Office at Dollis Hill.

Probably one of the most important applications of the laser is in the field of communications for the step from microwaves into the optical region offers an increase in bandwidth by the factor of 10 000 or more and for this reason both the General Post Office and the communications industry are actively studying the possibilities.

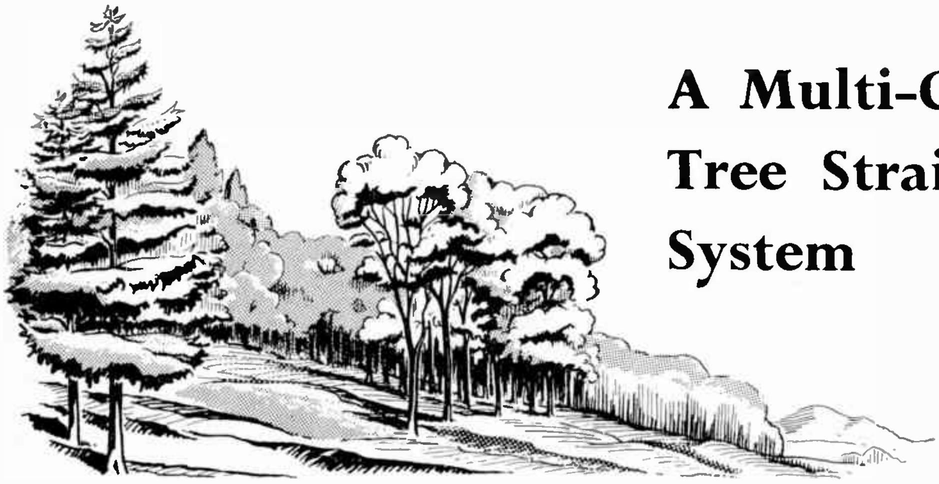
It is known that microwave links and coaxial cables are only just keeping pace with the present day demand for communication channels but with the still greater increase in the near future for circuits for telephone, television, data transmission and other facilities it is obvious that an entirely new means of communication must be obtained.

Although the realization is still some years ahead the laser appears to offer a solution in providing an optical system in two forms.

One form is by radiation in free space but due to the high attenuation of the optical beam through the atmosphere such a method would be limited to short ranges. However there are distinct opportunities for the development of portable line-of-sight communication links using laser beams with their high directivity.

Round the corner perhaps is the high capacity open system using balloons suspended above the weather but much nearer to hand and more promising is the alternative form using guided propagation, along optical fibres or evacuated pipes.

Other laser applications which are already in existence and fully operational are the micro-welding and micro-machining techniques together with a new type of range finder for surveying and possible military applications. Here the laser provides a method of fast accurate range finding of passive targets which overcomes some of the limitations of optical and radar measurements and quite recently a portable range finder of this type has been made available with a range up to 10km and an accuracy of 10 metres while at the other end of the scale the National Physical Laboratory has devised a method of checking one metre scales to an accuracy of half a millionth of an inch.



A Multi-Channel Tree Strain Recording System

By C. I. Sach*

This article describes a strain recording system developed specifically for tree response studies, but which has general application to laboratory and field work where a number of resistance strain gauge records are required. It is battery operated and is calibrated in such a way that it is not affected by supply voltage variations. The particular unit described has a non-linearity of less than 1 per cent of maximum output and is suitable for measuring dynamic strain having frequency components up to 10c/s.

(Voir page 798 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 805)

DURING the winter of 1963, a test, known as Operation Blowdown†, was held in North Queensland to investigate the effect of a large explosion on a tropical rain forest. An important part of this project was the investigation of the response of trees to the shock wave resulting from the charge, the particular quantity of importance being the dynamic strain occurring at the base of the trees. This required measurements on a number of trees at different distances from ground zero so that their deflexion, strain and damage could be correlated with the parameters of the blast wave. As the characteristic period of the trees was of the order of one second it would have been wasteful to use separate high frequency oscillograph or tape recorder channels, so a system using time division multiplex was employed. This system is based on a motorized switch of a type normally used for rocket telemetry. It allows 20 tree response signals to be multiplexed into one composite signal which can be recorded on one high speed channel. To obtain an accurate reconstruction of the original signals after demultiplexing, a sampling rate for each channel of 50 per second is used.

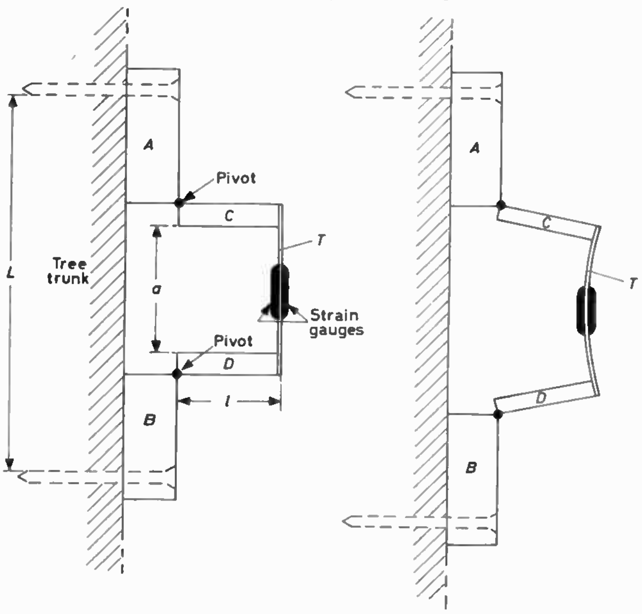
System Description

The strain is detected by a pick-up using strain gauge elements attached to the base of the tree and operated in conjunction with an electronic unit situated nearby. The output signal is then fed via a coaxial cable to the multiplexing unit which was located in the test area. The output of the multiplexer was then fed back to a recording bunker via a single coaxial cable. Rather than have the multiplexer motor running continuously, it was switched on by a sequencing unit in the bunker about 10sec before the charge was initiated. This allowed operators in the bunker adequate time to monitor and confirm its correct operation. The control signal used to start the multiplexer was also used to generate calibration command signals of 4sec duration which were sent to all the associated strain units. This signal was used to switch in calibrating components which effectively defined the electrical sensitivity of each recording channel.

Strain Unit

The jig to which the strain gauge elements are attached was designed by Raymond and Woolward of these laboratories. A diagram of the unit in the strain free condition is shown in Fig. 1(a). The strain gauges are attached to the flexible beam T whose ends are clamped to the rigid sections C and D. These in turn are pin jointed to the rigid sections A and B which are firmly attached to the tree. When that part of the tree to which the jig is attached is in tension, the jig assumes the shape shown in

Fig. 1(a). Strain jig in rest position
(b). Strain jig in strained position



* *Australian Defence Scientific Service.*
† *This was a joint Department of Supply—Australian Army project.*

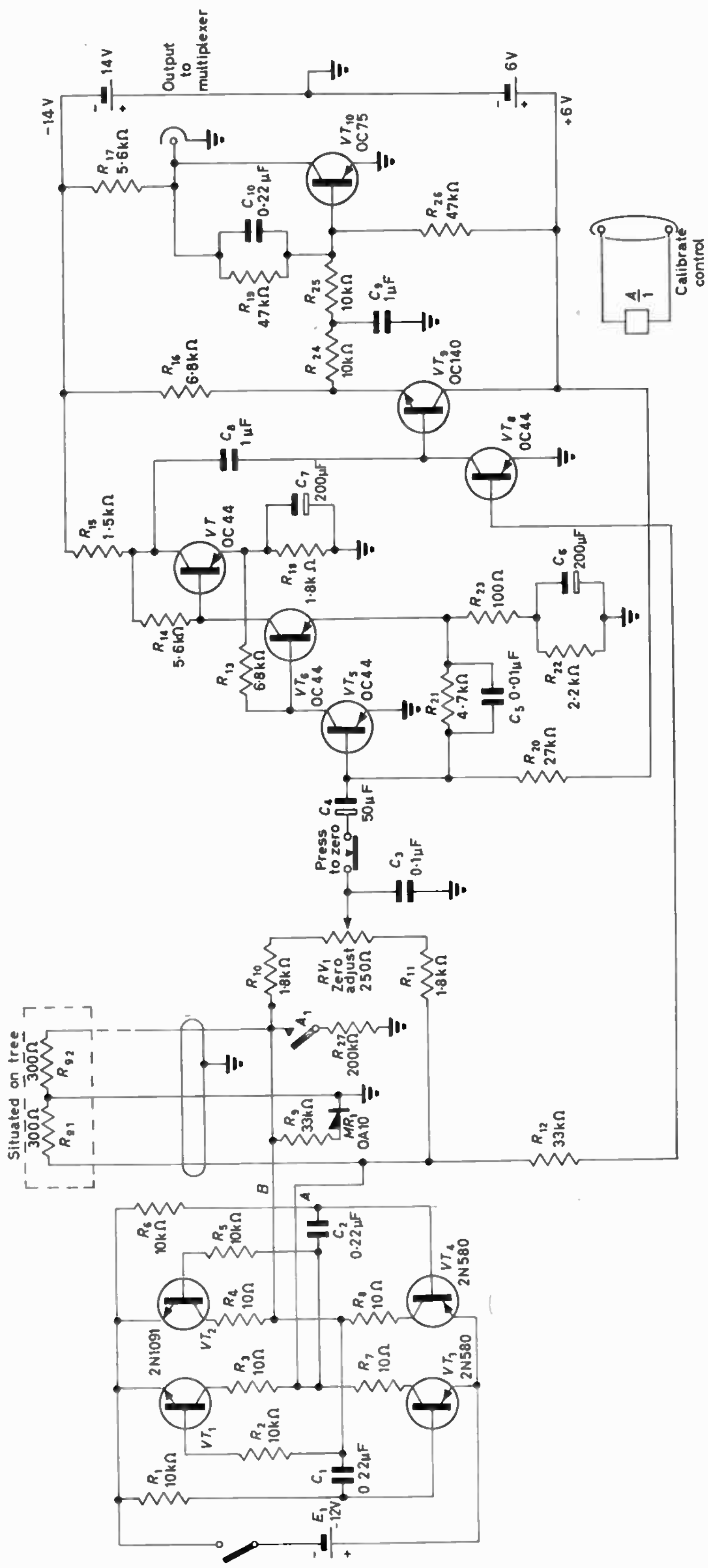


Fig. 2. Strain gauge unit

Fig. 1(b) where the beam T undergoes bending. This produces strains in the strain gauges which therefore change their resistances. The strain in the gauges is related to tree strain by the dimensions L, l and a which can be modified to obtain a desired strain magnification in the jig.

The strain gauge elements, designated R_{g1} and R_{g2} have nominal resistances of 300Ω and form a four arm bridge in conjunction with R_{10}, R_{11} and RV_1 situated inside the electronic unit whose circuit diagram is given in Fig. 2. The strain elements are arranged so that one undergoes compression while the other is in tension. In order that the common point of the gauges may be earthed, it is necessary to excite the bridge by an isolated supply. This normally involves the use of a transformer between the oscillator and the bridge, but in this case separate supplies were used, thus allowing the transformer to be dispensed with. The oscillator supply was allowed to float with respect to the amplifier supplies and hence direct drive to the bridge was possible. Such a system would not be suitable for high frequencies due to the effect of stray capacitances, but was quite acceptable at the chosen operating frequency of about 300c/s. The oscillator itself is a high efficiency multivibrator circuit using transistors VT_1, VT_2, VT_3 and VT_4 in a switching mode. In this way a square wave excitation voltage V_{BA} of approximately $\pm E_1$ volts was obtained, where E_1 is the oscillator supply voltage which in this case was 12V. A low impedance, high efficiency source is obtained because either VT_1 and VT_4 are saturated, thus allowing current to flow into the bridge at B and out at A or, on the alternate half cycle, VT_2 and VT_3 are saturated, thus allowing the current to flow in the opposite direction. The 10Ω series resistors R_3, R_4, R_7 and R_8 limit the currents occurring during the switching process where it is possible for all transistors to be partially conducting for a short period.

The output signal from the bridge is taken from the wiper of potentiometer RV_1 which acts as a 'zero adjust' control and fed to a low input impedance a.c. amplifier having a transfer resistance of $Z = 300$V/mA. This gain is quite stable, being defined by a current gain of about 50 in the feedback pair VT_5 and VT_6 and a transfer resistance of 5·6V/mA in the shunt feedback stage VT_7. The bandwidth, defined by R_{21} and C_5 in the current amplifier, is sufficient to pass the square wave signal. A shunt capacitance C_3 at the amplifier input helps to eliminate spikes produced by the oscillator in its transition state. As far as the amplifier is concerned the bridge acts as a voltage source given by:

$$E_b = E_1/2\ (\Delta R_g/R_g) \quad \ldots\ldots (1)$$

where ΔR_g is the resistance change in each strain element.

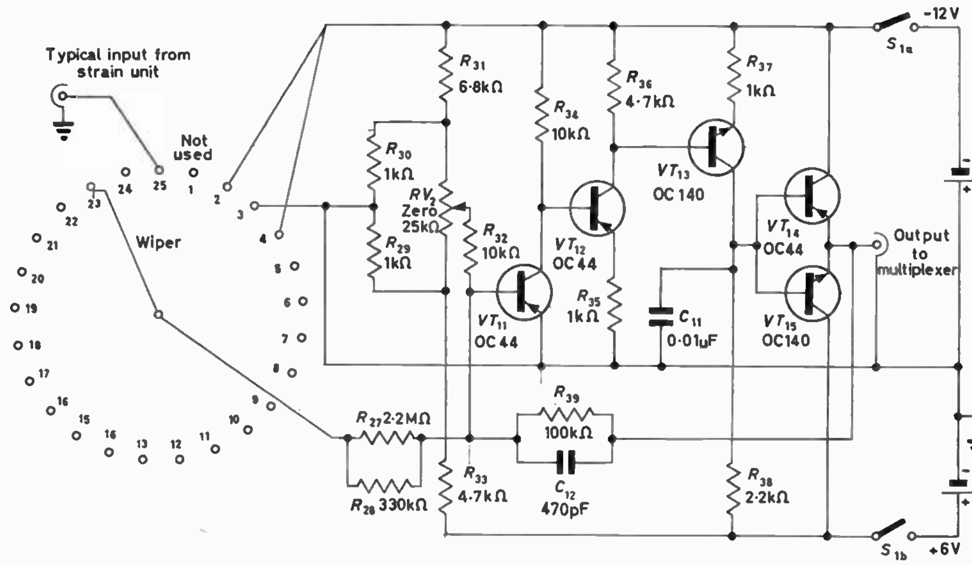


Fig. 3. Multiplexer sampler and buffer amplifier

In series with this generator is a resistance:

$$R_b = \frac{R_g + R_{10}}{2} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

where R_{10} here refers to the total resistance in the arm between point B and the output of RV_1 or between A and the output.

The current into the amplifier is then:

$$I_B = \pm \frac{E_1(\Delta R_g/R_g)}{R_g + R_{10}} \quad \ldots\ldots\ldots\ldots \quad (3)$$

The amplitude of the square wave occurring at the amplifier output is then given by:

$$E_0 = \frac{ZE_1(\Delta R_g/R_g)}{R_g + R_{10}} \quad \ldots\ldots\ldots\ldots \quad (4)$$

The signal is then phase sensitive detected using the transistor VT_8 as a clamping switch under the control of the oscillator via the base resistance R_{12}. To balance the load this control presents to the oscillator, a similar load (R_9 and MR_1) is connected to the other side of the bridge supply. The signal occurring at the collector of VT_8 is then the same as at the collector of VT_7 except that one limit is clamped to earth potential in the latter case. The signal then has a d.c. component which is amplified and filtered by VT_{10}. The d.c. component has a value:

$$E_{dc} = \pm \frac{ZE_1(\Delta R_g/R_g)}{R_g + R_{10}} \quad \ldots\ldots\ldots\ldots \quad (5)$$

which is amplified by a factor $R_{19}/(R_{24}+R_{25})$ to give an output variation of:

$$E_{out} = \pm \frac{R_{19}}{R_{24} + R_{25}} \; \frac{ZE_1(\Delta R_g/R_g)}{R_g + R_{10}} \quad \ldots\ldots \quad (6)$$

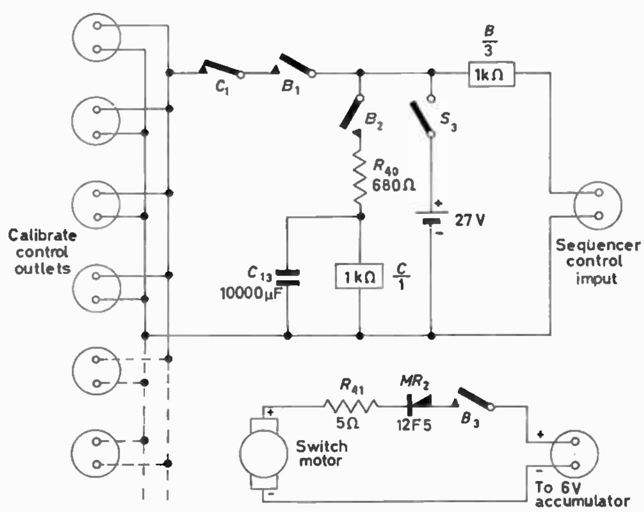
The values of the components as given in the circuit diagram were selected so that for a fractional resistance change of $\Delta R_g/R_g = 0{\cdot}075$ per cent a 3V output excursion would be obtained. The jigs associated with the strain elements are therefore designed to have a magnification factor which would produce this resistance change for the maximum strains expected to occur at the various trees.

Of the factors in equation (6) which determine the output signal, the least certain is the oscillator supply voltage E_1. Any change in this voltage produces a proportional change in the sensitivity of the system. Rather than resorting to a regulated supply it was decided to calibrate the electrical sensitivity of the system by connecting a resistance across one arm of the bridge (in this case R_{g2}) such that an output close to that expected from a maximum strain would be produced. This is achieved by using a 200kΩ resistor (R_{27}) which simulates a strain of 0·15 per cent in R_{g2} which thus gives the same output as complementary strains of 0·075 per cent in both R_{g1} and R_{g2}. This calibrating resistance, which is located in the electronic unit, is operated by relay A under the control of a voltage from the multiplexer. It is emphasized here that the strain gauge elements are assumed to be stable, and that changes in sensitivity are due to the electronics, i.e. oscillator amplitude and amplifier gain variations.

Before any test or calibration, the bridge must be balanced as follows. With the 'press to zero' switch depressed read the output d.c. level which should be about −6·5V. Then release the switch and adjust the zero control until the same reading is obtained. This can only be obtained when there is no bridge output signal, i.e. when the bridge is balanced.

To calibrate the device it is initially zeroed under

Fig. 4. Multiplexer control circuit



strain free conditions. Then a test jig is arranged to produce a known strain in the pick-up and the corresponding excursion of the output noted. Then the jib is removed and the calibration resistance switched in and the new output excursion noted. The ratio of these two readings can then be used to calibrate a strain recording taken at any subsequent time providing another calibration record is made close to the time at which the new strain recording is made. This is necessary since the important circuit parameters as given in equation (6) must have the same values for both records.

The two stages of filtering between the emitter-follower VT_9 and the output gives an effective step response time of 30msec. The non-linearity of the unit is less than 1 per cent of full scale over the whole range, the greater part of this error being due to the simple phase sensitive detector employed.

Multiplexer

The multiplexer is required to perform two functions. The first is its basic one of sampling the input channels and transmitting the composite signal back to the bunker while the second is its role of controlling the calibration process in the strain units.

The circuit diagram of the sampling unit and the buffer amplifier is shown in Fig. 3. There are 24 contacts on the switch, the output being taken from the wiper arm. In the switches used, contact 1 is connected internally to the motor frame while numbers 2, 3 and 4 are used for inserting synchronizing or identification pulses which help in separating the channels during the demultiplexing process. The outputs of the strain meters are usually in the range $-6\text{V} \pm 3\text{V}$ and these are fed to the remaining 20 inputs (5 to 24). The amplifier following the buffer has a stable input resistance of almost 300kΩ which is defined by R_{in}, the parallel combination of R_{27} and R_{28}. The overall gain of the amplifier is $-\frac{1}{3}$ which is defined by:

$$A = -(R_{39}/R_{in}) \quad \text{.............. (7)}$$

where R_{39} is the feedback resistance.

The forward transfer resistance of the amplifier is about 5 000V/mA so with R_{39} equal to 100kΩ an open loop gain of 50 is achieved which ensures a stable gain. The output which is taken from the emitter-follower pair VT_{14} and VT_{15} is used to drive the coaxial cable feeding the recorder which presents an effective terminating resistance of 10kΩ. The shunt capacitance due to the cable is of the order of 0·1μF. The bandwidth is determined by the parallel combination of R_{39} and C_{12} which give a rise time of about 60μsec which is helpful in reducing some inter-channel spikes. With the motor operating at 50rev/sec the sampling time per channel is nearly 1msec. The capacitance C_{11} is included to shape the open loop transfer function so that amplifier instability is avoided. The zero control is used to inject a direct current into the input circuit so that the output moves $0 \pm 1\text{V}$ for an input of $-6\text{V} \pm 3\text{V}$.

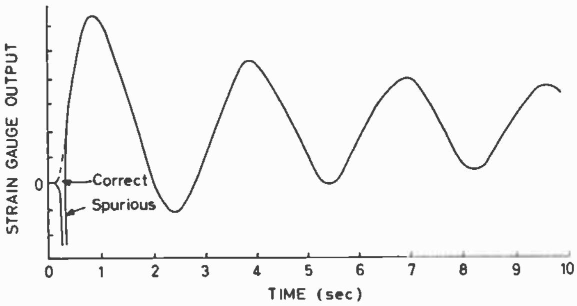


Fig. 5. Sample strain gauge record

The multiplexer motor is started, and the calibration control signals generated, using the system shown in Fig. 4. Normally the only supplies switched on are those for the amplifier and the relay supply switch S_3. A 6V accumulator is connected to the motor drive circuit but is ineffective while the contacts B_3 are open. When a short-circuit is applied by a sequencing unit to the control cable, relay B is energized, thus closing contacts B_1, B_2 and B_3. The latter allows the motor to start and the multiplexing process begins. With B_1 closed, 27V d.c. is fed to each of the calibrate control terminals which are connected by screened twin cable to all strain units. The closure of B_2 causes a rising voltage to be produced across relay C. The rate of rise is a function of R_{40}, C_{13} and the resistance of the relay coil and these are arranged so that the relay is energized 4sec after the original control signal is received. This then causes the C_1 contacts to open, thus removing the calibration control voltages.

Shock Protection

Since the units were required to operate in a severe shock environment some care was taken in housing them. In the case of the multiplexer, the unit was housed in a steel box which contained a heavy plate attached by rubber anti-shock mountings to the case. The bulk of the circuits were mounted on this plate except for the relays which were further isolated from the plate by anti-shock mountings. The case was housed in an outer wooden box using sponge rubber packing, the complete unit being located in a hole which was roofed and covered with earth.

Even these precautions were not completely adequate as the multiplexer motor appeared to falter; and spurious signals were introduced, at the time corresponding to the arrival of the shock front. It recovered, however, and was able to provide important information as shown in the typical demultiplexed waveform shown in Fig. 5. The negative spike is spurious while the final output differs from the start showing a residual strain in the tree.

Conclusion

A linear multi-channel strain recording unit has been described which is simple to operate and which can be calibrated electrically to reduce the influence of supply voltage and amplifier gain variations. It proved reliable in operation except for a tendency to be sensitive to severe shock conditions. This is a mechanical problem which could be overcome. Manual demultiplexing was used in this instance but in future work a more sophisticated system will be used. This uses a multiplexer, which has already been developed, whose speed is locked to the output of a 50c/s transistor oscillator, thus ensuring a constant sampling rate. Synchronizing pulses have been included in the waveform so that on replaying from tape a multiplexer can be phase locked to the histogram and operate gates to extract the pulses corresponding to various channels. These are stretched in box car circuits and then filtered to recover the original waveforms. The constant speed of the multiplexer eases the requirements on the lock-in circuit of the demultiplexer.

Acknowledgments

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Toroidal Ratio Transformers

Their Equivalent Circuits and the Measurement of their Parameters

By A. J. Binnie*, B.Sc., and T. R. Foord*, B.Sc., Ph.D., A.M.I.E.E.

This is a review article in which published analyses of equivalent circuits which represent the most useful types of toroidal transformer used in precision low-frequency bridge networks are collected together. Methods for measuring various parameters used in these equivalent circuits are also given. The effects of interwinding capacitances are not considered so that the circuits and equations given in this article do not necessarily hold at the higher audio frequencies.

(Voir page 798 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 805)

TOROIDAL transformers operating at low audio frequencies yield voltage or current ratios which are extremely close to their turns ratios. When such transformers are used as two of the branches in the four-arm Wheatstone type of a.c. bridge network, very high measurement accuracy and stability can be achieved. These advantages have been emphasized by a number of authors[1-6], who have also considered various aspects of the theory of the ratio devices and of bridge networks in which they may be usefully employed.

The closeness with which the terminal voltage or current ratio of any transformer agrees with its turns ratio depends on the leakage impedances of the windings, on the magnetization and core-loss currents, on the loading effect of the bridge network and the stray earth admittances of the transformer, and on the winding connexions, for example, on whether the windings are series aiding or in series opposition. These sources of error can be minimized by appropriate design. In general, it is necessary to use a toroidal magnetic core of high permeability strip wound material (Supermalloy or Supermumetal) with the ratio windings symmetrically disposed relatively to the core. This symmetry is closely fulfilled by winding a cable consisting of several insulated conductors, randomly stranded together, uniformly round the toroidal core. A winding of this type ensures that the coupling coefficients between pairs of sections are constant and very nearly equal to unity, e.g. 0·999 998.

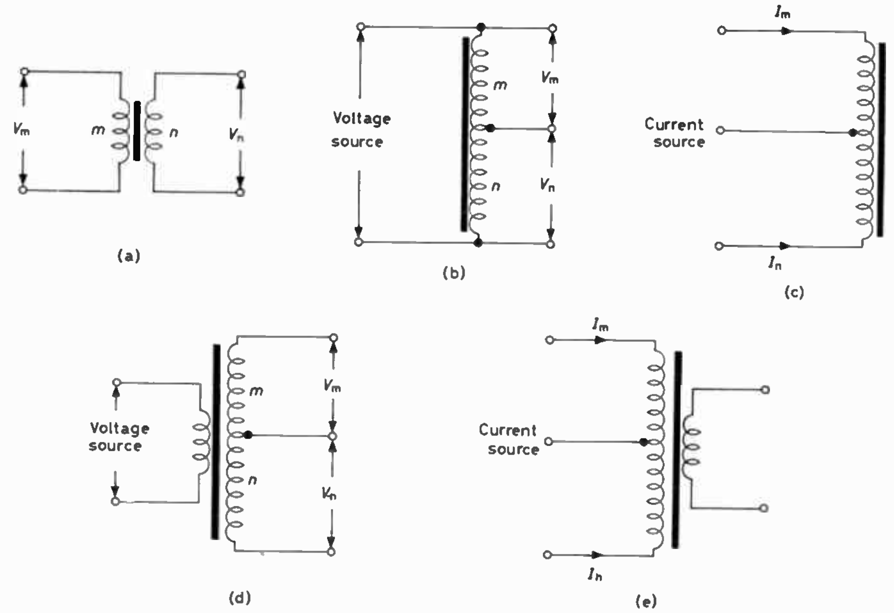

Fig. 1. Main types of ratio-transformer circoits
(a), (b) and (c) Two-winding transformers
(d) and (e) Three-winding tranformers

The different forms of ratio-transformer which are discussed are shown in Fig. 1 and are arranged to emphasize the quality between them.

When the auto-transformer is connected to a voltage source (Fig. 1(b)) and used in a bridge circuit to compare two impedances, the current in the impedances is fed directly from the source, i.e. there is no loading effect on the ratio transformer. The auto-transformer used in this way is, therefore, more suitable than the three-winding transformer (Fig. 1(d)) for comparing low impedances.

When a ratio transformer is connected as inductive ratio arms in a balanced bridge network, i.e. as if connected to current source (Fig. 1(c) and 1(e)), then in addition to providing a precise and stable current ratio the three terminals of the ratio winding are very nearly at the same potential. Thus if the tapping point is earthed, both detector terminals are also virtually at earth potential: earth admittances from these points, therefore, have very little effect on the bridge balance condition. Use is made of this feature in certain commercial bridges which permit the measurement of very small capacitances even when connected at the end of very long leads. Long leads of relatively high capacitance are also often necessary between the h.v. and l.v. components in the high voltage Schering bridge, and can be the cause of large measurement errors. This problem, however, is practically eliminated in bridge circuits incorporating current ratio transformers, such as that recently described by Baker[12].

Simple Theory of Ratio Transformers

The effects of the core-loss current and winding resistances are neglected in this initial simple theory but are

* *University of Glasgow.*

considered in later, more detailed, analyses of the various ratio-transformer equivalent circuits.

Consider a transformer with two ratio windings which consist respectively of m and n similar sections, with each section having a self-inductance, L, and suppose that the coupling coefficient between any two sections is constant and denoted by k: these conditions, also used by Hill and Miller[5], clearly imply fully symmetrical windings. Let the self-inductance of these windings be L_m and L_n, and let the mutual inductance between them be M.

Leslie[6] has shown that for such a transformer L_m and L_n can be represented by $(L_{mi} + l_m)$ and $(L_{ni} + l_n)$ respectively where l_n and l_m are the leakage inductances of these windings and where the product of L_{mi} and L_{ni}, equals the square of the mutual inductance between these two windings.

Therefore:

$$M = \sqrt{(L_{mi} L_{ni})} = mnkL$$

$$L_m = L_{mi} + l_m = L(m+m(m-1)k)$$

and

$$L_n = L_{ni} + l_n = L(n+n(n-1)k)$$

Therefore:

$$l_m = mL(1-k) \text{ and } L_{mi} = m^2kL$$

also

$$l_n = nL(1-k) \text{ and } L_{ni} = n^2kL$$

Thus the leakage inductance of each winding is proportional to the number of sections in the winding, i.e. proportional to the number of turns[17].

Two-winding toroidal transformers can be of three forms (Figs. 1(a), 1(b), 1(c)).

The first form with completely separate windings Fig. 1(a) is of little practical importance for precision measurement purposes as it provides the least accurately known voltage or current ratio. The main causes of this inaccuracy are that: (1) the magnetization and core-loss currents, which are voltage and frequency dependent, flow in one of the ratio windings (primary), so causing a potential drop even when the secondary winding is open-circuited; and (2) when the secondary is loaded, an additional drop occurs in both the primary and secondary windings. This may be shown by analysis of the circuit shown in Fig. 1(a) which gives:

$$V_n \simeq (n/m) V_m \left[1 - (1/m)\left(\frac{1-k}{k}\right) - \frac{j\omega L}{mkZ}(n(1-k)(1+km+kn-k)) \right]$$

when $k \simeq 1$. Z is the load impedance connected across the n-section winding. It follows that only if the coupling is perfect, i.e. if $k = 1$, that $V_m/V_n = m/n$.

The remaining two forms (Fig. 1(b) and 1(c)) are duals and can provide extremely precise voltage and current ratios. Thus if the two windings in Fig. 1(b), are connected together in series-aiding, and a voltage is applied across them so that voltages V_m and V_n are developed across the separate coils, then:

$$V_m/V_n = \frac{L_m + M}{L_n + M} = \frac{L(m+m(m-1)k)+mnkL}{L(n+n(n-1)k)+mnkL} = \frac{m((1-k)+(m+n)k)}{n((1-k)+(m+n)k)} = m/n$$

Similarly, if two windings are connected together as in Fig. 1(d) and currents I_m and I_n flow through them so that the voltages across the coils are equal, then:

$$I = I_m j\omega L_m - I_n j\omega M = I_n j\omega L_n - I_m j\omega M$$

therefore:

$$I_m/I_n = \frac{L_n + M}{L_m + M} = \frac{L(m+m(m-1)k)+mnkL}{L(n+n(n-1)k)+mnkL} = \frac{m((1-k)+(m+n)k)}{n((1-k)+(m+n)k)} = m/n$$

This form of transformer when used in bridge circuits is frequently referred to as the inductive ratio-arms.

This initial study of two-winding ratio transformers serves to indicate that, in order to obtain either voltage or current ratios, precisely proportional to the transformer turns ratio, one requires to have perfect coupling between coils in the case of the transformer with separate windings. The coupling coefficient between any two sections of the windings must be constant, in the case of the auto-transformer.

Three-winding ratio transformers comprise two ratio-windings and a third winding which is used either as a voltage supply winding or as a detector winding (see Figs. 1(d) and 1(e)). Let the coupling coefficient between either of the ratio windings and the third winding be k_i and the self-inductance of the third winding be L_p.

If a voltage is applied to the supply winding Fig. 1(d) so that a current I_p flows in it, then the open-circuit voltages, V_m and V_n, developed over the ratio windings, are:

$$V_m = I_p j\omega k_i m \sqrt{(L_p L)} \text{ and } V_n = I_p j\omega k_i n \sqrt{(L_p L)}$$

Hence:

$$V_m/V_n = m/n$$

Consider now the dual case (Fig. 1(e)). If currents I_m and I_n flow in the ratio windings the voltage across the detector winding will be zero when:

$$0 = I_m j\omega k_i m \sqrt{(L_p L)} - I_n j\omega k_i n \sqrt{(L_p L)}$$

Hence:

$$I_m/I_n = n/n$$

So it follows that to construct a three-winding voltage or current transformer whose voltage or current ratio is precisely proportional to the turns ratio of the ratio windings, the coupling coefficient between the supply or detector winding and any section of the ratio windings must be constant, the self-inductance of equal sections of the ratio windings should be equal, and the coupling coefficient between any pair of sections of the ratio winding should be constant.

The above conditions are satisfied with a transformer constructed thus. The supply or detector winding is wound symmetrically round a high permeability toroidal core and encompassed by magnetic and electrostatic screens. The ratio windings are wound on top of these screens, and these windings are constructed by winding a cable of randomly-stranded, insulated conductors, symmetrically round the toroid. Each conductor in the cable becomes one section of the ratio windings.

McGregor and others[14] have described a similar transformer whose ratio windings consisted of edge-wound strips of copper soldered into position over the screens. With such a construction the resistances of the ratio windings are very low.

Practical Theory for Two-Winding Ratio Transformers

Because of its practical importance the auto-transformer form of connexion (Figs. 1(a) and 1(c)) is discussed in this section. In the previous section only the effect of magnetic coupling between sections in the ratio windings is considered. Here, however, two additional factors, namely the winding resistance and the core power loss current, are taken into account. As stated earlier, interwinding capacitances are neglected in this theory although the measurement techniques described later do include the effects of these capacitances.

THE AUTO-TRANSFORMER

The magnetizing and core-loss currents flow in the ratio winding of the auto-transformer and so it is essential that the leakage impedances of these windings should be proportional to their respective member of sections for a precise open-circuit ratio to be obtained. This requirement is partly satisfied, i.e. so far as the leakage inductances are concerned, by the use of a stranded-cable winding on a high permeability core. It is, however, clearly necessary for the winding resistances also to be proportioned to the number of section. The stranded-cable form of winding construction tends to give equality of resistance in each section, though in practice it is difficult to reduce differences in resistance below 1mΩ. It is desirable, therefore, to minimize the core-loss current by using low-loss magnetic material for the core, e.g. Supermumetal strip.

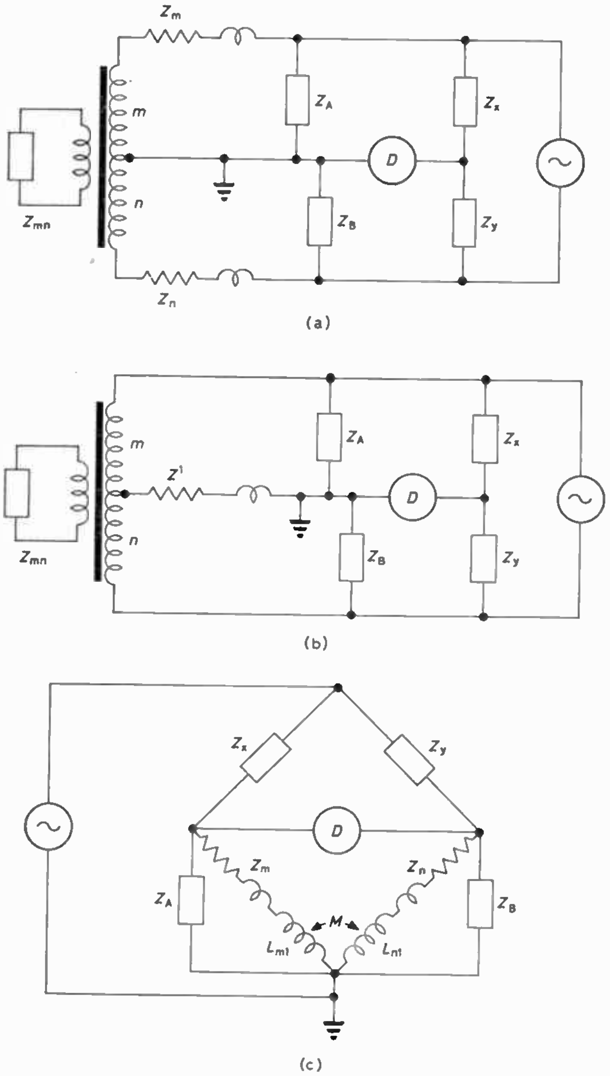

Fig. 2. Thomson's auto-transformer bridge networks

When an auto-transformer is connected into a bridge circuit, it is generally easiest to consider it to be composed of ideal sources with separate output impedances. Thompson has used this method to discuss two alternative equivalent circuits for an auto-transformer bridge network. These circuits are shown in Figs. 2(a) and 2(b) in which Z_X and Z_Y are the impedances being compared, Z_A and Z_B are the stray earth impedances, Z_{mn} is the mutual impedance between the windings, and $m/n = p$ is the turns ratio. In Fig. 2(a), z_m and z_n are the separate leakage impedances of the windings, and in Fig. 2(b) these are represented by the equivalent output impedance z'.

The balance equation for the circuit shown in Fig. 2(a) is:

$$Z_X = pZ_Y \left[1 + \frac{z_m - pz_n}{(1+p)Z_{mn}} + ((1/pZ_B)-(1/Z_A))\frac{z_m + z_n p^2}{1+p}\right]$$

This equation also holds at balance of the conjugate circuit shown in Fig. 2(c).

Balance for the alternative equivalent circuit shown in Fig. 2(b) is achieved when:

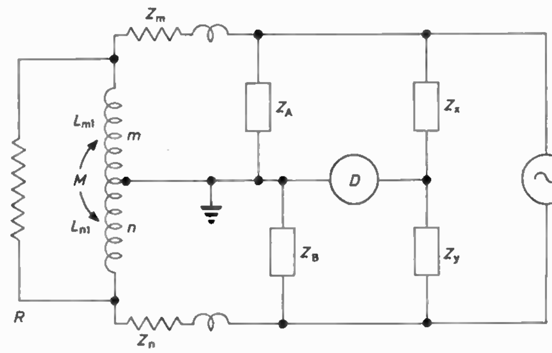

Fig. 3. General equivalent circuit for auto-transformer bridge

$$Z_X \simeq p'Z_Y \left[1 + ((1/p'Z_B) - (1/Z_A))(1+p')z'\right]$$

here p' is the open-circuit voltage ratio.

An alternative more general circuit is shown in Fig. 3. Here the nomenclature for the inductances is the same as that used earlier.

If R represents the core loss resistance, L the total input inductance and z_m and z_n are the leakage impedances, consisting of the winding resistances and leakage inductances then balance for this circuit is given by:

$$Z_X/Z_Y \simeq m/n\left[1 + ((1/R)+(1/j\omega L))((z_m/p_1)-(z_n/p_2)) + ((1/p_1Z_B) - (1/p_2Z_A))(z_m p_2^2 + z_n P_1^2)\right]$$

where

$$p_1 = \frac{m}{m+n} \text{ and } p_2 = \frac{n}{m+n}$$

It follows from this equation that the auto-transformer should have $z_m : z_n = m : n = Z_A : Z_B$ in order that its voltage ratio, when connected in a bridge network, should be equal to its turns ratio.

The first condition, i.e. $z_m : z_n = m : n$, is very close met with the type of construction discussed here, i.e. a winding of randomly-stranded conductors round a high permeability toroidal core.

THE THREE WINDING RATIO TRANSFORMER

This form of transformer can be used in bridge circuits in two dual ways shown respectively in Figs. 4(a) and 4(c).

In Fig. 4(a) the supply winding is separate from the ratio windings. No error in ratio can therefore ensue from potential drops in the leakage impedances of the ratio windings caused by magnetization and core-loss currents. A difference between the secondary terminal-voltage ratio and the turns ratio may occur, however, due to the fact that the secondary leakage impedances are directly in series with the impedances being compared. Furthermore, earth admittances across the ratio windings may also significantly contribute to a ratio error, for they, too, carry currents which flow in the leakage impedances.

In Fig. 4(c) the detector winding is on the same core as the ratio windings. Since there is no flux in the core

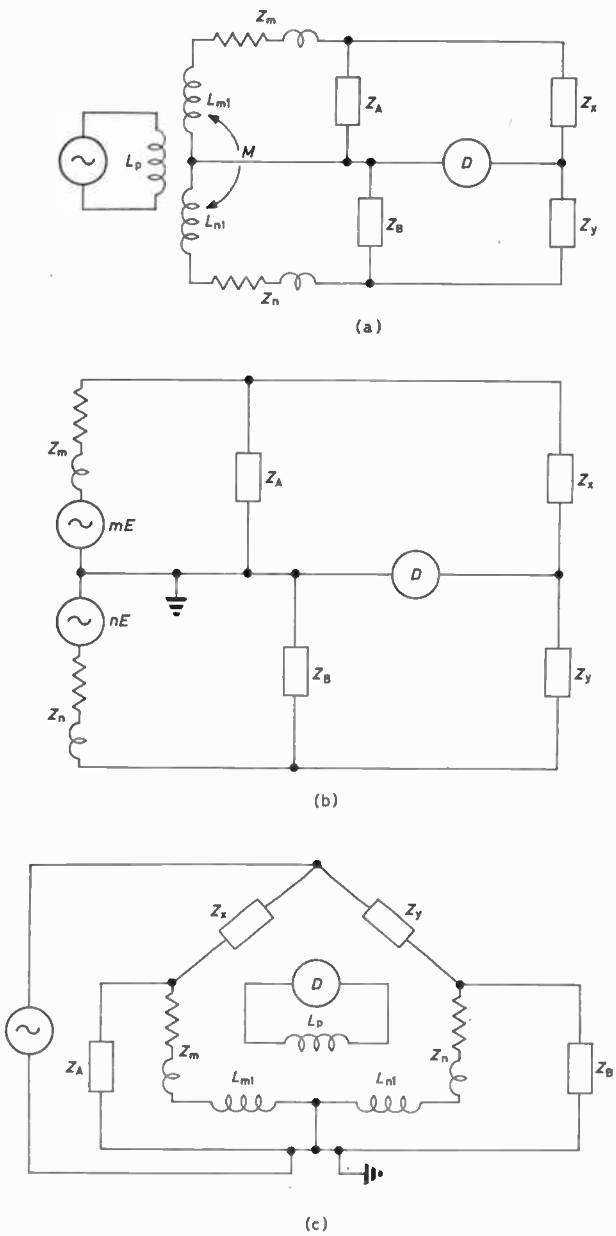

Fig. 4. Three-winding transformer bridge circuits

at balance there can be no core-loss current in the ratio windings and so no potential drops in the leakage impedances. In addition, there will be virtually no voltage across the ratio windings, and so any stray earth impedances across these windings will have little effect on the ratio. These advantages have been exploited in the current comparator[15,16] for precision current transformer testing.

A difference between the current ratio in the ratio windings and the turns ratio may occur, due to the fact that the leakage impedances of the ratio windings are directly in series with impedances being compared; the earth impedances across the ratio windings may also contribute to the ratio error. Direct analysis of either circuit, Fig. 4(a) or Fig. 4(b), gives:

$$\frac{Z_X(1 + (z_m/Z_A)) + z_m}{Z_Y(1 + (z_n/Z_B)) + z_n} = m/n$$

for the balance condition. Thompson[4] has shown that the simple circuit shown in Fig. 4(b), in which the secondary windings are considered to be two voltage sources, mE and nE (i.e. having an open-circuit ratio m/n) with their respective output impedances z_m and z_n, is equivalent to circuits Fig. 4(a) and 4(c). The balance equation is precisely the same as that given above. The advantage of this equivalent circuit is that its parameters are easily measured.

For the balance equation to reduce to $Z_X/Z_Y = m/n$, the conditions $Z_A/Z_B = z_m/z_n = m/n$ must be satisfied. The condition $z_m/z_n = m/n$ can be fulfilled by making the ratio windings from a cable of randomly-stranded conductors, as described earlier, but it is impossible always to ensure that $Z_A/Z_B = m/n$.

Measurement of Parameters

Of the equivalent circuits discussed, those shown in Figs. 2(b) and 4(b) are probably the most useful in practice because, in order to use their balance equations, it is only necessary to measure the voltage ratio and either the output impedance z' (Fig. 2(b)) or the leakage impedances z_m and z_n (Fig. 4(b)). In general, however, more complete information about the transformer parameters is required. In this section, practical methods for the measurement of open-circuit ratio, (m/n), the leakage impedances, (z_m and z_n) and the stray impedances, (Z_A and Z_B) are given.

OPEN-CIRCUIT RATIO

A number of recent papers[5,8-11] have been devoted to the problem of measuring, with high precision, the open-circuit ratio, m/n, of ratio transformers. In general, it is necessary to compare the unknown ratio with a reference divider of the same nominal ratio. The inevitable inaccuracies in the reference divider are eliminated by making a series of measurements with permuted connexions to nominally equal sections of the reference divider. It is assumed that errors in each section of the reference divider are either constant or all change identically during the sequence of measurements.

Most of this work has been applied to ratio autotransformers. However, the methods used may also be applied to three-winding voltage transformers, provided the reference divider does not cause a significant loading error on the ratio windings of the test transformer.

Cutkosky and Shields[9] have reported the most precise measurement of m/n which has probably so far been made. They used eleven nominally equal three-terminal capacitors in a reference divider and measured three-winding transformer ratios of 10:1 to a few parts in 10^9. This very high precision was only achieved by extremely elaborate and careful measurement techniques, and by evaluation of corrections for load impedances and earth admittances of the reference divider.

A simpler, though not so accurate, method of evaluating the ratio of a three-winding voltage transformer is based on a technique reported by Pinckney[11]. In this method, a transformer of several sections has an inductive ratio divider connected across its first two sections and so the error in ratio between these two sections can be evaluated when this circuit is balanced. Balance can be achieved with great accuracy, since inductive ratio dividers have now been constructed which can discriminate to 1 part per 10^7 with an accuracy of 1 part per 10^8 over most of their range. The loading error caused by this divider will be insignificant, since its input impedance usually equals a resistance of 500kΩ in parallel with an inductance of 150H, whereas the output impedance of each section will be about equal to a resistance of 5mΩ in series with an inductance of 1μH, and so the error caused by the inductive ratio divider will never be as much as 1 part per 10^7. After this first measurement, the second and third sections are compared, and so on until all adjacent sections in the ratio have been compared. The open-circuit voltage ratio between any two adjacent sections in the transformer being thus evaluated, it is quite easy to evaluate the ratio between any two combinations of sections in the transformer.

Perhaps the simplest of these methods to apply in practice and one which is also very precise is that described by Hill and Miller[5].

A transformer having ten sections was compared against a very stable, 900:100Ω, oil-immersed, non-inductive, resistance ratio. Balance was achieved by adjusting a resistor S and the capacitor C connected in parallel with the 100Ω resistor. The sections of the transformer winding were then cyclically permuted so that each section was adjacent, in turn, to the 100Ω resistor.

At each permutation the bridge was balanced, and the values of S noted. Taking $a_1, a_2 \ldots a_{10}$ to be the errors in each section, $S_1, S_2 \ldots S_{10}$ the corresponding balance settings of S, and k_r the error in the 100Ω resistor.

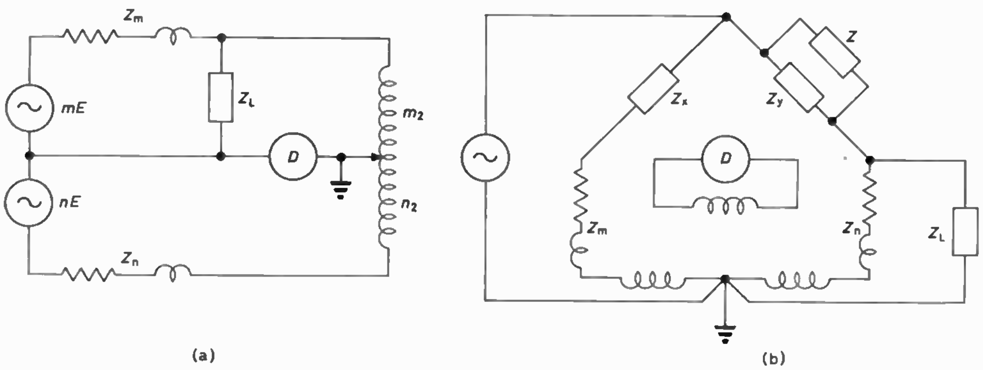

Fig. 5. Measurement of leakage impedances

Then:

$$a_1 = 9/10\,(k_r - x_1) \text{ where } x_1 = \frac{100\,(1 + k_r)}{S_1}$$

and similarly for a_2, a_3 etc.

a_1 is initially assumed to be zero, and the relative values of $a_2, a_3 \ldots a_{10}$ determined. By taking the average of these values of a from the values of a themselves, the error in each section was obtained.

Leakage Impedances

The basic procedure to determine the leakage impedance of any ratio winding is to measure the change of transformer ratio produced when the winding is shunted by a known impedance. This procedure may be illustrated for the case of a three-winding transformer by considering the circuit shown in Fig. 5(a) in which the test transformer is directly compared against a multi-decade inductive divider. z_m and z_n are the leakage impedances to be measured. The input impedance, Z, of the inductive divider will be very high, e.g. megohms compared to the leakage impedance, e.g. 0·05Ω. The bridge is initially balanced without Z_L connected. The ratio obtained on the divider is say m_1/n_1 so that Z is effectively divided by the variable tapping into the components m_1Z and n_1Z. Thus the balance equation is:

$$m/n = \frac{m_1\{1+(z_m/m_1Z)\}}{n_1\{1+(z_n/n_1Z)\}} \simeq (m_1/n_1)\,\{1+(1/Z)\,((z_m/m_1)-(z_n/n_1))\}$$

since z_m/Z and $z_n/Z \ll 1$.

To achieve a precise balance in practice, a trimmer capacitor of a few tens of picofarads connected across one of the sides of the divider will generally be necessary. Such a small quadrature-balance component, however, has a very small effect on the divider ratio.

If the impedance Z_L is now connected across one of the ratio windings of the test transformer, and the bridge re-balanced to yield a divider ratio m_2/n_2, the balance equation is:

$$m/n = m_2/n_2\,\frac{\{1+(z_m/m_2Z)+(z_m/Z_L)\}}{\{1+(z_n/n_2Z)\}} \simeq$$
$$m_2/n_2\,\{1+(1/Z)\,((z_m/m_2)-(z_m/n_2))+(z_m/Z_L)\}$$

since z_m/Z and $z_n/Z \ll 1$.

Equating these two expressions for m/n and using the fact that the change in divider ratio will only be of the order of a few parts in 10^6, i.e. $(m_1/n_1) \simeq (m_2/n_2)$, then:

$$z_m = Z_L\,\frac{(m_1/n_1)-(m_2/n_2)}{(m_2/n_2)}$$

z_m comprises a small resistance r_m in series with a small inductance l_m so that when Z_L is made resistive, R say, there will be a small quadrature unbalance caused by l_m. This has to be balanced out by a trimmer capacitor connected across the inductive decade divider. If the divider ratio at balance for this case is (m_2'/n_2'), then:

$$r_m/R = \frac{(m_1/n_1)-(m_2'/n_2')}{(m_2'/n_2')}$$

If, however, Z_L is made capacitive, giving a value C at the balance setting (m_2'/n_2') of the divider, then:

$$\omega^2 l_m C = \frac{(m_1/n_1)-(m_2''/n_2'')}{(m_2''/n_2'')}$$

Thus, as these two equations show, it is possible to separate the resistive and inductive components of the leakage impedances.

Fig. 5(b) shows the dual circuit in which the reference divider comprises Z_x and Z_y. The value of Z_n is found by loading the winding with the impedance Z_L as shown and noting the change of ratio by adjustment of Z.

Stray Earth Impedances

Rayner[3] has furnished information on how to evaluate the stray impedances Z_A and Z_B. In this method the detector has one of its inputs disconnected from the impedances, Z_X and Z_Y, and it is then connected to the earth point as shown in Fig. 6. This circuit is then balanced by applying the impedance Z_{C1}, thus giving

$$p_1 = \frac{Z_A}{Z_B Z_{C1}/(Z_B+Z_{C1})} = \frac{(1/Z_B)+(1/Z_{C1})}{(1/Z_A)}$$

If another ratio point, say p_2, is chosen on the auto-

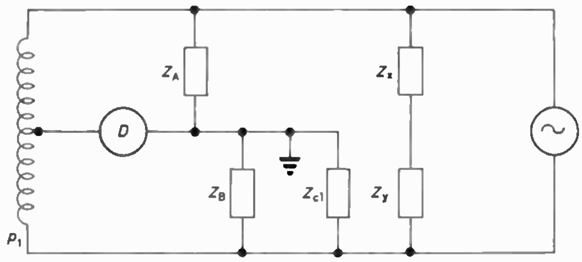

Fig. 6. Measurement of stray impedances

transformer and balance is obtained by adjusting Z_{C1} to a new value Z_{C2}, then

$$p_2 = \frac{Z_A}{Z_B Z_{C2}/(Z_B + Z_{C2})} = \frac{(1/Z_B) + (1/Z_{C2})}{(1/Z_A)}$$

Hence:

$$Z_A = \frac{p_2 - p_1}{(1/Z_{C2}) - (1/Z_{C1})} \text{ and } Z_B = \frac{p_1 - p_2}{p_2/Z_{C1} - p_1/Z_{C2}}$$

Although this analysis has been worked out for an auto-transformer the method applies equally well to the three-winding transformer.

More complete information on the measurement of ratio-transformer parameters is given in an N.E.L. Report[17].

Conclusion

Before an assessment can be made of the possible measurement accuracy of a particular ratio-transformer bridge circuit under specified working conditions of voltage, frequency, etc., it is necessary to know the appropriate values of the stray and leakage parameters for that circuit. In this article the theory of some methods of measuring these parameters is described. In general the practical measurement techniques are simple. The bridge circuits chosen for analysis are those which are of practical use.

Acknowledgment

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Self-Tuning Communication Receivers

The Marconi Co. Ltd has announced a completely new range of radio receivers which will cut both capital and running costs for high-grade h.f. communications. Frequencies are selected on simple decade dials, and the receiver then performs all tuning operations entirely automatically, enabling one man, at a central point, to control every receiver in a major communications centre.

The Marconi MST h.f. communications system has been developed to take every advantage of new techniques and methods of construction. It provides smaller, more reliable and more stable equipment, automatic in operation as far as possible, and very rapid frequency changing. All units are fully transistorized, and two major sources of unreliability, the valve and the telegraph relay, have been completely eliminated. Printed circuit techniques and modular construction are used extensively throughout the equipment, and the overall size has been reduced to a third of that of previous designs.

To operate an MST receiver, the operator simply sets decade dials on a frequency synthesizer unit to the required frequency, and presses a button. This initiates the self-tuning action which, within 24sec, accurately tunes all stages of the receiver to the new frequency. Any one of 250 000 frequencies (in steps of 100c/s) can be selected, and when automatic frequency control is used, the capture range is sufficient to give complete coverage of all frequencies by setting the synthesizer to the nearest 100c/s step. The synthesizer units are mounted separately from the receiver in standard racks, a large number of which can be accommodated at the station control position. The receivers, installed in a different part of the station, are less than one quarter of the size of previous designs, up to three being mounted in a single rack.

The MST receiving equipment range comprises three separate receivers, and a range of associated units to make up a complete station of any given size. The associated equipment includes the frequency synthesizer, the display and control unit, and a master frequency source.

The three receivers are a dual-diversity telegraph receiver, type H2002, suitable for diversity reception of f.s.k. or four frequency diplex signals; a dual diversity i.s.b./s.s.b. receiver, type H2102, and a single path i.s.b./s.s.b. receiver, type H2112.

The frequency synthesizer, type H1500E, covers the band 3·0000 to 30·9999Mc/s, and provides the high-stability first oscillator input to any of the MST receivers and a 100kc/s signal used for carrier re-insertion and for generating the second oscillator signal where a.f.c. is not used.

The required frequency may be selected in a matter of seconds by adjusting a series of decade controls indicating the receiver frequency.

The front panel of the display and control unit carries the controls and indicators needed for extended operation of a receiver. Labels are provided for recording the identity of the controlled receiver, the service or signal being received, the aerials in use and the line, v.f.t. or other channel connecting the receiver output to the central terminal station.

The controls enable the operator to initate the self-tuning process, adjust the amount of attenuation at the input to the receiver, select one of three available aerials, switch the a.f.c. system, and select a path for monitoring.

Signal lamps indicate the position of the aerial input attenuator, supply failures, receiver drift to the limits of effective a.f.c. control and fading below threshold of the received signal.

The synthesizer requires a 1Mc/s master frequency signal of a high accuracy and stability. This can be provided by the master frequency source type H1605 which is used with the MST transmitting system. Other master frequency sources with even higher standards of frequency stability are available.

A Transistorized Resistance-Capacitance Selective Network

By T. H. Appleby*, A.M.I.E.R.E.

It is now generally accepted that microelectronics will eventually improve the reliability and decrease the size of electronic equipments. The solid circuit and thin film techniques do not allow for the use of inductors for frequencies below 30Mc/s so that transistorized resistance-capacitance networks simulating LC type responses must be found. The circuit described here was the result of an investigation of such circuits.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 805)

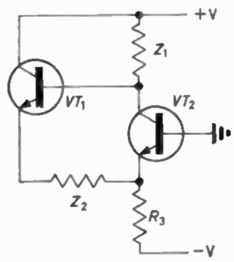
THE bulk of electronic equipments operate at frequencies below 30Mc/s so that the simulation of *LC* type responses in this range is of great importance. Various resistance-capacitance selective networks already exist and when used with transistors may be used to make oscillators or selective amplifiers. Initially the standard circuits capable of operation in this band were investigated, using where possible, thin film exponentially tapered distributed resistance capacitance networks to improve the performance. The standard techniques were found to be unsuitable because of poor stability and the use of distributed resistance capacitance networks did not improve the stability enough to make the circuits very useful.

Resistance-Capacitance Frequency Selective Circuits

Frequency selective circuits using capacitance, resistance and active elements may be divided into three groups, according to the manner in which the desired effect is achieved. The first group consists of those circuits in which negative feedback is applied via a frequency selective network. The most commonly used network in this application is the twin-T and since a distributed resistance capitance, tapered exponentially to the power of five may be used to make a network with a sharper null than the twin-T, it was hoped to produce a highly selective circuit. It was found that due to oscillation, the open loop gain was limited so giving the network a poor skirt rejection. Also the response was peaky as opposed to the required gaussian shape of an *LC* circuit. When the null network was arranged so that only a partial null occurred, a stable amplifier with a Q_E in the region of five resulted. Using the network with maximum rejection resulted in a maximum Q_E in the region of fifteen for reasonable stability, although amplifiers with a Q_E greater than a hundred were operated for several hours under laboratory conditions. The type of circuit in this first group will then probably be useful for narrow band application where frequent attention to the equipment is possible, or where a very low Q_E is admissible.

The second group comprises those in which phase-shift through an *RC* network is used to obtain a total loop phase-shift of 360°. In this group it is usual to obtain 180° of phase-shift through the *RC* network and 180° by using a grounded emitter transistor. The reason for this is that an *RC* network used to give 360° of phase-shift has a too large attenuation. The parameter controlling the Q_E, thus stability in this application is the rate of change of phase-shift at the frequency at which 180° phase change occurs in the *RC* network. Exponentially tapered distributed *RC* networks are considerably better than lumped networks in this respect and were investigated. However for various reasons although there is an improvement over lumped *RC* circuits in this group, the improvement does not make the circuits very much more useful. The main reason for this is that a grounded emitter transistor is used in the loop which tends to result in poor stability.

The third group comprises those in which active elements are used to transform the impedance across one pair of terminals into the negative of the impedance across another pair of terminals. Various workers have produced some circuits but those investigated did not seem very stable and are not in general use.



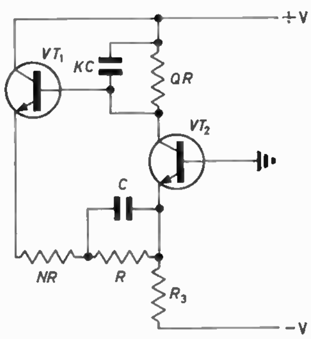


Fig. 1. Negative impedance convertor circuit
Fig. 2. Negative impedance convertor oscillator

Negative Impedance Oscillator and Selective Amplifier

Negative impedance convertors such as that shown in Fig. 1 are well known. Referring to Fig. 1; if the voltage at the collector of VT_2 increases, then the voltage at the emitter of VT_1 will follow it by by the same amount. The resulting increase of current through the impedance Z_2 may only be achieved by a decrease of the current through VT_2 if the current through R_3 is constant. The net effect is that an increase of voltage at VT_2 collector causes a decrease of current through VT_2 which is inversely proportional to Z_2, so giving an impedance of $-Z_2$ at the collector of VT_2. Referring to Fig. 2; the impedance of *R*, *C* and *NR* is, at any frequency, equivalent to a capacitor in series with a resistor. The impedance seen looking into the collector of VT_2 is the negative of this and equivalent to an inductor in series with a negative resistance, at any frequency. When the series connexion equivalent of *QR* and *KC* is added, the total impedance at VT_2 collector is a lossy capacitor, in parallel with an inductor in series with a negative resistance. If it is arranged that at a certain frequency the inductive and capacitive impedance are equal, then the circuit will have frequency selective properties. Oscillation will occur

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at this frequency of resonance if the negative resistance is bigger than the positive resistance. If the positive resistance is slightly larger than the negative resistance it will act like a normal tuned circuit with the Q_E determined by the net positive resistance. The following will show that by including the resistor NR it is possible to make a stable oscillator. Also it will be shown that NR determines the Q_E of a selective amplifier and that as the resistor is not in the frequency determining equation, a variable Q_E, constant frequency amplifier may be made.

Circuit Analysis

Appendix 1 shows the derivation of the oscillation frequency equation.

Appendix 3 shows the derivation of S.

The fundamental frequency of the oscillator is given by C and R as $f_o = 1/2\pi CR$ and if $x = f/f_o$ the frequency of zero reactive impedance (f_{osc}) is given by:

$$(f_{osc}/f_o)^2 = x^2 = (1/Q^2K)\left(\frac{S - Q^2K}{1 - SK}\right) \quad \ldots\ldots\ldots (1)$$

$$S = \frac{1}{1 - \alpha_1 - \alpha_2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

$\alpha_1 \alpha_2$ are common base current gains of VT_1 and VT_2. The circuit will oscillate at the frequency given by equation (1) if the real part is negative as explained above and it may be shown the real part is.

$$\mathrm{Re} = R \, . \, \frac{(Q - S - SN(1 - x^2QK))(1 - x^2QK) + (x^2)(1 + QK)(Q(1 - SK) - SN(1 + QK))}{(1 - x^2QK)^2 + x^2(1 + QK)^2} \quad \ldots (3)$$

Stability of Damping

The resistance NR in equation (3) is not merely the resistor NR in Fig. 2 but has the sum of the emitter resistances of VT_1 and VT_2 added to it. Since the stability of the oscillator depends on this fact, the operation will now be examined more closely. By setting $S = Q$ in equation (1) the real part is much simplified and it is possible to show that oscillation will occur if K is less than

$$(1/S)\frac{1 - N}{1 + N} \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

Now when the circuit is oscillating, the amplitude will be determined by the amplitude that has to be reached in order to make N sufficiently large to give stabilization as in equation (4). Taking the case when NR of Fig. 2 is zero, the effective value of N in equation (4) is given by the sum of the emitter resistors. Since the sum of the emitter currents of VT_1 and VT_2 is constant, the sum of the emitter resistors will increase for positive and negative swings of the oscillatory waveform. Also if the quiescent currents of VT_1 and VT_2 are equal, the value of N will obviously increase at the same rate for positive and negative swings. This circuit configuration approximates to the case where there is a range of easy oscillation bounded equally on both sides by a sharp onset of damping, which results in an output sensitive to power supply changes. However, by adding the resistor NR the rate of change of effective NR with amplitude is more gradual, so giving equal and gradual damping for positive and negative swings of the oscillatory waveform.

Effect of Power Supply Variation on Frequency

It is obvious that a constant current through R_3 is required so that V_2 and R_3 should be as large as possible. Also by differentiating the equations giving the collector VT_2 voltage, with respect to V_1 it may be shown that QR should be as small as possible compared with $R(1 + N)$, to minimize the change of transistor currents with power supply voltage. The minimum value of QR is determined by the transistors since KC is in parallel with the combined collector base capacitances of VT_1 and VT_2. If QR is reduced, KC must be reduced to allow oscillation, and the limit is reached when the collector base capacitance is considered to be too large a fraction of KC. Measurements clearly show the high degree of stability when equal emitter currents are used, with QR as small as possible, the measurements given in Tables 1 and 2 are on a 295kc/s oscillator. It may be seen that a ± 5 per cent change of power supply gives a $\pm 0{\cdot}05$ per cent change of frequency and this isolation ratio of 100:1 is much better than transistor phase shift oscillators investigated.

Effect of Temperature on Frequency

Since the emitters of VT_1 and VT_2 are connected to each end of the $(NR + R)$ combination, change of base emitter voltage will have no effect on the current through it. However a change of base emitter voltage will change the current I_3, although this may be minimized by using a high R_3 and V_2. The main cause of frequency change is due to the variation of α_1 and α_2 with temperature. Silicon transistors have a gain-temperature coefficient of about $+7000$ in 10^6 so giving S a negative temperature coefficient of -130 in 10^6 for $\alpha_1 = \alpha_2 = 0{\cdot}99$. The resulting change in frequency due to this will be of the order of -50 in 10^6 which is a considerable improvement over phase shift oscillators.

Temperature coefficient compensation may be achieved over a small range of temperature by using a positive temperature coefficient 'solid circuit' type of R_3. The effective decrease of the negative power supply voltage tends to increase the frequency as shown in Table 2. A 97kc/s oscillator was compensated in this way and the frequency changed by only 2c/s over a range of 10°C, oscillation being extinguished over a range of 20°C.

Example of Oscillator Design

The circuit constants were $R = 3400\Omega$, $QR = 1600\Omega$, $C = 820$pF, $KC = 165$pF, $\alpha_1 = \alpha_2 = 0{\cdot}083$ and $S = 1{\cdot}017$. The fundamental frequency was found from C and R to be 56·6kc/s. In practice the Q_E of the thin film capacitors at 300kc/s is in the region of 100 and the loss resistance associated with them slightly reduces R and to a lesser extent QR. Taking the Q_E into account the frequency of

TABLE 1

NEGATIVE POWER SUPPLY=20V I_3=2mA				
Positive Power Supply (V)	5	5·5	6·0	6·6
Transistor VT_1 Current (mA)	1·4	1·2	1·1	0·63
FREQUENCY CHANGE (c/s)	−420	−200	0	+10
OSCILLATION RANGE (V)	4·4 to 7·4			

TABLE 2

POSITIVE POWER SUPPLY=6V				
Negative Power Supply (V)	18	19	21	22
Transistor VT_1 Current (mA)	0·8	0·9	1·3	1·45
FREQUENCY CHANGE (c/s)	+50	+70	−70	−400
OSCILLATION RANGE (V)	17·5 to 30			

oscillation was calculated as 297kc/s compared with the measured 295kc/s. A simple calculation, ignoring the capacitor Q_E gave a frequency of 305kc/s.

Variable Frequency Oscillator

The frequency may be varied by changing the capacitors C and KC at the same rate to give a constant K. Since the frequency is inversely proportional to the capacitance a 10:1 frequency change is possible with a normal tuning capacitor, compared with 3:1 for an LC oscillator.

Selective Amplifier

Appendix (2) shows the equation giving the Q_E of the circuit and the variation of Q_E with N, K, Q and S is shown in Fig. 5.

The design procedure for a selective amplifier is almost

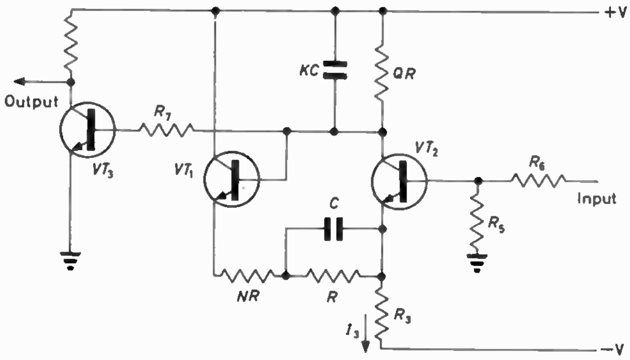

Fig. 3. Negative impedance tuned amplifier

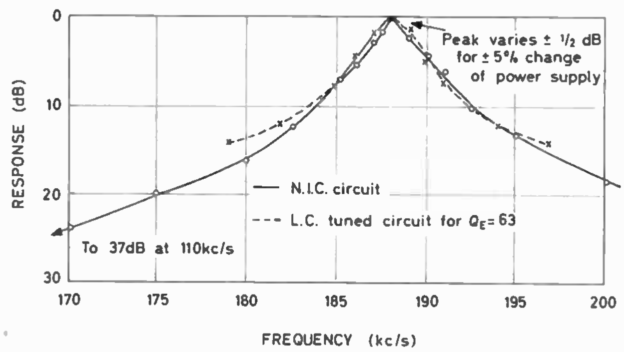

Fig. 4. Frequency response of N.I.C. amplifier

identical to that for the oscillator. The only differences being that NR is adjusted to make the circuit non-oscillatory and the emitter currents of VT_1 and VT_2 are increased so as to reduce the effect of emitter resistance on gain and Q_E. A convenient method of coupling into and out of the circuit is shown in Fig. 3. Very little coupling in the reverse direction is introduced so that the stages may be cascaded without oscillation. Selective amplifiers using RC feedback are not in general use, because as a class they suffer from instability due to variation in power supplies and temperature. Also the characteristics of the amplifiers do not generally repeat the useful frequency characteristics of the LC tuned circuit. However, as is shown in Fig. 4, the amplifier described here is relatively stable with respect to power supply change, a change of ±5 per cent in supply voltage changing the response by only ±0·5dB. The shape of the response is also very nearly the same as that for an LC circuit, giving a much better skirt attenuation than other circuits investigated during the work.

Conclusions

The circuit described here is considered to be a useful

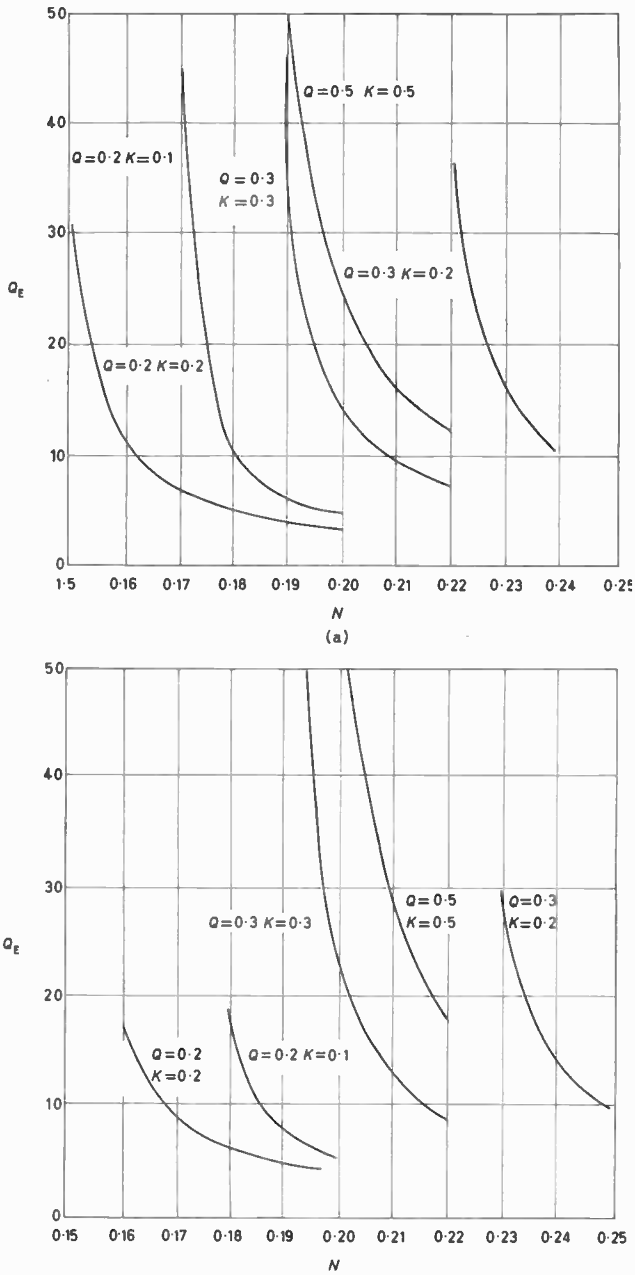

Fig. 5. Calculated variation of N.I.C. selective amplifier Q_E
(a) S = 1·05 (b) S = 1·02

selective network particularly when compared with other circuits that were investigated. When testing the circuit, the stability as compared with the other circuits was most striking, repeatable results being obtained over several weeks with an equivalent Q_E of 60. The other advantage of the circuit is that simple calculations allowed the easy design of circuits with an accurate result. This was a pleasant change from the arduous work involved when using tapered distributed RC networks.

Coupling of the circuits should allow the production of stable band pass amplifiers since the required type of response exists. Also the information needed to calculate the value of coupling impedance may be calculated from equation (5) of appendix (1).

Acknowledgments

The permission by the Directors of the Plessey Co. (U.K.) Ltd, to publish this article is gratefully acknowledged.

Appendix

(1) Derivation of N.I.C. Oscillation Conditions

Referring to Fig. 2 the total impedance (Z_t) seen by a generator inserted between the load and collector of VT_2 is as in equation (5)

$$Z_t/R = -S\left(N + \frac{1-jx}{1+x^2}\right) + Q\cdot\left(\frac{1-jK^2Q^2x^2}{1+K^2Q^2x^2}\right)\ldots\quad (5)$$

where x is the normalized frequency WCR.

Equating the imaginary part to zero:

$$Sx/1+x^2 - \frac{KQ^2x}{1+K^2Q^2x^2} = 0 = I.P.(Z_t/R)$$

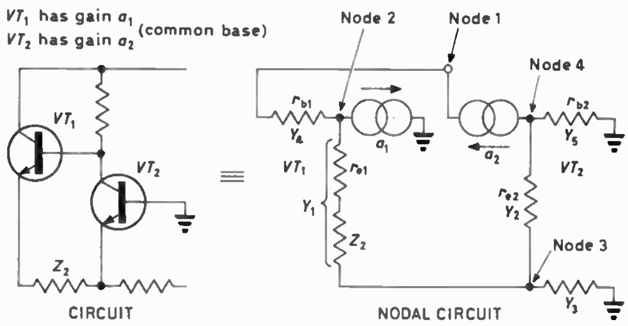

Fig. 6.

Solution of this for x gives:

$$x^2 = \frac{S-KQ^2}{KQ^2(1-SK)}$$

The real part is given by:

Re $(Z_t/R) =$

$$-S\left(N + \frac{1}{1+x^2}\right) + \frac{Q}{1+K^2Q^2x^2}$$

But when $S = Q$, $x^2 = 1/QK$ for zero reactive impedance. Thus the numerator becomes:

$$-N(1+SK)^2 - (1+SK)SK + (1+KS)$$

For this to be zero

$$K = \frac{1-N}{S(1+N)}$$

(2) Equivalent Q of N.I.C. Selective Amplifier

Since the response of the tuned amplifier is like that of an LC tuned amplifier it is reasonable to assume that the Q may be taken as the ratio of the positive or negative part of the reactive impedance to real part at resonance.

$$\text{Amplifier } Q_E = +\frac{\text{Im } Z/R}{\text{Re } Z/R}$$

$$= \frac{Sx}{(1+x^2) - S\left(N+\frac{1}{1+x^2}\right) + \frac{Q}{1+K^2Q^2x^2}}$$

At a frequency given by:

$$x^2 = \frac{(S-KQ^2)}{KQ^2(1-SK)}$$

Various values of S, N, K, Q, were inserted into this equation and the resulting Q_E plotted in Fig. 5. As seen in Fig. 5, no great advantage in low rate of change with Q_E with respect to N occurs for particular values of KQ. Since the d.c. stability is highest when QR is small, the optimum KQ for stability is in the region of 0·2 or less. The change of d.c. conditions owing to power supply and temperatures variation may easily be calculated and the Q_E stability predicted from Fig. 5.

N.B.—Q_E is the equivalent Q of the circuit taken as the ratio of midband frequency to the -3dB bandwidth. Q_E is also used in the text to represent the normal Q of a capacitor defined by $Q = \omega CR_P$ (R_P being the equivalent parallel resistance).

(3) Negative Impedance Convertor Circuit Analysis

The circuit is simplified by neglecting the collector to base capacitance and cut-off frequency of the transistor (Fig. 6).

Admittance Matrix Reduction

$+Y_4$	$-Y_4$	$-Y_2a_2$	$+Y_2a_2$
$-Y_4$	$Y_4+(1-a_1)Y_1$	$-(1-a_1)Y_1$	0
0	$-Y_1$	$Y_1+Y_2+Y_3$	$-Y_2$
0	0	$-(1-a_2)Y_2$	$Y_5+(1-a_2)Y_2$

Suppress Node 4

1	Y_4	$-Y_4$	$-Y_2a_2+\frac{Y_2^2a_2(1-a_2)}{Y_5+(1-a_2)Y_2}$
2	$-Y_4$	$Y_4+(1-a_1)Y_1$	$-(1-a_1)Y_1$
3	0	$-Y_1$	$(Y_1+Y_2+Y_3)-\frac{Y_2^2(1-a_2)}{Y_5+(1-a_2)Y_2}$

Suppress Node 3

1	Y_4	$-Y_4+Y_1\left[-Y_2a_2+\frac{Y_2^2a_2(1-a_2)}{Y_5+(1-a_2)Y_2}\right]\frac{1}{Y_1+Y_2+Y_3-\frac{Y_2^2(1-a_2)}{Y_5+(1-a_2)Y_2}}$
2	$-Y_4$	$Y_4+(1-a_1)Y_1-\frac{Y(1-a_1)}{Y_1+Y_2+Y_3-\frac{Y^2_2(1-a_2)}{Y_5+(1-a_2)Y_2}}$

From the 2×2 matrix the input admittance at node 1 is

$$Y_4 + \frac{Y_4\left(-Y_4+Y_1\left[-Y_2a_2+\frac{Y_2^2a_2(1-a_2)}{Y_5+(1-a_2)Y_2}\right]\frac{1}{Y_1+Y_2+Y_3-\frac{Y_2^2(1-a_2)}{Y_5+(1-a_2)Y_2}}\right)}{\left(Y_4+(1-a_1)Y_1-\frac{Y_1^2(1-a_1)}{Y_1+Y_2+Y_3-\frac{Y_2^2(1-a_2)}{Y_5+(1-a_2)Y_2}}\right)}$$

It may be shown that if Y_3 is much less than Y_2; Y_4 and Y_5 much greater than $1/Z_2 \times$ common emitter gains of transistors, the equation is much simplified and in practice.

Impedance at node 1 is given by:

Z input $=$

$$\frac{Z_1 + \text{emitter resistance } VT_1 + \text{emitter resistance } VT_2}{1-a_1-a_2}$$

A Novel Integrator and its Application to the Maintenance of Sensitivity in Null-Measurement Systems

By H. C. Bertoya*, A.M.I.E.R.E.

A novel integrator is described which is the electronic analogue of a servo motor. A particular application is then considered where the integrator is used to maintain the sensitivity of a servo-operated null-measurement system which compares the performance of an unknown (or 'test') element with a standard (or 'reference') element.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

THE basic circuit is shown in Fig. 1. Two grounded base npn and pnp transistors (simulating constant current sources) are connected across a capacitor. In the absence of leakage currents the voltage E_{AO} will remain constant if $i_1 = i_2$. If $i_1 \neq i_2$ then the charge on the capacitor—and hence E_{AO} will change. The sign of the change will depend on the magnitude of i_1 as compared with i_2. The circuit thus acts as a comparator as well as an integrator. The

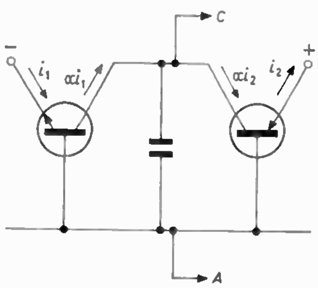

Fig. 1. Basic circuit of integrator and comparator

voltage E_{AC} may be monitored and used as a control in a manner analogous to a servo motor. E_{AC} simulates the position of the motor output shaft and $i_1 - i_2$ the current into the motor winding.

Since the transistors operate as constant current sources the rate of change of E_{AC} is linearly related to $i_1 - i_2$. That is:

$$Q = CE_{AC} = (i_1 - i_2)t$$

$$dE_{AC}/dt = (i_1 - i_2)/C$$

In practice there will be leakage across the capacitor and through the input resistance of the stage monitoring E_{AC}. The steady-state value of i_1 and i_2 should be large in comparison with this leakage. In the practical circuit of Fig. 2 the two transistor bases are taken to different potentials. The limit of voltage change is given by the minimum base-collector potential at which the transistors will operate—approximately the base potential. In the circuit these are about +0·7V and −10·6V. If an electrolytic capacitor is used it will thus be properly polarized over the working range.

In the circuit of Fig. 2 it will be seen that the current i_2 is set by a potentiometer RV_3. This sets a reference level. If it is assumed that the α of the two transistors is identical and leakage is negligible, then the voltage E_{AC} will remain constant when $i_1 = i_2$. Note that i_1 can either be d.c. or a.c. so long as in the latter case $i_{1(av)} = i_2$. In fact, in the sections which follow, it will be seen that a sinusoidal signal (the system carrier) is applied to the base of VT_6. VT_6 acts both as a buffer stage and, with VT_7, as a half-wave rectifier supplying $i_{1(av)}$. The voltage E_{AC} is monitored by the double emitter-follower stage VT_{10} and VT_{11}. The output may be taken directly from the emitter of VT_{11} for the purpose of control, although in the circuit of Fig. 2 adjustments are included whose purpose will be explained later.

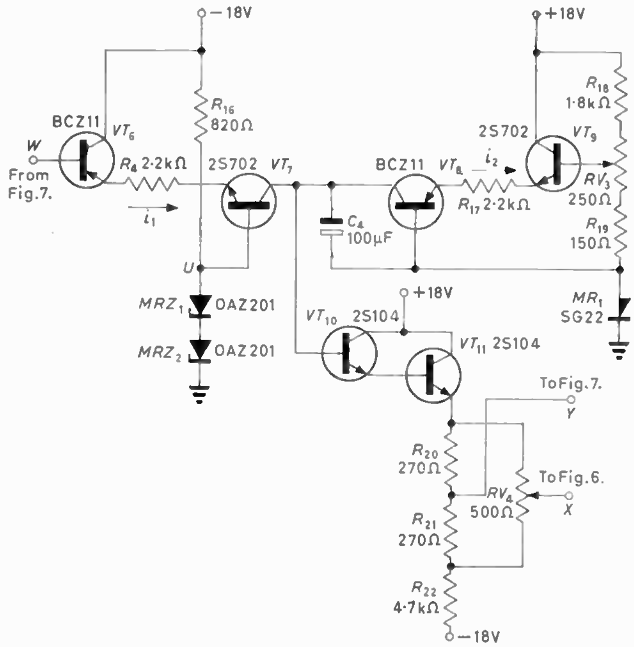

Fig. 2. Circuit of integrator and comparator

Application of Integrator and Comparator to the Maintenance of Sensitivity in a Servo-Operated Null-Measurement System

Though the problem is a general one, the particular apparatus for which the automatic gain control was required takes the form shown in Fig. 3. The apparatus measures the optical density of a medium.

In order to maintain a constant system sensitivity, i.e. to arrange that there will always be the same servo-drive available for a given percentage difference between the signals in the test and reference channels—the gain of the difference amplifier must be varied in an inverse manner to that of the absolute level of the signal in the test channel. It is not possible to use a conventional a.g.c. circuit[1] for the difference amplifier since, at null, the input to the amplifier is zero whatever the level of the signals in the inputs to the difference circuit.

The method of control adopted is shown in Fig. 4 and its operation is as follows.

The signal obtained from the measurement of absolute level in the test channel is attenuated in order to bring it within the range of operation of the gain-control amplifier.

* *British Scientific Instrument Research Association.*

The gain-control amplifier and difference amplifier have similar characteristics so that the gain-control changes the gain of both amplifiers equally. The output level of the gain-control amplifier is compared with a reference level. If the absolute level changes there is a difference output from the comparison circuit which drives the integrator and gain control. The gain-control amplifier is then adjusted until its output level is again equal to the reference level.

Since the absolute level of the signal in the test channel determines the system sensitivity, the change of gain of the gain-control amplifier is a measure of the change of

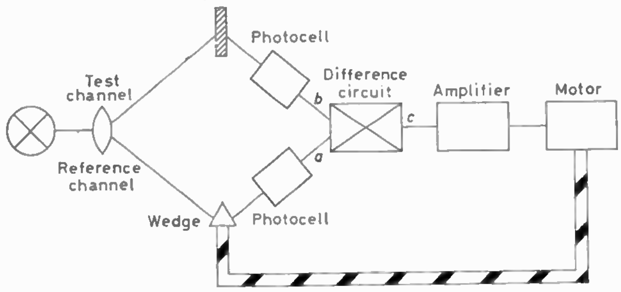

Fig. 3. Form of apparatus for measurement of optical density

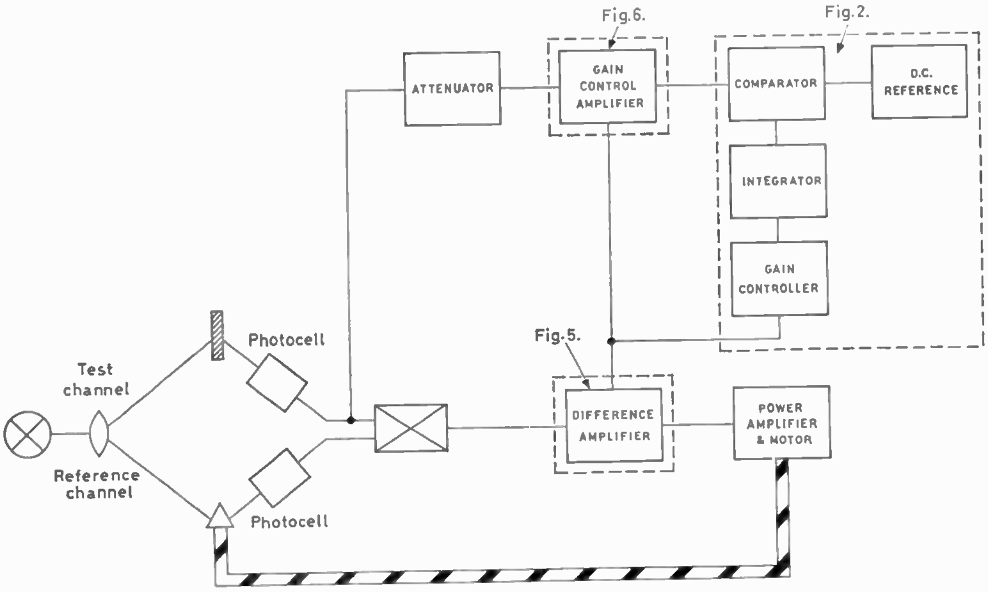

Fig. 4. Method of control and operation

gain required in the difference amplifier. The required change of gain in this latter is in fact obtained—since, as already stated, the gain control affects both the difference amplifier and gain-control amplifier equally.

Description of Circuits

The system operates with a 50c/s carrier and the circuits are shown in Figs. 2, 5 and 6.

Operation

The operation of the control loop (circuits of Figs. 2 and 6) is as follows.

The current i_2 is set by RV_3 which controls the base voltage of VT_9. This is the 'set reference level' control. The voltage across the capacitor C_4 will either increase or decrease and the voltage is transferred via the buffer stage (VT_{10}, VT_{11}) to the grid of V_1. The absolute level input signal passes through V_1 and the following amplifier The a.c. output signal at W is applied to VT_6 which acts both as an emitter-follower buffer stage and, with VT_7, as a rectifier. The average value of the half-wave of current through R_4 forms the current i_1. The magnitude of i_1 obviously depends on the amplitude of the a.c. signal at W.

The voltage across C_4 (and hence the voltage applied to the grid of V_1) will continue to change until the gain of V_1 has reached such a value that $i_{1(av)} = i_2$. Equilibrium is then achieved and the amplitude of the a.c. voltage at W remains constant at the value determined by the setting of RV_3.

The control voltage applied to V_1 in Fig. 6 (the control amplifier) is also applied to V_1 in Fig. 5 (the difference amplifier), thus altering its gain in proportion to the magnitude of the absolute level signal.

Variable-mu Pentode

This is a Mullard EF97 capable of working with the same order of anode voltage as is used for the transistor stages.

The a.c. signal at the anode must be kept small in order to prevent excessive distortion and is of the order of 20mV peak-to-peak. The maximum input level is also determined by the degree of distortion permissible.

The effect of differing characteristics between the two valves is minimized by carrying out two adjustments on the valve in the difference channel. The first adjustment, by means of RV_1 (Fig. 5) is to the screen voltage in order to obtain equal gain at high input signal levels (when the g_m is low). The second adjustment, by means of RV_4 (Fig. 2) is to the grid voltage in order to obtain equal gain at low input signal levels when the g_m is high).

The valve in the control amplifier requires no adjustment apart from the initial setting of the screen voltage to the specified level. This amplifier is, of course, inside the control loop. The effect of the time-constant $C_6 R_6$ may be neglected when considering the complete servo loop since $R_5 \gg R_6$. The output of the pentodes is taken through emitter-followers in order to present a low impedance to the following fixed amplifiers. The time-constant $C_8 R_1$ in the difference amplifier is larger than that of $C_1 R_1$ in the control amplifier. This is to reduce the value of the induced distortion as explained later.

The heaters were operated from a d.c. source in order to reduce hum.

Fixed Gain Amplifier in Difference Channel

While the two amplifiers in the difference and control channel are similar in design, there are some detail differences. The design employs both a.c. and d.c. feedback stabilization[2,3]. It can be shown that if an amplifier has the configuration shown in Fig. 7 then the d.c. output voltage is given by:

$$E_{AC} = E_{AD} R_{15}/R_9 h_{FE2}$$

where the h_{FE2} is the d.c. gain of VT_2.

If the source impedance is low, the gain of the amplifier will be given by:

$$V_o/V_i = R_{15}/R_2$$

To allow for differing values of h_{FE}, the required voltage E_{AC} is obtained by adjustment of the variable resistor RV_2. This alters the value of E_{AD} in the above equation. RV_2 and C_7 together also form decoupling components. The capacitor C_5 across the first collector load is inserted as accurately as possible. This was achieved as follows:

(1) The type of amplifier employed has inherently the minimum number of time-constants since—apart from the input capacitor—it is d.c. coupled throughout.

(2) The amplifier is fed from a low output-impedance emitter-follower so that the input time-constant is defined by

$$R_2C_2 = \tau_2$$

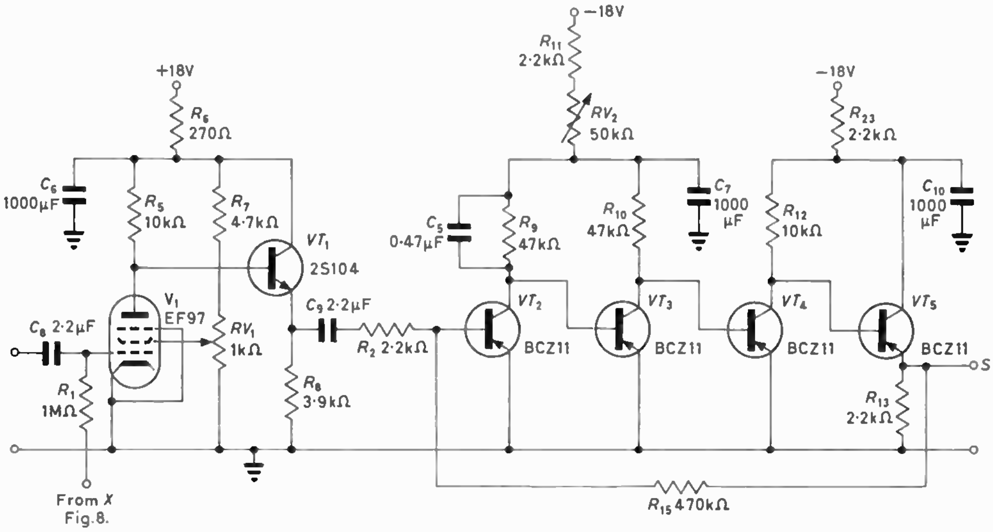

Fig. 5 (left). The difference amplifier

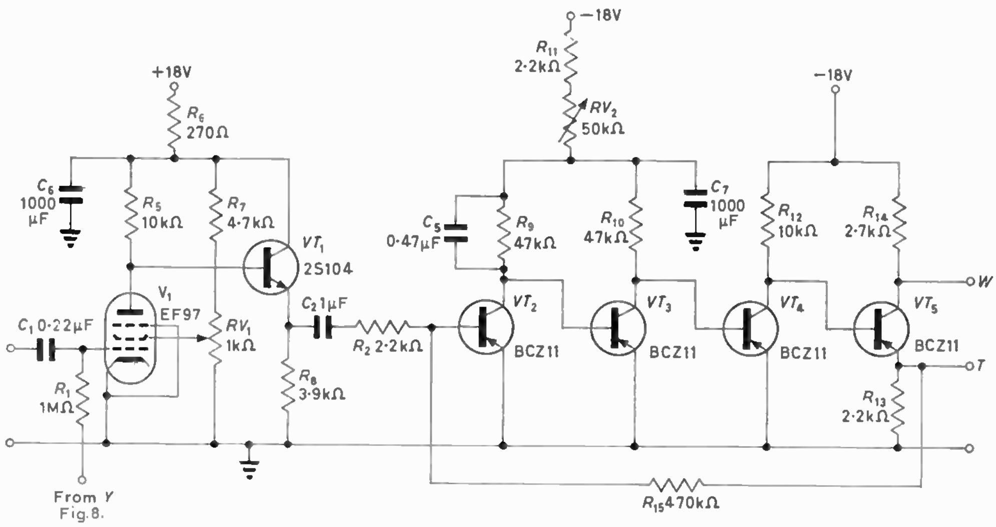

Fig. 6 (right). Gain control amplifier

for the purpose of amplifier loop stabilization. The various other time-constants given by coupling and decoupling capacitors etc. are not critical and have been chosen to suit the operating frequency of 50c/s. The components C_9R_2 (as well as C_8R_1 in the previous valve stage) make the response time of the difference amplifier longer than that of the control loop.

Fixed-gain Amplifier in Control Channel

This amplifier is similar to the amplifier in the difference channel but since it forms part of a complete servo loop the various time-constants must be taken into account in order to determine conditions for stable operation. The first aim was to reduce the number of time-constants as far as possible and then to determine those remaining

(3) Since the first two stages (VT_2 and VT_3) operate into low impedances (the input resistance of grounded emitter transistors) the effect of the time-constant given by the decoupling components RV_2 and C_7 can be neglected.

(4) The final stages VT_4, VT_5 are not decoupled.

The output of the control amplifier is taken from the collector of VT_5. This is done in order to obtain the correct phase relationship around the complete control loop and to provide the correct d.c. level for operation of the following unit.

The gain of both amplifiers is 200 and the bandwidth (determined primarily by the value of C_5, and the open loop gain) is about 8kc/s.

INTEGRATOR AND COMPARATOR

This unit has already been described but some additional points should be noted.

The diode MR_1 is inserted so that the maximum positive voltage at Y will drive the grid of the variable-mu pentode as near to ground as possible without entering the grid current region. If this happens loss of control will result. No attempt was made to match VT_7 and VT_8.

Conditions for Loop Stability

The complete loop consisting of the gain control amplifier and integrator/comparator may be represented by the

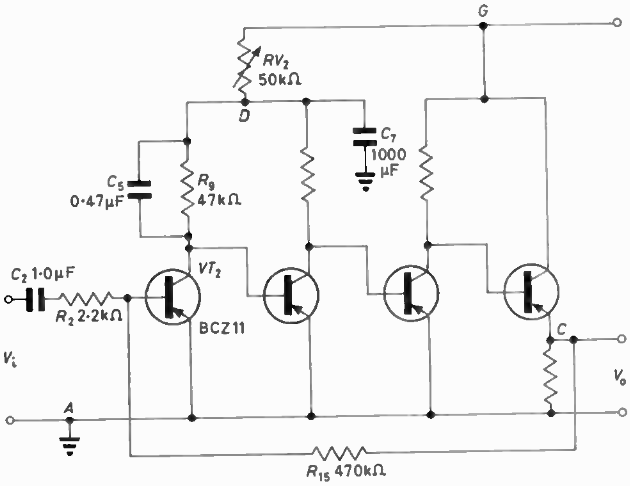

Fig. 7. Basic circuit of amplifier with a.c. and d.c. feedback stabilization

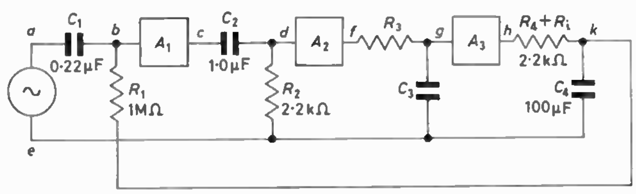

Fig. 8. The control loop showing time-constants

equivalent circuit of Fig. 8. A_1 represents the variable-mu valve stage and A_2 the transistor amplifier. A_3 represents the buffer stage VT_6 which has unity gain. The time-constants (referred to the circuit diagram of Figs. 2 and 6) are represented in the equivalent circuit as follows:

$$C_1R_1 = \tau_1, \quad C_2R_2 = \tau_2$$

τ_3 is the effective time-constant of the amplifier given approximately by $\tau_3 = 1/\omega_c$, where ω_c is the point where the response is 3dB down. τ_4 and τ_5 are determined from the circuit of Fig. 9. Consider the circuit of Fig. 9(b). The transfer function of this latter written in operational form is given by:

$$E_{ek} = E_{eh}\,\frac{1/C_4(R_i + R_4)}{p + 1/C_4(R_i + R_4)}$$

where p is the Laplace variable.

$$\text{put } C_4(R_i + R_4) = \tau_4 \text{ then } E_{ek} = E_{eh}\,\frac{1/\tau_4}{p + 1/\tau_4} \quad \ldots\ldots (1)$$

The equations for Fig. 9(a) are:

$$E_{eh} = i_3(R_4 + R_i), \quad E_{ek} = i_5/pC_4$$

$$i_5 = i_4 R_o/(R_o + 1/pC_4), \quad i_4 = \alpha i_3 \simeq i_3$$

Therefore:

$$E_{ek} = 1/pC_4 \cdot \frac{R_o}{R_o + 1/pC_4} \cdot \frac{E_{eh}}{R_4 + R_i}$$

$$= E_{eh} \cdot \frac{1/C_4(R_4 + R_i)}{p + 1/C_4R_o}$$

$$\text{put } C_4R_o = \tau_5, \text{ then } E_{ek} = \frac{E_{eh}\,1/\tau_4}{p + 1/\tau_5} \quad \ldots\ldots\ldots (2)$$

By comparing equation (2) with equation (1) it is seen that it is of the same form but with τ_5 replacing τ_4 in the denominator.

The equations for the equivalent circuit of Fig. 8 are:

$$E_{eb} = E_{ea}p/(p + 1/\tau_1) + E_{ek}\,1/\tau_1(p + 1/\tau_1)$$

$$E_{ec} = -A_1E_{eb}, \; E_{ed} = E_{ec}\,p/(p + 1/\tau_2)$$

$$E_{ef} = A_2E_{ed}, \; E_{eg} = E_{ef}\,1/\tau_3(p + 1/\tau_3)$$

$$E_{eh} = E_{eg} \text{ (since } A_3 = 1\text{)}, \; E_{ek} = E_{eh}/[\tau_4(p + 1/\tau_5)]$$

Writing the above in terms of E_{eb} yields:

$$E_{eb}\left[1 + \frac{pA_1A_2}{(p+1/\tau_1)(p+1/\tau_2)(p+1/\tau_3)(p+1/\tau_5)}\right]$$

$$= E_{ea}\,p/(p + 1/\tau_1) \quad \ldots\ldots\ldots\ldots (3)$$

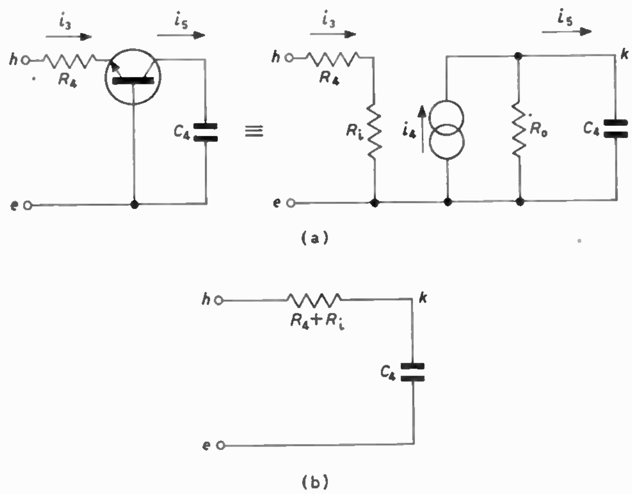

Fig. 9. Equivalent circuits of VT_7 (integrator stage)

By making $E_{ea}(p)$ a step input E_{ea}/p and re-arranging equation (3):

$$E_{eb} = \frac{E_{ea}(p+1/\tau_2)(p+1/\tau_3)(p+1/\tau_5)}{(p+1/\tau_1)(p+1/\tau_2)(p+1/\tau_3)(p+1/\tau_5) + pA_1A_2/\tau_1\tau_3\tau_4} \quad \ldots\ldots\ldots\ldots (4)$$

The denominator of equation (4) may now be shown as a polynomial.

$$p^4 + p^3[1/\tau_1+1/\tau_2+1/\tau_3+1/\tau_5]$$
$$+ p^2[1/\tau_1\tau_2+1/\tau_1\tau_3+1/\tau_2\tau_3+1/\tau_1\tau_5+1/\tau_2\tau_5+1/\tau_3\tau_5]$$
$$+ p[1/\tau_1\tau_2\tau_5+1/\tau_1\tau_3\tau_5+1/\tau_2\tau_3\tau_5+1/\tau_1\tau_2\tau_3+A_1A_2/\tau_1\tau_3\tau_4]$$
$$+ 1/\tau_1\tau_2\tau_3\tau_5$$

Let the coefficients of the descending powers of p be a, b, c, d and f, then Rouths criterion may be used to establish the limiting conditions for stable operation.

The criteria are:

$$d > 0, \; b > 0, \; bc - ad > 0, bcd - ad^2 - fb^2 > 0, \; f > 0.$$

The work of evaluation is seen to be extremely laborious but one may proportion the time-constants so that a solution is more easily achieved. Apart from stability there are other factors affecting the choice of time-constants which will be referred to in the next section. Luckily, this further consideration assists in the simplification.

Suppose the various time-constants are proportioned so that $\tau_5 \gg \tau_1 \gg \tau_2 \gg \tau_3$, the coefficients may then be re-written as follows:

$$a = 1,\ b = 1/\tau_3,\ c = 1/\tau_2\tau_3$$
$$d = A_1A_2/\tau_1\tau_3\tau_4,\ f = 1/\tau_1\tau_2\tau_3\tau_5$$

As far as criteria are concerned, d, b and f are all > 0. Now consider $bc - ad > 0$.

$$bc - ad = 1/\tau_2\tau_3 - A_1A_2/\tau_1\tau_4 \quad \ldots\ldots\ldots\ldots (5)$$

If τ_4 is taken as the variable, the limiting condition for stable operation is given by:

$$\tau_4 = A_1A_2\ \tau_2\tau_3/\tau_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

Now consider $bcd - ad^2 - fb^2 > 0$. Put $\tau_5 = k\tau_4$

$$bcd - ad^2 - fb^2 = A_1A_2/\tau_1\tau_2\tau_3^3\tau_4 - A_1^2A_2^2/\tau_1^2\tau_3^2\tau_4^2 - 1/\tau_1\tau_2\tau_3^3\tau_4k$$
$$= (1/\tau_1\tau_2\tau_3^3\tau_4)(A_1A_2 - 1/k) - A_1^2A_2^2/\tau_1^2\tau_3^2\tau_4^2 \quad \ldots\ldots\ldots (7)$$

Now $\tau_4 = C_4(R_1 + R_4)$ and $\tau_5 = C_4R_o$.

In the circuit $R_1 + R_4$ was about 3kΩ and R_o about 800kΩ. Thus $k = 266$. Since $A_1A_2 = 1\,200$ and $A_1A_2 \gg 1/k$, equation (7) may be rewritten:

$$bcd - ad^2 - fb^2 = A_1A_2/\tau_1\tau_2\tau_3^3\tau_4 - A_1^2A_2^2/\tau_1^2\tau_3^2\tau_4^2 \ldots\ldots (8)$$

The limiting condition for stability is thus given by:

$$A_1A_2/\tau_1\tau_2\tau_3^3\tau_4 = A_1^2A_2^2/\tau_1^2\tau_3^2\tau_4^2$$

or:

$$1/\tau_2\tau_3 = A_1A_2/\tau_1\tau_4 \text{ and } \tau_4 = A_1A_2\tau_2\tau_3/\tau_1 \ldots\ldots\ldots (9)$$

This relationship has already been given in equation (6). Thus, if k is large, when $bc - ad > 0$ is satisfied $bcd - ad^2 - fb^2 > 0$ is also satisfied.

Selection of Time-Constants

The factors governing the choice of time-constants are determined by the need to:

(1) Achieve loop stability

(2) Obtain maximum speed of response

(3) Introduce the minimum acceptable amount of distortion via R_1 to the grid of V_1 (Fig. 10).

The question of loop stability has been discussed in the previous section, where—in addition—the time-constants have further been determined by the need to simplify computation.

The speed of response will decrease as the time-constants decrease; so that the starting point for the calculation is τ_3 the time-constant of the amplifier. In other words for maximum speed of response the amplifier A_2 should have as wide a bandwidth as possible so that the relationship $\tau_5 \gg \tau_1 \gg \tau_2 \gg \tau_3$ can be satisfied with the minimum value for each time-constant. Care must be taken, of course, to ensure that any other bandwidths—that of A_1 for example—are wide in comparison with that of A_2.

By examining the circuits of Fig. 2 it is seen that the sinusoidal carrier is rectified by VT_6. A half-wave of current is then supplied by VT_7 to the capacitor C_4. The half wave of voltage developed across C_4 is fed via the buffer stages VT_{10} and VT_{11} to the resistor R_1. Consider now the input circuit when fed from a zero-resistance source (see Fig. 10).

The voltage at the grid of V_1 is given by:

$$E_{af} = E_{ad}\frac{R_1}{R_1 + 1/pC_1} + E_{ac}\frac{1/pC_1}{R_1 + 1/pC_1}$$

R_1 and C_1 are seen to form an adding circuit adding the sinusoidal input signal (E_{ad}) to the half wave across $C_4(E_{ac})$. Since the magnitude of E_{ac} is constant over the operating range, the percentage of distortion introduced will depend on the magnitude of E_{ad}, the ratio of R_1 to C_1, the frequency of operation and the magnitude of C_4. The estimation of the value of E_{ad} is better carried out by experiment than calculation since it involves—among other things—the ratio $(R_o + R_1)/R_4$ (Fig. 9) which is difficult to determine. It will be seen that C_4, in addition to its other duties, also operates as a smoothing capacitor. The practical value of C_4 may be much larger than that determined by considerations of stability simply in order to reduce the value of E_{ac}. Another consideration is the magnitude of E_{ac} must not extend beyond the limits of operation of the collectors of VT_7 and VT_8, the constant current sources. All these phenomena are easily examined with an oscilloscope, where malfunctioning is readily apparent.

Setting Up

Gain Control Amplifier

With no input signal, set the screen of V_1 to 1·6V. Note

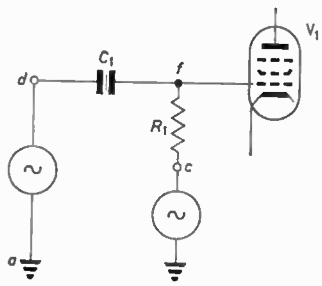

Fig. 10. V_1 input circuit

the voltage at point U (Fig. 2) and adjust RV_2 (Fig. 6) to bring point W to 1V above this voltage.

Integrator and Comparator

Connect 300mV r.m.s. to input of gain control amplifier. Adjust RV_3 (Fig. 2) to give desired output level at W (Fig. 6)—between 3V and 6V peak-to-peak.

Difference Amplifier

Short input and adjust RV_2 to give 6V d.c. at point S, connect input of difference amplifier and control amplifier in parallel and adjust RV_1 (Fig. 5) and RV_4 (Fig. 2) somewhere near the maximum and minimum input signal level in order to obtain the optimum degree of constancy of output level for change of input level.

Performance of System

Stability

The various time-constants for the 50c/s system were as follows:

$$\tau_1 = C_1R_1 = 0{\cdot}22\mu\text{F} \times 1\text{M}\Omega = 0{\cdot}22\text{sec}$$
$$\tau_2 = C_2R_2 = 1{\cdot}0\mu\text{F} \times 2{\cdot}2\text{k}\Omega = 2{\cdot}2 \times 10^{-3}\text{sec}$$
$$\tau_3 = 1/\omega_c = 1/(2\pi \times 7{\cdot}8\text{kc/s}) = 0{\cdot}02 \times 10^{-3}\text{sec}$$
$$A_1A_2 = 1{\cdot}3 \times 10^3$$

If these values are inserted into equation (6):

$$\tau_4 = 0{\cdot}26 \times 10^{-3}\text{sec}$$

Now $\tau_4 = C_4(R_4 + R_1)$ and $R_4 + R_1 = 3\text{k}\Omega$. Therefore $C_4 = 0{\cdot}087\mu\text{F}$.

The nearest integral value of C_4 which reliably stopped loop oscillation was 0·2μF. The difference from the theoretical value may be reasonably ascribed to a number of causes. These are: the fact that the theoretical value is a *limiting* condition, the approximation $\tau_3 = 1/\omega_0$, stray circuit capacitance etc. and the difficulty of observing the cessation of the oscillatory condition. If the integrator transistor VT_7 is 'bottomed' the loop gain will fall considerably and if it is not, then care must be taken to ensure that the variable-mu pentode is operating with maximum gain. The test procedure adopted was to short the input to the control amplifier and then to adjust RV_2 in order to apply the maximum negative voltage to point W—thus passing current through VT_6. RV_3 was then adjusted so as

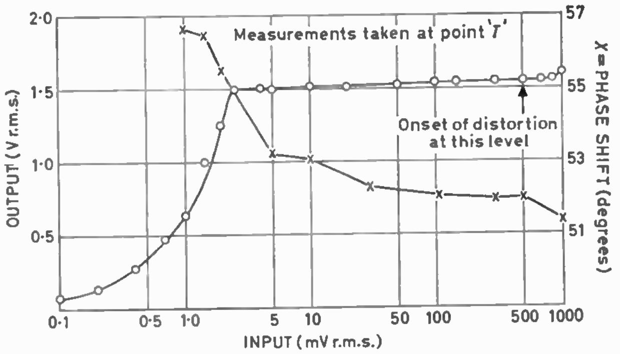

Fig. 11. 50c/s control amplifier. Amplitude control and phase shift characteristics

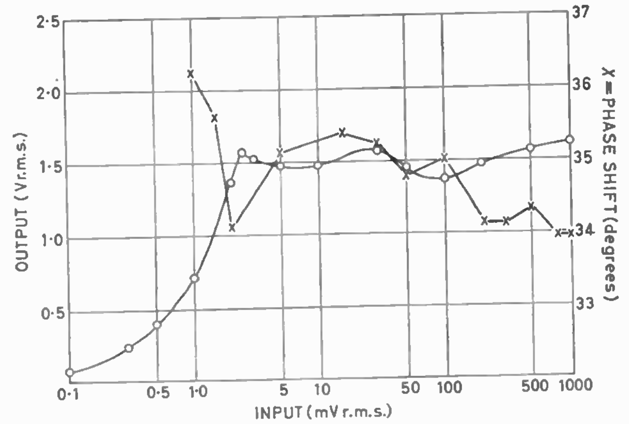

Fig. 12. 50c/s difference amplifier. Amplitude control and phase shift characteristics

to pass a current through VT_8 in such a manner as to swing the variable-mu pentode grid voltage slowly in a positive direction through the range. As the control grid passed through the region of maximum gain—prior to VT_8 bottoming—the output at W was monitored with an oscilloscope to observe whether oscillation took place.

However, the discrepancy is not serious since, though the order of capacitor is correctly obtained, the value adopted will be very much greater.

Control Characteristics

The measurements of the control characteristics were all carried out with a sinusoidal signal and with the level set to produce 4V peak-to-peak at point W on the control amplifier and with $C_4 = 100\mu F$.

Control Amplifier

The a.g.c. performance at 50c/s is shown in Fig. 11 together with the phase-shift over the control range.

Time of Response of Control Amplifier (to control to with 3dB of the set level)

Input change (mV r.m.s.)	*Time (sec)*
300-3	4·4
3-300	1·8

Difference Amplifier

The a.g.c. performance and phase shift over the control range is shown in Fig. 12.

Time of Response of Difference Amprifier

Input change (mV r.m.s.)	*Time (sec)*
300-3	11·8
3-300	6·5

Temperature Characteristics

The input voltage to the control and difference amplifiers was set to 300mV r.m.s. The three units were placed in an oven and the change in output voltage was −15 per cent for a rise in temperature from 20°C to 50°C. The change is mainly due to the change in d.c. level (at point W) of the fixed gain amplifier in the control channel.

Time Stability

The equipment was run for 24 hours and the output level of the difference and gain control channels recorded. The change in output level in both channels was less than 2 per cent.

Operation with Square Wave Carrier

It has been assumed so far that the carrier signal has been a sinusoid. The system is not directly suitable for dealing with a square wave carrier because of the differentiation carried out by the components C_2, R_2 (Fig. 6) and C_9, R_2 (Fig. 5).

The effect of differentiation may be overcome, however, and the two amplifiers should be considered separately. The values of C_2 and R_2 in the gain control amplifier cannot be altered since they form part of the time-constants already calculated. The solution is to pass the square wave signal through a low-pass filter before applying it to the grid of V_1 (Fig. 6). A simple RC filter only is required and its time-constant should be such as to remove the worst of the 'spike' produced by C_2 and R_2. So long as no overloading takes place the shape of waveform passing into the integrator is unimportant.

So far as the difference amplifier is concerned, the solution is to increase the value of C_9 as this is not part of the complete servo loop.

The 50c/s system already described was operated in this manner and the value of C_9 was changed to 25μF. The response time for an input change from 1V to 10mV peak-to-peak was 15sec.

The a.g.c. performance was substantially the same as that given in Figs. 14 and 17.

Whatever modifications are made to the inputs into the amplifier circuits it should be remembered that they must operate from as low a source impedance as possible.

Acknowledgments

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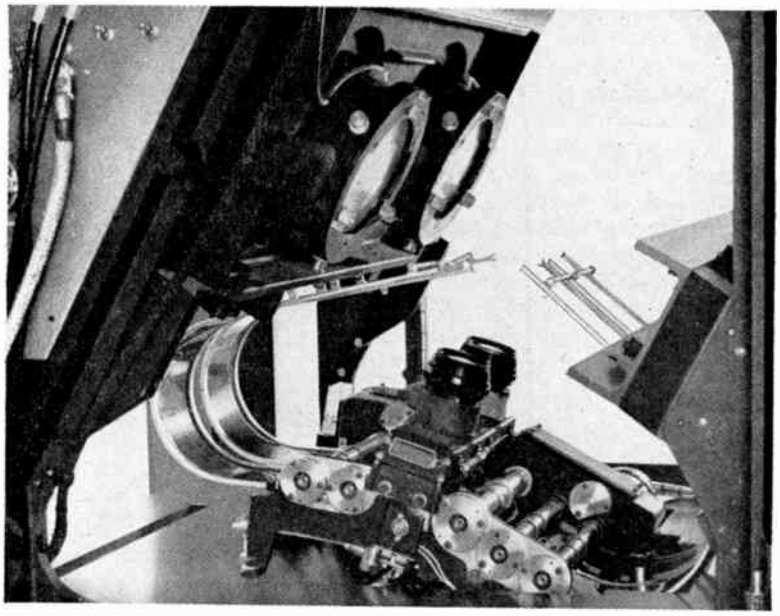
Xerography Applied to a High Speed Computer Printer



By P. F. T. C. Stillwell*, M.A., and R. H. Dagnall†, M.A.

A brief review of the requirements for commercial computer output printing is given. A printer for on-line or off-line operation from a digital computer, which uses xerography to produce a continuous printed record is described. Selected items of fixed data (e.g. form outlines) are projected from a film store and printed simultaneously with the variable data which are displayed as characters on a pair of cathode-ray tubes. Form selection and form layout are under programme control; monitoring circuits detect certain errors or machine misfunctions and cause an erroneous form to be clearly marked.

The paper moves at a constant speed of 8in/sec, permitting a printing speed of 2 800 lines/min.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 8□6)

ELECTRONIC digital computers were originally developed for scientific use, but during the last ten years they have been increasingly used for automatic control of industrial processes and in commercial applications. When used for the latter purpose the quantity of printing required to record the output from the computer is usually greater than in other applications and the operating speed of the output printer becomes a significant factor in the economic use of the computer.

Computer output printing involves two quite separate and largely independent operations one or both of which in the conventional mechanical printer necessitates the rapid movement of parts which have considerable inertia. The first operation is the selection or positioning of a type carrier and the second is the production of some kind of hammer blow to produce a visible impression on the paper. In the slower printers, for many decades associated with punched card tabulators, character selection is carried out by causing a type bar or type wheel to move to a position in which the required character is opposite the platen. In more rapid printers having a continuously rotating type drum, selection of a character involves a timing operation which is done electronically. In both types of printer a rapid hammer blow is necessary to produce the visible image on the paper without blurring.

One primary requirement of commercial data processing is that the output document should be a 'form'—that is to say, it must contain two kinds of information. One is the printing (e.g. the ruling, heading and fixed information) which is common to all forms of a group and the other is the variable information (e.g. account details, serial number, name and address) which can be different on every form. The latter is the content of the computer output data and the former is usually provided by the use of pre-printed stationery. Xerography, however, offers a process whereby the fixed part of the form, the form background or form outline as it will be called, may be printed simultaneously with the variable data and is the basis of the printer to be described.

Xerography

Xerography, a process invented in 1937 by C. F. Carlson[1] and developed by the Xerox Corporation of Rochester, U.S.A., makes use of the photo-conductive properties of selenium and the movement of charged particles in an electric field. Fig. 1 shows schematically a xerographic mechanism. An aluminium drum has a coating, about 80 microns thick, of amorphous selenium; it rotates at a constant speed and all the operations necessary to produce the printed image take place around its periphery. In darkness the selenium surface passes under a charging grid (called a scorotron), as a result of which it acquires and retains a positive charge at about 600V. The charged surface is then exposed to a light image projected through a slit, and the illuminated areas lose their charge by conduction to the earthed drum, while the areas that have received no illumination retain positive charge.

The charge pattern on the selenium is developed by cascading over it a powder—the developer—having two components. One part, the carrier, consists of hard spherical particles about 600 microns in diameter, the other part is the toner which is a fine thermo-plastic powder. The toner is the material from which the final image on the paper is formed and the carrier provides means for charging and distributing toner over the surface of the drum. The carrier and toner are agitated in the developing chamber and, by tribo-electric action, they acquire

The above photograph shows the dual formhead, showing cathode-ray tubes above

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charges of opposite polarity. The toner, being positively charged, is repelled from the positively charged areas of the drum, but is attracted to the high field caused by the steep potential gradient at the boundaries of the uncharged areas. The carrier particles being heavier and having considerable momentum, do not cling to the charged areas but roll off the drum and are returned to the developing chamber.

As the drum rotates it comes into contact with paper fed continuously from a roll and the toner image is transferred from the drum to the paper as a result of a negative charge applied to the paper by a transfer scorotron. The thermo-plastic constituent of the toner is fused by radiant heat and binds the toner particles to the fibres of the paper.

To prepare the selenium to receive a new charge, the remaining particles of toner are brushed away and the surface is discharged by illumination.

As explained above, the developing process is a boundary effect and this precludes the printing of large solid areas. It is, however, excellent for line printing because it increases line contrast, thereby improving the clarity of the printed character, which in turn permits latitude in exposure.

The toner is inert and unaffected by the usual bleaching agents used by forgers—an important feature when the forms are payment warrants or insurance policies.

The Xerographic Printer

In the printer the form outline and variable data are projected on to the xerographic drum simultaneously from two different sources[2]. The form outline is selected by programme from a number of such outlines stored on strips of photographic film and is then projected by the formhead to form an image at the surface of the selenium drum. Since the drum is rotating continuously the image thrown on it must move in synchronism with its surface to maintain alignment and prevent blurring. The firm is given a slow scanning motion for this purpose.

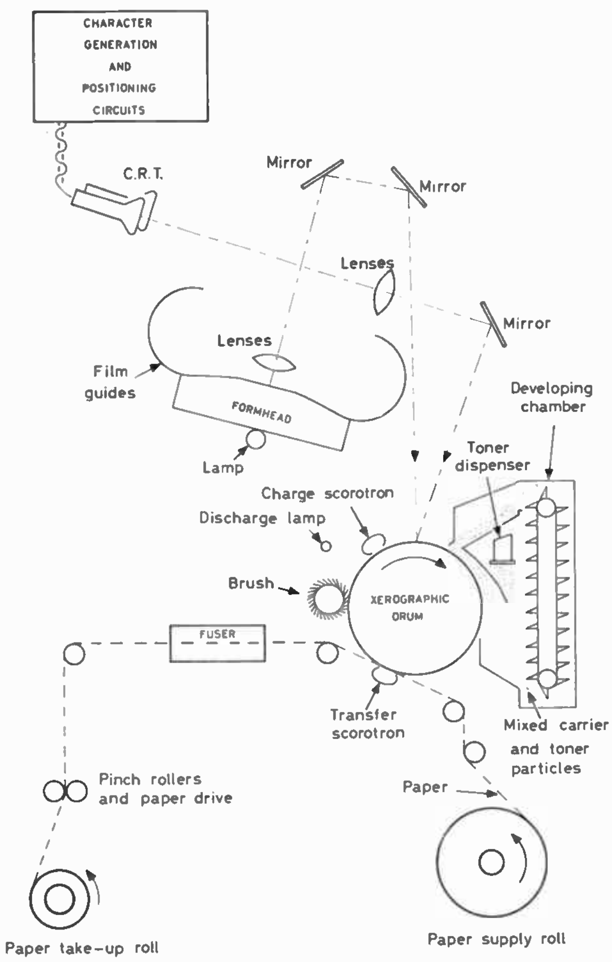
The variable data are displayed as legible characters on the face of a cathode-ray tube, and a separate optical system forms images of these characters on the surface of the selenium drum. In order to compensate for the movement of the surface, the character positions are shifted on the face of the cathode-ray tube by a voltage which varies linearly with time and which is reset at the beginning of each printed line.

Duplicate forms are a usual requirement of office procedure and it is arranged that data printed on the left-hand half of the paper can also be printed simultaneously on the right-hand half, two cathode-ray tubes being used for this purpose.

The paper web moves through the printer at a constant speed of 8in/sec, a rate that determines, in one way or another, the timing of all the printer functions. A form 4in deep is completed in 0·5sec and the printing of lines spaced one-sixth of an inch apart must be completed within 20msec, allowing for timing tolerances. This continuous movement of paper demands that the printer receives the next item of input data (e.g., a printable character) whenever it requires it, without undue delay. If the computer can emit successive data items 'on demand' then on-line operation of the printer is possible, otherwise off-line working from a magnetic tape unit through a buffer store is adopted.

Individual forms are severed from the roll in a separate cutting operation on an off-line paper guillotine. The form outlines on the film include a cutting mark that is used to trigger the guillotine control circuits.

Fig. 1. The xerographic printer



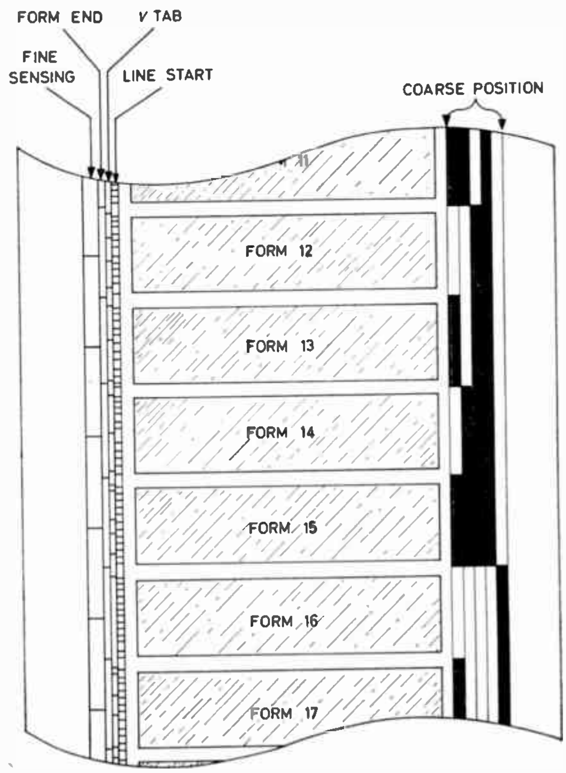


Fig. 2. Portion of form master showing relative positions of marker tracks

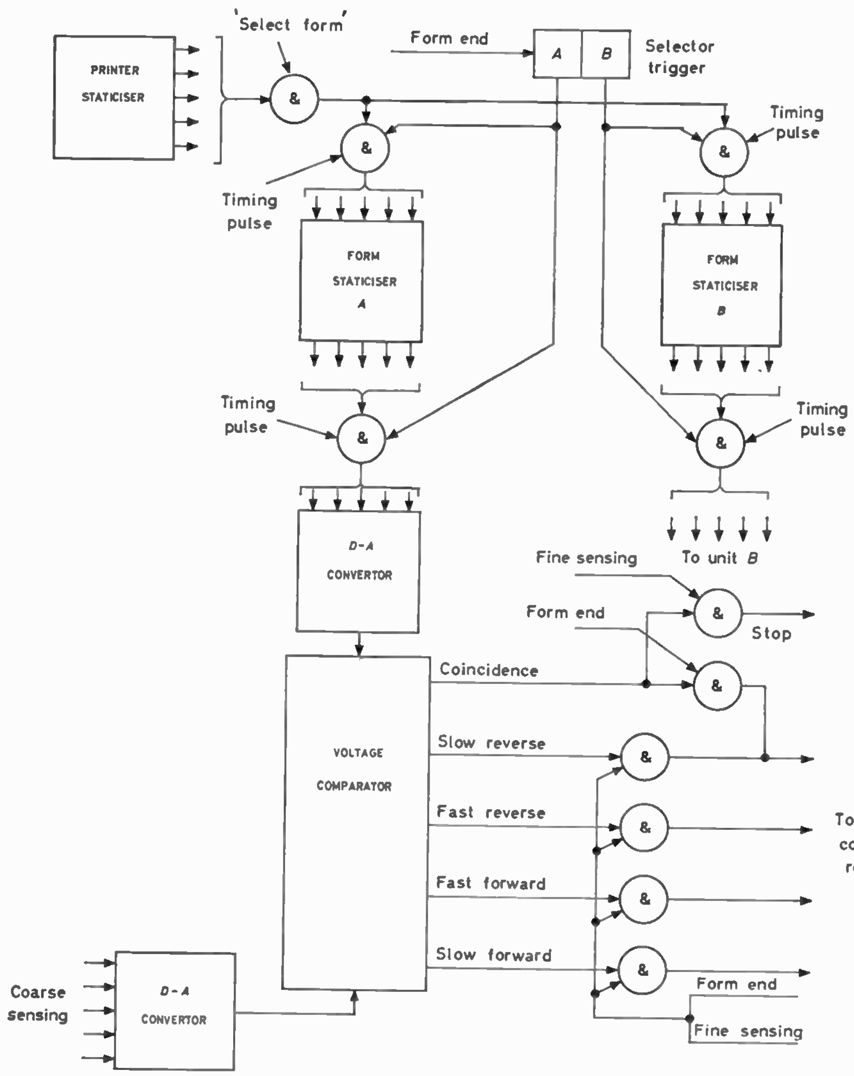


Fig. 3. The form positioning

a magnification of 11:1 to give an image 24in wide on the selenium drum. Five coarse-position markers serve to identify the form number and are sensed by photocells to determine the necessary direction and speed of drive during a form selection sequence. The fine-position marker is also sensed by a photocell and the output signal is used both to strobe the signals from the coarse-sensing circuits (thus avoiding ambiguities when the code changes), and also to stop the drive when the coarse-sensing outputs correspond with the desired form. The 'form end', 'V-tab' and 'line-start' markers are projected and are sensed by photocells effectively at the surface of the xerographic drum. Only the form-end timing signal is used in the formhead control circuits.

The logical arrangements for controlling a form master (*A*) are shown in Fig. 3. The form numbers of two consecutive forms, namely the one being printed and the next one being positioned, are stored in staticizers *B* and *A* respectively. The contents of a staticizer are converted to an analogue voltage and compared with a voltage similarly derived from the outputs of the coarse-sensing photocells. If the voltages are equal to within $\pm\frac{1}{2}$ unit (where a unit is the voltage change corresponding to a binary change of one) a coincidence output is obtained, otherwise the appropriate slow or fast, forward or reverse output is generated. A fast drive output is necessary if the difference between the form number stored and form position is greater than three.

Control is transferred from one unit to the other alternately by gates operated from a selector trigger which is switched from one state to another by the form end signal. Assuming that the next number has just been set up in form staticizer *A*, a form end signal gates the output from the comparison circuit and initiates the appropriate drive, for example the fast forward drive. The film moves and each time a fine position marker is detected, the comparison outputs are sampled. When the film is within three forms of the required position, the fast drive is disengaged and the slow drive becomes effective. Finally when coincidence is detected the AND gate operates and removes the drive. The vacuum drag on the film ensures rapid deceleration and the film is left in the correct position for starting a scan.

Printing of the form positioned in unit *A* is initiated by a 'scan' signal which opens the projection shutter and engages the scan drive, the film then moves at a rate which ensures that the projected image remains stationary relative to the moving surface of the selenium drum. At the end of the form a form end signal is sensed and the positioning cycle recommences in unit *A*.

It is not possible within the scope of this article to give a complete account of the design of the printer, but the sections which follow describe the main features of the formhead from which the timings are derived, the printer control and the character display.

The Formhead[3]

The form outlines are stored on a strip of film, called the form master, 3¼in wide and about 22in long, which is driven backwards or forwards to bring the beginning of the required form into the projection aperture and then mechanically scanned during exposure. To avoid gaps between forms in the output print the form master and its control are duplicated allowing the next form master to be positioned while the previous one is being scanned.

The two films are mounted side by side just out of contact with five sets of rotating rollers of which four are for positioning and the fifth for scanning. Co-operating with these driving rollers are two sets of solenoid operated pinch rollers which can be operated individually to drive either film.

When the film is stationary or moving slowly (4·5in/sec) the perforated film gate is connected to a vacuum line so that the film is held in close contact with the gate, but during a fast drive (70in/sec) pressure is applied and the film runs on an air cushion. Fig. 2 shows part of a form master. The form area is 2·2in wide and is projected with

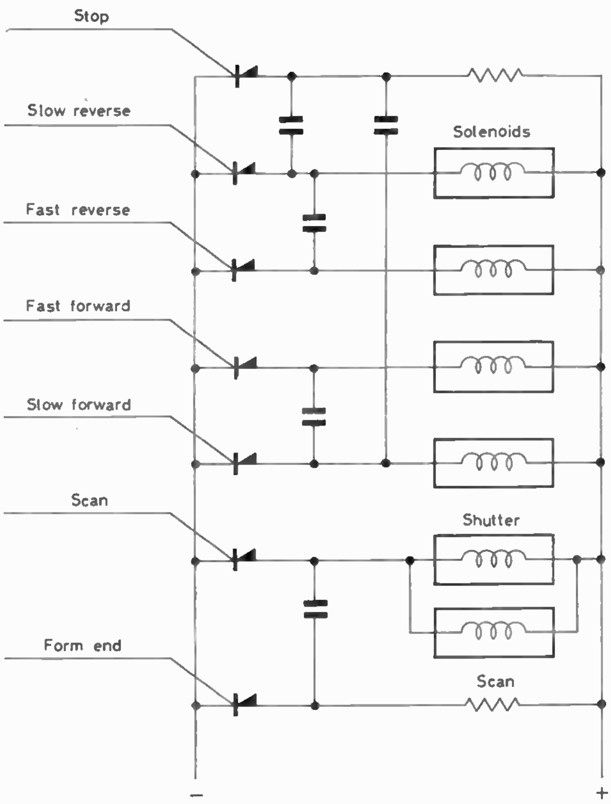


Fig. 4. Switching circuit for formhead drives, using silicon controlled rectifiers

The scan signal which starts the scanning operation can either be derived from the form end signal associated with the previous form or generated independently. If the former then the sheet length varies with the length of printed form, otherwise the sheet length is independent of the length of form.

Fig. 4 shows the drive for the film-positioning pinch rollers. The solenoids of the film-positioning drives are controlled by silicon controlled rectifiers which are mutually exclusive.

A single high-pressure mercury vapour lamp, driven by a 1·2kc/s solid state invertor, is used as the projection lamp for both films and also provides the light source for the photocells. Since the light output from the arc has both a steady and a ripple component, the a.c.-coupled photocell amplifiers give outputs both when the film is stationary and when it is moving rapidly.

Print Control, Character Generation and Display

The printer logic demands one code combination (i.e. an array) at a time, the input is, therefore, serial by arrays and parallel by bits. Input arrays can represent either printable characters or commands for control purposes. When read into the printer logic the latter arrays are decoded on to separate lines which control the appropriate functions.

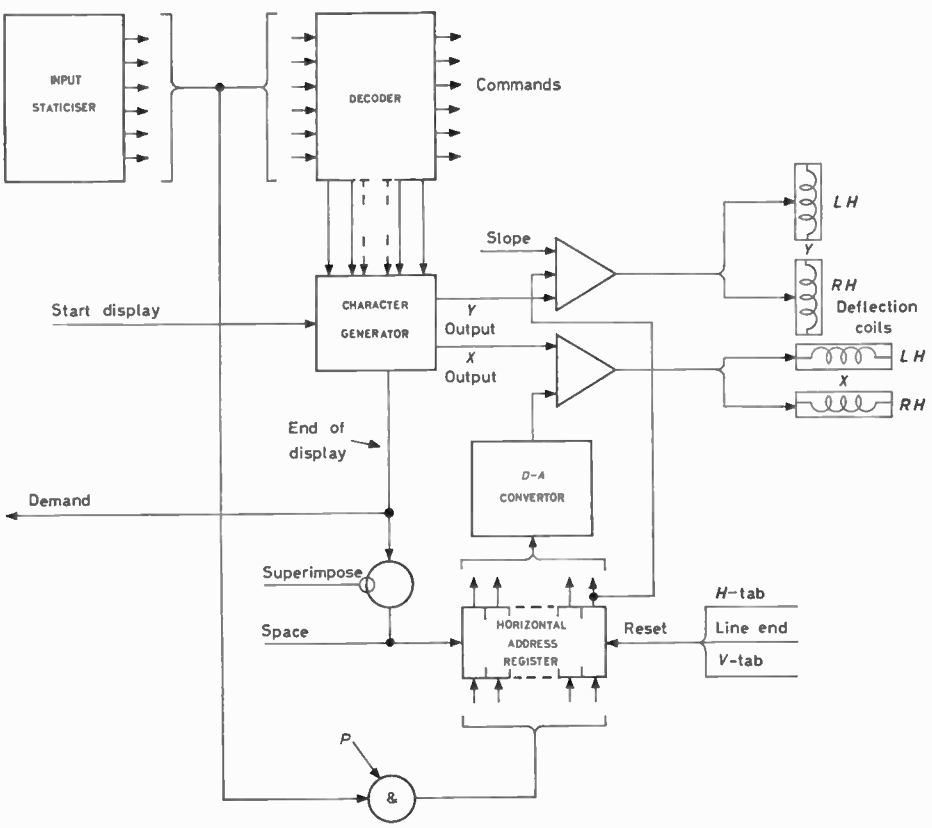
Certain commands are composed of two or more arrays. The first array defines the command and the others convey numerical information concerning the command, the desired form number or a tabulation position, for example.

Timing of the printer functions is controlled by the three tracks of timing marks on the form master (see Fig. 2). The form end marker denotes the end of the current form. Line start markers which control the time at which the 'demand' for the first data array of a line is emitted, are placed wherever variable printing may be required on the form, with irregular spacing if necessary. The V-tab markers are used to define vertical positions and, except for the first one, which is used logically, are optional.

The column position of any character is determined by the decoded contents of the horizontal address register (Fig. 5). The register is reset to zero by the commands 'line end' and 'V-tab', both of which return the printing position to the left-hand margin. It is also reset by 'H-tab' as a preliminary to horizontal tabulation, but during the second array of an 'H-tab' command a control voltage P sets the horizontal address register to correspond with the binary value held in the input staticizers. A 'space' command advances the horizontal address register by one position as does the 'end of display' pulse. The latter may be inhibited by a 'superimpose' command which then allows two characters to be superimposed (for example for underlining or the insertion of accents).

Vertical positioning of the variable data with respect to the form outline is controlled by generating the first 'demand' of a line only when a 'line-start' marker is

Fig. 5. Arrangement of character generation and horizontal tabulation



sensed. Vertical tabulation is effected by inhibiting the normal 'line start demands' during a 'V-tab' command. The second array of the command, indicating the V-tab position required is compared with the contents of a counter which counts the V-tab markers. When the two are equal, the next line start marker initiates the demand for the next array.

Characters are built up on the faces of two cathode-ray tubes as a series of dots[4]. In order that the tubes may be small and efficiently used the line of characters is split into two half lines of 64 character positions which are displayed one above the other, as shown in Fig. 6. Associated with each tube is a pair of lenses which produces a pair of images on the selenium drum, butted side by side but offset vertically. The upper line of one image is aligned with the lower line of the other image to form a complete line of 128 columns and the two unused images are masked off by the slit aperture as shown in Fig. 7.

The display of a character is initiated by a 'start display' signal generated whenever a character array is decoded.

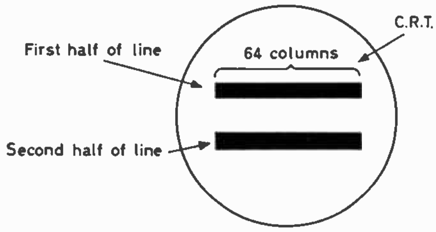


Fig. 6. Layout of line on c.r.t.

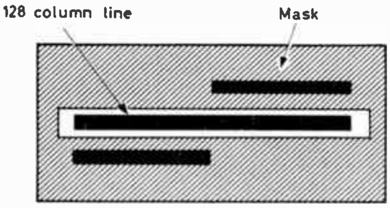


Fig. 7. Image of 11½in wide line

Subsequently X and Y voltages for the required character shape are generated together with a bright-up waveform. The X character deflexion voltage is added to an analogue voltage derived from the contents of the horizontal address register; the Y character deflexion voltage is added to a 'slope' waveform which compensates for the motion of the selenium drum and ensures that the line is printed axially along the drum. The Y waveform also contains a voltage derived from the horizontal address register to provide the double-line display.

The two cathode-ray tubes are fed with identical deflexion waveforms and consequently it is not possible to print different characters in corresponding columns of the two halves of the paper simultaneously, but because the bright-up waveforms of the tubes can be independently controlled different symbols can be placed in corresponding columns by a serial process.

The cathode-ray tube images are projected on to the selenium drum at a magnification of approximately 1:1 and each line on the drum and on the final print measures 11½in for 128 characters. Magnetic deflexion is used and the deflexion amplifiers contain circuits which compensate for the time-constant of the scanning coils as far as the available power supplies and the circuit voltage limits allow, but during horizontal tabulation and flyback periods 'demand' is delayed by approximately 200μsec to allow the beams to reach the next required position.

Printer commands are normally completed in 25μsec except those requiring paper movement ('V-tab') and those which call for a large change in magnetic deflexion ('H-tab'); a character display is completed in 200μsec.

Error Detection

In a printer used to provide forms which will be sent out to the public, reliability and freedom from error are most important. Failure in these two repects would offset the advantage that increase of speed makes possible, besides causing much inconvenience to the user and possible damage to his reputation. The main basis of mechanical reliability has already been mentioned, namely the absence of any fast moving parts in the printer mechanism. The electronic part of the equipment embodies numerous monitoring and checking circuits which are of three types:

(a) On the validity of information and control signals. Parity checks are made on the data; the waveforms appearing across the deflexion coils are compared with the setting of the horizontal address register; the bright-up waveform applied to the grid of the cathode-ray tube is monitored to ensure that the tube is operative during a character display and cut-off during a command.

(b) On the sequence of certain control signals. The sequence of scan and form end in the formhead is checked; the alternation of line start signals with 'line end' commands is monitored.

(c) On the mechanical operation and electrical circuits of the xerographic printer.

A fault which makes it impossible or dangerous to continue printing (e.g. formhead lamp failure or a fault in the c.r.t. deflexion or bright-up circuits) is arranged to stop the printer. Other faults, which may only effect one form, cause the error mark printer, which is an electronic flash lamp focused on the selenium drum, to imprint a distinctive mark which can be optically sensed in subsequent paper handling, on the form effected.

Results

Some minor problems—and not a few major ones—have had to be solved due to the very nature of the methods chosen to effect the final result.

For example the combination of two optical systems, one having a magnification of 1:1 and the other of 11:1, could lead to difficulties in registration between the variable printing and the form outline. Dimensional stability of the film must be of the highest order and a special film base, originally developed for aerial map making, is used. Registration also depends on the accuracy with which the characters are positioned on the cathode-ray tubes; this can be affected by variations in the e.h.t. voltage, in the performance of the deflexion amplifiers and in the departure of the magnetic deflecting field from strict linearity. Nevertheless it has been found possible to achieve very accurate and stable horizontal and vertical registration between the variable data and the form outlines.

Acknowledgments

The printer described in this article is manufactured by the Rank Organization under their trade mark 'Xeronic'.

The authors wish to thank their colleagues who helped to develop the printer, in particular Messrs. J. H. Ede, P. J. Greaves, A. S. Pratt, T. C. Reeve, R. Robertson and B. C. Sewell. They also wish to acknowledge the permission of the Rank Organization to publish this article.

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An Electronic Device for Counting and Recording Insects in Agricultural Research

By E. J. Brach* and W. J. Mason*

To combat the insects in wheat and flour, entomologists must study their habits. For example, how many insects are in a unit of wheat or flour and when is their period of greatest activity? An instrument is described for determining these factors. It consists of an insect trap, optical transducer, counting circuit, timing unit and printer which records the number of insects emerging from a container of flour or wheat to forage for water.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

To determine the number of insects in samples of wheat and flour the number of insects per sample are counted. Manual methods are time-consuming and an automatic device is therefore required.

The insects are forced to emerge from the sample for counting by a simple device (Fig. 1). Insects climbing a tube in an attempt to reach water suspended above, fall down in a chute (*b*) placed in the path of an optical transducer. The insect interrupts a light beam thus actuating an impulse counter. The system described will count insects whose size is from 0·25mm to 5·0mm long.

Photohead

A 14V lamp (*e*) with a V-shaped filament operated by a regulated power supply is focused on a photo transistor (*h*). The photo transistor has a sensitive area of 40mm^2 and is a germanium junction transistor. Its light sensitivity is 15μA/ft-c. Its h_{fe} value is 160.

For reliable operation, the output change appearing across R_3 of the optical transducer should be a minimum of 180mV or 12μA (Fig. 2).

The quantity of light Φ falling on the sensitive area A of the photo transistor, will eventually determine the sensitivity of the instrument.

$$\Phi = \frac{I \times A}{d^2}$$

where I is the candlepower of source, and d is the distance between the light source and the photo transistor. By selecting a lamp whose output is 1·4 candlepower, and placing it 70mm from VT_1, Φ is 11×10^{-4} lumens. Since the sensitivity of VT_1 is 350mA/lumen, 11×10^{-4} lumen will introduce 318μA. This is the current that will flow in the collector circuit of VT_1 when illuminated by the above calculated amount of light.

The beam of the light source to the sensitive area of VT_1 is parallel and much wider than the minimum insect size of 0·25mm. Because of the small size, the current change actuated by the insects will be small.

The sensitivity is improved by focusing the light beam so that an insect 0·25mm long will blank out most of the light beam falling on the sensitive area. This is achieved by inserting a lens between the light source and the photo transistor. The equation of Φ will change to:

$$\Phi_L = \frac{I \times A_L}{d^2}$$

where Φ_L and A_L is the amount of light falling on the lens, and the area of lens respectively. If the lens is 40mm from the light source and has a focal length of 15mm, Φ_L will be 0·162 lumen and magnification of the lens is 147.

If the focal length of the lens is f and d' the distance between the lens and the beam focus point:

$$(1/f) = (1/d) + (1/d') \text{ or } d' = 29\text{mm}.$$

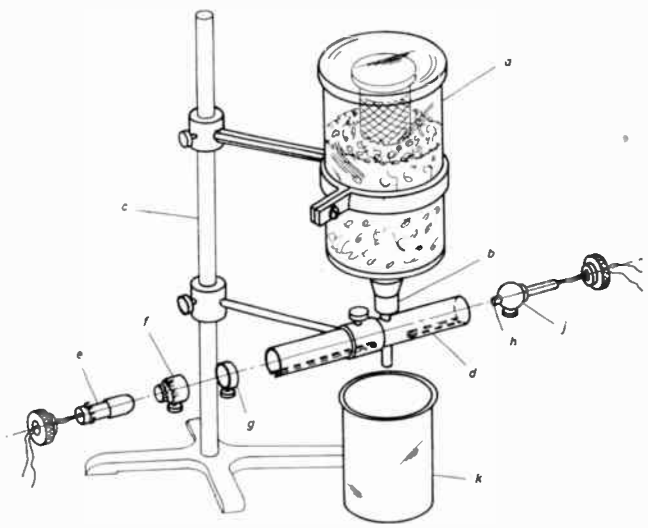

Fig. 1. Insect trap and photohead

(*a*) *insect trap*	(*f*) *lampshade*
(*b*) *funnel and chute*	(*g*) *lens*
(*c*) *stand*	(*h*) *phototransistor*
(*d*) *photohead assembly*	(*j*) *phototransistor holder*
(*e*) *lamp*	(*k*) *collecting dish*

The photo transistor is placed 30mm from the lens so that the image of the lamp element is slightly out of focus and the photo transistor is not overheated.

When light falls on a photo transistor, an I_{ph} photoelectric current develops an internal base current. If the base is an open-circuit, the current is amplified by $(1 + a)$ at the collector where a is the current amplification factor. To ensure stability at a wide temperature range, a voltage dividing network R_1 and R_2 in conjunction with an emitter resistor R_5 is used. Any increase in emitter current causes a large voltage drop across R_5 and reduces the base-emitter voltage, thus reducing the collector leakage current.

To reduce the dark current and improve the light to dark current ratio, a resistance (R_4) is inserted between the base and point *a* at the voltage dividing network. The dark current without R_4 in the circuit is 225μA, and with R_4 in the circuit it is reduced to 22μA. The signal from VT_1 is coupled through C_1 to a two-stage switching circuit.

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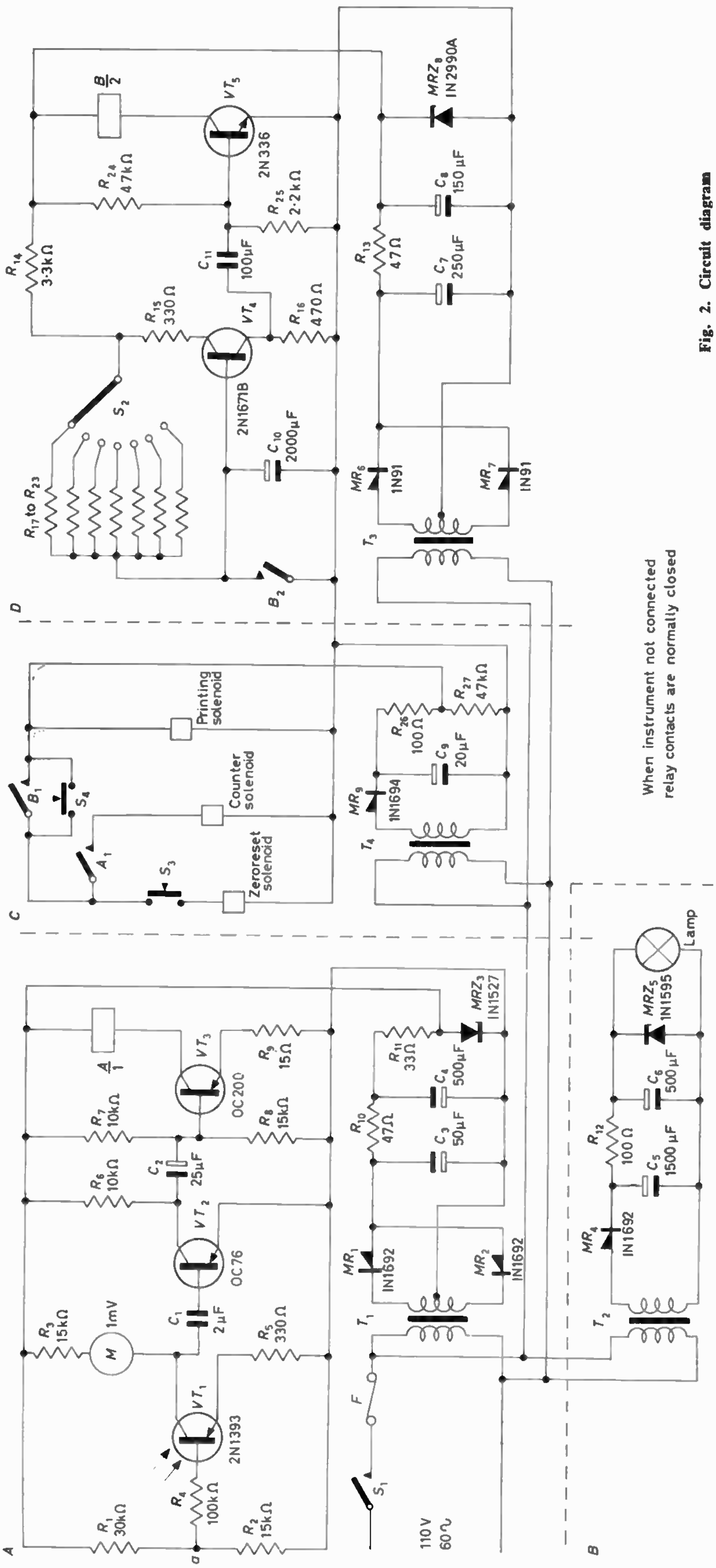

Fig. 2. Circuit diagram
(A) counting circuit (B) lamp and lamp supply
(C) circuit of electromechanical counter and printer
(D) timing circuit

Counter Circuit Description

When photo transistor VT_1 (Fig. 2) is illuminated, $400\mu A$ photo-current flows through it generating $-22V$ at its collector. Interrupting the light beam reduces the voltage at the collector to approximately $-10V$. This $-12V$ pulse generated at the collector of VT_1 is coupled through C_1 to VT_2. VT_2 is an alloy junction transistor of the germanium pnp type, and is operating in the bottomed or current saturated condition. This is achieved by lowering its collector by a resistive load (R_6) from its normal condition $a'I$ where a is the large signal d.c. amplification factor. Both the collector and emitter diodes are forward biased. The pulse introduced on the base of VT_2 switches the collector voltage between the bottoming voltage and a voltage nearly equal to the supply voltage. The pulse introduced by this voltage change is coupled through C_2 to VT_3, the relay switching circuit, with the relay A being the load resistance. In the on condition VT_3 is bottomed. The current in the load rises to V_s/R_L where V_s = supply voltage and is equal to $-22V$. The d.c. resistance of the relay coil is $2{\cdot}5\Omega$ and the load current is:

$$I_L = 22/2500 = -8{\cdot}8\text{mA}$$

When the pulse at the base of VT_2 is positive, VT_2 will cut off. A positive base current of approximately $2\mu A$ flows into VT_3 bringing it out of bottoming and tending to switch it off. The induced voltage makes the collector potential rise almost immediately to the supply voltage, cutting off the relay. The contacts of relay A actuate an impulse counter.

When the light is interrupted with an object the size of 5mm, a 20mV pulse appears at the base of VT_1. At the collector, the voltage is 800mV. The pulse is amplified by $A_{VT1} = 800/20 = 40$.

At the base of VT_2, the pulse is attenuated to 400mV and an amplified pulse of 1 000mV appears at the collector. The amplification of this stage is $A_{VT2} = 1\,000/400 = 2{\cdot}5$. The pulse at the collector of VT_3 is 50 000mV. The amplification of the third stage is $A_{VT3} = 50\,000/1\,000 = 50$. The total amplification for an object intersecting the light 5mm diameter or larger is:

$A_{total} =$

$$A_{VT1} \times A_{VT2} \times A_{VT3} = 5\,000$$

Intersecting the light with an object 0·1mm diameter, the pulse height at the base of VT_1 is 7mV, at the collector it is 75mV. $A_{VT1} = 75/7 = 10{\cdot}7 \simeq 11$. The pulse height at the base of VT_2 is 30mV, at the collector it is 600mV, $A_{VT2} = 20$. The amplification of VT_3 is $A_{VT3} = 50\,000/500 = 100$. The total amplification for an object intersecting the light with 0·1mm is:

$$A_{total} = A_{VT1} \times A_{VT2} \times A_{VT3} = 22\,000$$

The circuit is powered by a −22V regulated supply based on Zener diode MRZ_3. The regulation is ±1 per cent with a line voltage change of ±15V. The ripple is 0·01 per cent.

To achieve reliable performance, the lamp supply voltage must also be regulated. A second Zener diode MRZ_5 in conjunction with a rectifier circuit (Fig. 2(*B*)) is installed for this purpose with the same performance as the main power supply.

A third power supply is used for the Sedeco 1Tpb3 impulse counter and printer (Fig. 2(*C*)). The unit is a six-digit electromechanical counter with an electrically operated zero reset (S_3) and a printing mechanism controlled by a timer. The voltage needed to energize the coils is 110V, the load and energy consumption of each of these coils are 15W.

The adjustment of the instrument for maximum sensitivity is made by connecting a 1mV panel meter in the collector circuit of VT_1, and adjusting for maximum collector current.

Description of Timing Mechanism

To determine the period of maximum activity of the insects, a timer operates the printing mechanism of the counter at predetermined time intervals.

A timer was designed to suit the specifications of the printing coil so that the duration of the operating impulse shall not be smaller than 150msec and shall not be larger than 5sec.

The timer is a solid state time delay circuit using a unijunction transistor VT_4 (Fig. 2(*D*)) in connexion with an npn silicon transistor VT_5. Timing is initiated by applying supply voltage to the circuit, the timing capacitor C_{10} is charged through R_{17} to R_{23} until the voltage across C_{10} reaches the V_p (peak point voltage) of VT_4, at which time VT_4 will fire, generating a pulse across R_{16}, and will turn off VT_5. VT_5 is initially in the on position, being in the bottomed state. The load of VT_5 is the 5kΩ coil of relay *B*, when VT_5 is turned off relay *B* is de-energized. One set of the contacts actuates the printing coil, the second set is connected across C_{10}.

The time delay of the circuit depends on the time-constant of $R \times C_{10}$. *R* represents any of the selected resistors from R_{17} to R_{23}. Capacitor C_{10} is charged through resistance *R*, from the voltage supply V_B to a voltage V_e where V_B is 24V and V_e is 3·2V, when V_e reaches 3·2V, C_{10} will discharge through R_{16} and the base resistance R_{B1} of VT_4, and this will end the time cycle. If the loading effect of VT_4 on V_e is neglected, the charging time *T* of the capacitor is given by:

$$T = R \times C_{10} \ln \frac{V_B}{V_B - V_e}$$

This equation states that the change in time produced by a change of *R* or *C* is linear, while a change in V_B or V_e is exponential. Fixing the values of V_B, V_e and C_{10}, *T* will depend only on *R*.

Selecting the appropriate resistance with S_2, T_1 will vary from 60 to 1800sec. The accuracy of the time interval depends on the stability of *R*, C_{10}, V_B, and V_e. V_B is regulated and resistor R_5 with Zener diode MRZ_8 provides a stable power supply for the unijunction transistor. The maximum time delay in that circuit depends on I_p, the peak point current of VT_4, the maximum ambient temperature and the leakage current of the capacitor and I_{eo} of VT_4 at the maximum ambient temperature. The upper leakage of *R* is determined by the expression:

$$R < \frac{(1 - \eta)\, V_{B(min)}}{25\, \dfrac{I_p + I_c}{V_{B(min)}}}$$

where η = the maximum value of intrinsic stand-off ratio. Thus the current flowing into the emitter at the peak point must be greater than the peak point current for VT_4 to turn on.

The lower limit of *R* is determined by the requirement that the load line formed by *R* and V_B intersect the emitter characteristic curve to the left of the valley point, otherwise VT_4 will not turn off.

To temperature stabilize VT_4, R_{15} is connected in series with R_{B2}.

In this circuit a high value capacitor is used for the timing capacitor. It would be desirable to use a tantalum capacitor, but it is difficult to find a tantalum capacitor of the required value of 2 000μF. The leakage current of the capacitor and of R_{B1} of VT_4 had to be taken into account, and was corrected by multiplying the *R* values of the timing circuit by a factor *k*. To improve the accuracy of the timing, one contact of relay *B* is connected across C_{10}. When relay *B* releases momentarily, contact B_2 closes, shorting the timing capacitor for a rapid discharge, and for setting a uniform initial voltage at the beginning of the timing cycle.

Tests Performed on the Circuit

100 per cent counting and printing, and ±3 per cent accuracy for the timing device are required. These accuracies were verified by the following tests:

(a) Life
(b) Temperature
(c) Accuracy
(d) Effect of line voltage
(e) Rate of counting.

For the life test, the instrument was operated continuously for 336 hours. At the end of this period, voltage measurements were taken, and compared with measurements taken before the life test, and they showed exact similarity. The components were checked for overheating and the maximum component temperature was 40°C. After the life test, the instrument was taken to a chamber where the temperature was varied from −15°C to +60°C.

TABLE 1

POSITION (SEC)	WORST READING (SEC)	ERROR (PER CENT)	NO. OF TESTS PERFORMED
60	58·9	1·88	100
300	295	1·66	100
600	588	2	75
900	833	1·88	75
1200	1178	1·83	75
1500	1472	1·86	50
1800	1764	2	50

The performance of the instrument was not changed at any time during this test although the semiconductors tended to overheat at ambient temperatures above 60°C.

During the test time, accuracy was maintained. While the above tests were being performed, the instrument was closely watched to verify its accuracy. A dish was placed under the photohead (Fig. 1(k)) to collect the counted insects. Every hour these insects were compared with the readings of the impulse counter and printer. There was not one case where a miscount could be found. Periodically, the timer was tested for repeatability and accuracy, and the worst deviations from the times set are shown in Table 1.

The response of the instrument to line voltage variation was tested by connecting the instrument to a variable transformer. The accuracy of the counting and printing section of the instrument was not affected by changing the line voltage from 103V to 130V.

Tests were made to determine the counting speed of the instrument, and it was found that 300 counts/min on the smaller insects was obtained without affecting the accuracy of the instrument.

Conclusion

The instrument proved very useful in determining the number of insects in a unit of wheat, barley or flour. Every effort has been made in the designing of the photohead to obtain maximum sensitivity. By choosing the correct working point in the design of the switching circuit; a relatively simple circuit performs with accuracy and is reliable.

A unijunction transistor, in connexion with an npn transistor, provides a wide-range variable time delay, with acceptable accuracy.

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Voltage Stabilization in Hybrid Electronic Apparatus

By P. Visontai*, A.M.I.E.R.E.

In electronic apparatus using both valves and transistors, the provision of supply voltages for the transistors presents special problems. If one of the valves passes as much d.c. as the transistorized parts of the apparatus need, it may be possible to use a single power supply. An arrangement is described here making possible the stabilization—virtually at no extra cost—of the low supply voltage of one group of transistors against variations in the d.c. consumption of a different group of transistors, fed from the same supply.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

IN partly transistorized electronic equipment, i.e. in equipment using both valves and transistors, generally two power supplies are required, one to provide the high voltage for the valves, and another, of much lower voltage, to feed the transistors.

Under certain conditions, however, it is possible to use the same power supply to serve both the valves and the transistors, resulting in substantial saving in the cost and weight of the equipment.

The purpose of this article is to describe one such arrangement and to demonstrate how the low supply voltage of one group of transistors can be stabilized—virtually at no extra cost—against variations in the d.c. consumption of a different group of the transistors, fed from the same supply.

Assume a hypothetical case, where one of the valves draws as much cathode current (i.e. anode plus screen current in the case of a pentode) as the transistorized part of the circuit needs. It is further assumed, that the cathode of the valve can be bypassed to earth at the frequency of operation. Also, that the valve is able to fulfil its original function satisfactorily at a somewhat reduced anode and screen voltage.

Fig. 1 shows the arrangement in simplified form. There is only the high voltage power supply, providing the current needed by all the valves.

The voltage and current needed for the transistors is taken from the cathode of valve V. If the grid is returned to a fixed potential, it is a straightforward cathode-follower arrangement. In those cases where the current consumption of the transistorized circuits is essentially constant, this is a very satisfactory solution. If, on the other hand, the transistorized circuits need a varying amount of current, the voltage supplied to them will be subject to variation, which may be undesirable.

By defining the mutual conductance of the valve related to the cathode current—rather than to the anode current —i.e.

$$g_c = dI_c/dV_{gc} \quad \ldots\ldots\ldots\ldots\ldots\ (1)$$

where

g_c is the mutual conductance,

I_c is the cathode current,

V_{gc} is the voltage between grid and cathode,

it can be very easily shown (assuming linear valve characteristics) that the variation of the cathode voltage V_c with cathode current I_c is:

$$\Delta V_c = -(1/g_c)\Delta I_c \quad \ldots\ldots\ldots\ (2)$$

If the transistorized circuits operate in such a manner

* *Radio and Allied Industries Ltd.*

that some of the transistors draw constant current and need constant voltage, whereas other transistors draw varying amounts of current and can tolerate changing supply voltage, a development of the basic arrangement of Fig. 1 can give a solution. (An example of such circuit is a multi-stage tuned amplifier, where the gain-control is effected by altering the d.c. of one or more of the stages.)

In Fig. 2 the variable resistor R_1 symbolizes the group of transistors consuming a varying amount of current, and R_3 substitutes the transistors drawing constant current. The grid of the valve is returned to a potential divider across R_1. It will be shown that at a particular division ratio—i.e. at a particular value of m, V_c is independent of the value of R_1 and consequently of I_1.

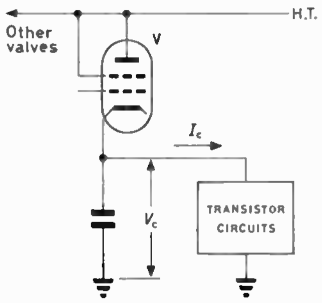

Fig. 1. Basic arrangement

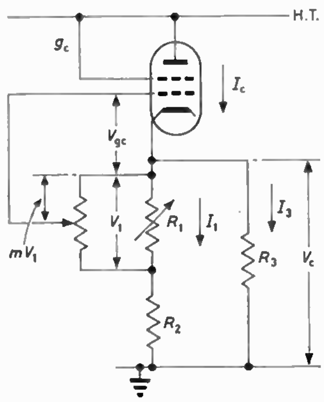

Fig. 2. Stabilizing circuit

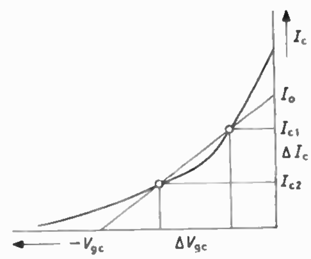

Fig. 3. $I_c - V_{gc}$ characteristic

With the notations of Fig. 2:

$$V_c = I_cR_c, \text{ where } R_c = \frac{(R_1 + R_2)R_3}{R_1 + R_2 + R_3} \quad \ldots\ldots\ldots\ldots (3)$$

$$I_c = I_o - g_cV_{gc}; \text{ where } I_o = I_c \text{ at } V_{gc} = 0\text{V} \quad \ldots\ldots (4)$$

and

$$V_{gc} = mI_1R_1 = mV_c(R_1/(R_1+R_2)) \quad \ldots\ldots\ldots (5)$$

Substituting equations (4) and (5) into equation (3):

$$V_c = (I_o - g_cV_{gc})R_c = I_oR_c - g_cR_cmV_c(R_1/(R_1+R_2)) \quad \ldots\ldots (6)$$

and

$$V_c = \frac{I_oR_c}{1+g_cR_cm\,(R_1/(R_1+R_2))} = I_o\frac{(R_1+R_2)\,R_3}{R_1+R_2+R_3+g_cmR_1R_3}$$
$$= I_o\frac{R_2R_3}{R_2+R_3}\;\frac{1+(R_1/R_2)}{1+(R_1/(R_2+R_3))\,(1+mg_cR_3)} \quad \ldots\ldots (7)$$

From equation (7) it can be readily seen that V_c is constant, if:

$$R_1/R_2 = \frac{R_1}{R_1 + R_2}(1 + mg_cR_3) \quad \ldots\ldots\ldots\ldots (8)$$

Solving equation (8):

$$m = 1/R_2g_c \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (9)$$

and substituting equation (9) into equation (7):

$$V_c = I_o\frac{R_2R_3}{R_2 + R_3} \quad \ldots\ldots\ldots\ldots\ldots\ldots (10)$$

The voltage across R_1:

$$V_1 = V_c - I_1R_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

Equation (9) is the most important conclusion of this investigation, which shows that the value of m necessary to keep V_c constant is indeed independent of R_1.

The design procedure is as follows:

(1) The values of I_o and g_c have to be established from the published data of the valve in question (see Appendix).

(2) From I_o; the required supply voltage V_c; and the constant current need I_3 of the circuit represented by resistor R_3; the value of R_2 can be calculated.

Re-arranging equation (10):

$$R_2 = \frac{V_cR_3}{I_oR_3 - V_c} = \frac{V_c}{I_c-(V_c/R_3)} = \frac{V_c}{I_o - I_3} \quad \ldots\ldots (12)$$

(3) From g_c and R_2, using equation (9), the value of m is obtained.

(4) Combining equations (11) and (12), the voltage across the circuit represented by R_1 can be calculated:

$$V_1 = V_c - I_1\frac{V_c}{I_o - I_3} = V_c\left(1 - \frac{I_1}{I_o - I_3}\right) \ldots (13)$$

As seen, this voltage depends on the voltage V_c and the currents I_o, I_1 and I_3 only. From these, I_1 alone is not constant, but changes between a maximum and a minimum value, depending on the operation of the circuit, as was originally assumed.

It has to be checked, if the minimum value of V_1, corresponding to the maximum value of I_1, is still sufficient to ensure the satisfactory functioning of the circuit. If this is not so, V_c must be increased and if necessary, a corresponding voltage drop introduced into the branch supplying R_3.

Appendix

In conjunction with equation (4), I_o was defined as the value of the cathode current at V_{gc}=0V. In the case of a valve having a curved $I_c - V_{gc}$ characteristic, I_o cannot be measured simply by shorting the grid to the cathode. Assuming that the valve will operate between I_{c1} and I_{c2} (see Fig. 3), the value of I_o is at the intersection of the extended line connecting points I_{c1} and I_{c2} on the actual curve, with the ordinate of the diagram. This discrepancy arises from the approximation implicit in the calculations, that $g_c = \Delta I_c/\Delta V_{gc}$ is constant.

Obviously, the value of g_c obtained in this manner will not be sufficiently accurate if much smaller or much larger currents are considered than those within the range of operation.

Noise Spectrum Measurements on Tunnel Diodes in the Frequency Range 5kc/s to 10Mc/s

By R. A. Giblin*, B.Sc.

Measurements are described of noise in 1mA peak-current germanium tunnel diodes from 5kc/s to 10Mc/s in the bias voltage range 0 to 450mV, with particular reference to the negative conductance region of the characteristic. A comparison method was used which was applicable over the entire frequency range, and the special difficulties encountered in the noise measurements are discussed. The problem of stabilizing the diode is considered in detail, and a circuit is given which enabled the static characteristic and the negative slope conductance to be measured simultaneously. The results indicate that the spectrum obeys a 1/f law up to 10Mc/s for bias voltages above 300mV, and reduces gradually to a flat spectrum below 50mV.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

If the charges carrying currents in a diode move without interactions then the full shot effect is exhibited in each current and the diode can be represented as a noise source by a current generator i in parallel with the small-signal impedance of the diode (Fig. 1).

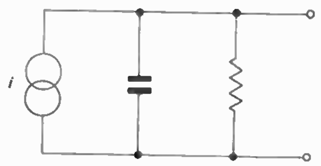

Fig. 1. Noise equivalent circuit of a junction diode or a saturated thermionic diode

$$\overline{i_{(mn)}^2} = 2q\,\Delta f\,\Sigma I_o$$

where $q = 1{\cdot}6 \times 10^{-19}$ coulombs

$\overline{i_{(mn)}^2}$ = mean-square noise current

Δf = bandwith of measuring channel

ΣI_o is the numerical sum of the d.c. in the mechanism. Thus for a saturated thermionic diode $\Sigma I_o = I_a$, the anode current, and for a junction diode $\Sigma I_o = (I + 2I_s)$. It is often convenient to compare the mean square noise current with the shot noise of a single equivalent polarizing current I_{eq}, where

$$\overline{i_{(mn)}^2} = 2q\,\Delta f\,I_{eq}$$

Interaction can either increase or decrease the noise. For instance, if the charges move in groups, q is effectively increased (e.g. the flicker effect in a thermionic diode), while if they affect each other singly the noise decreases (the space-charge effect).

If currents flow in the diode circuit when the bias voltage is reduced to zero the noise can be calculated either as the shot effect of these currents or as the thermal noise in the small-signal resistance (r) of the diode. With a junction diode this condition corresponds to the origin ($I = 0$) and $\overline{i_{(mn)}^2} = 2q\,\Delta f\,2I_s$

and $r = kT/q\,I_s$

giving $\overline{i_{(mn)}^2} = 4kT\Delta f/r$

With a thermionic diode, the anode current I_a under retarding-field conditions is given by

$I_a = B\exp(qV/kT_c)$ (T_c = cathode temperature)

and $1/r_a = q\,I_a/kT_c$

giving $\overline{i_{(mn)}^2} = 2I_a\,q\,\Delta f$

$= 2kT_c\Delta f/r_a$

which corresponds to the thermal noise in r_a at one-half the cathode temperature.

In general the noise increases at low frequencies, i.e. below a few kilocycles per second in thermionic and junction diodes. The excess noise is frequently associated with thin films and point contacts. Reverse-biased junction diodes exhibit considerable excess noise.

The excess noise varies widely from sample to sample, suggesting that surface effects are involved and processing could reduce the noise. Measurements on junction diodes indicate that this noise only exists in the first few kilocycles.

With semiconductor diodes, two exceptional features are encountered: the resistance in the forward direction is low, and the temperature can increase. Measurements of noise in junction diodes are limited to currents of a few hundred microamperes because the levels are so low. For 100μA the equivalent noise resistance is about 100Ω and for 1mA it is only 10Ω; these values are difficult to resolve accurately because they are low compared with the equivalent noise resistance of most noise measuring channels.

Little experimental work has been published on tunnel diode noise in the frequency range[1-4] 5kc/s to 10Mc/s and none which covers measurements made throughout the negative conductance region of the diode characteristic. In this investigation particular attention has been paid to this region as it is the one of greatest interest in amplifier and oscillator applications of tunnel diodes, and also because the 'excess current' mechanism predominates over much of the region. A study of the noise characteristics may help to clarify the nature of the excess current.

Measurement of tunnel diode noise over the whole forward characteristic presents special experimental difficulties for two reasons: the problem of stability, and the problem of the magnitudes involved. Both of these will be considered in more detail.

To make measurements in the negative conductance region the bias circuit must be designed to prevent the diode from switching. (Fig. 2 (a) and (b)). $|R|$ is the modulus of the greatest negative slope resistance, and if $R_L > |R|$ the diode will switch from one positive conductance region to the other, but will not settle at the intermediate point. Therefore $|R|$ is an upper limit to the series resistance of the bias supply, and for a 1mA peak-current diode it is about 100Ω. For higher peak currents, $|R|$ falls rapidly. When the diode is biased at a point in the negative conductance region the circuit is capable of sustained oscillations because the series inductance of the package and leads can form a tuned circuit with the junction capacitance. The negative resistance can cancel the

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losses in this tuned circuit unless the dynamic resistance is less than $|R|$. (Fig. 2(c)). This leads to the a.c. stability condition:

$$\frac{L_s}{C(R_s + R_L)} < |R| \text{ or } R_L + R_s > L_s/C|R|$$

where L_s is the total series inductance, R_s is the resistance of the bulk semiconductor material and C is the junction capacitance.

The complete stability criterion is therefore:

$$|R| > R_L > (L_s/C|R|) - R_s$$

This sets a lower limit to the permissible series resistance,

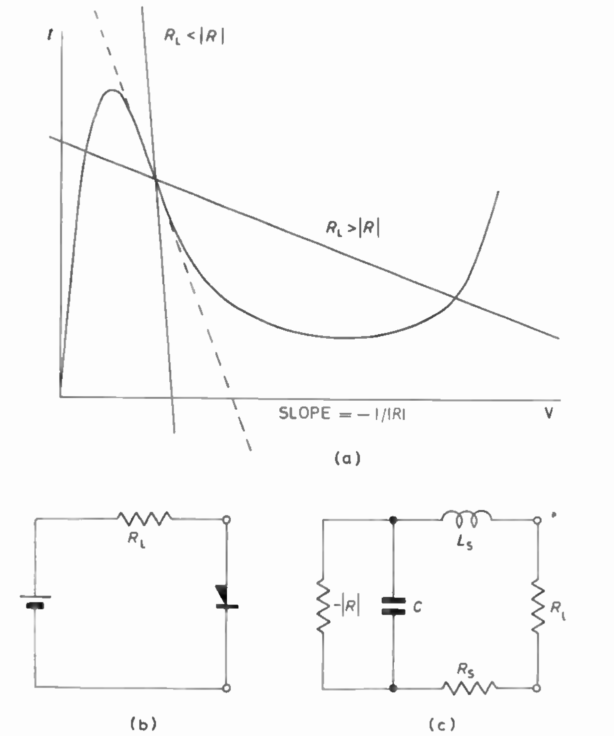


Fig. 2(a). Tunnel diode static characteristic showing two possible load lines
(b). Circuit for (a)
(c). A.C. equivalent circuit of a tunnel diode biased to an operating point in the negative conductance region

and unless the series inductance is kept as low as possible there may be no value of R_L to satisfy both conditions.

The second difficulty arises from the great variations in the diode conductance over the forward characteristic, and the equally great variations in the noise over the frequency range of this investigation. The conductance varies from $+50\text{m}\Omega^{-1}$ to $-10\text{m}\Omega^{-1}$ in the voltage range 0 to 450mV for the diodes used, and the equivalent saturated diode current I_{eq} varies from about 1 000mA at the lowest frequencies to less than 0·5mA at 10Mc/s. In selecting a single experimental technique for the entire series of measurements, substitution methods using standard noise diodes were excluded because of the unusually high anode current required below 1Mc/s. The noise voltages to be measured were very low in both positive conductance regions owing to the shunting effect of the high conductance. Thus from 0 to 25mV the noise voltage to be measured was equivalent to the thermal noise in less than 50Ω, and fell again to 50Ω or less above about 425mV. In the negative conductance region the noise voltage is much higher, and the greatest precision of measurement is possible in the neighbourhood of the maximum negative conductance, which occurs at about 85mV forward bias.

The precision of the measurement in the positive conductance regions is degraded by the presence of a parallel stabilizing resistance, but in the negative conductance region a relatively large noise voltage is produced at the diode terminals because the resistance can be selected to give a small total positive conductance in parallel with the noise current generator.

Theoretical Basis

The method used is a comparison one where the known noise level is that due to the thermal noise in the input circuit and the equivalent noise resistance of the measuring channel (Fig. 3).

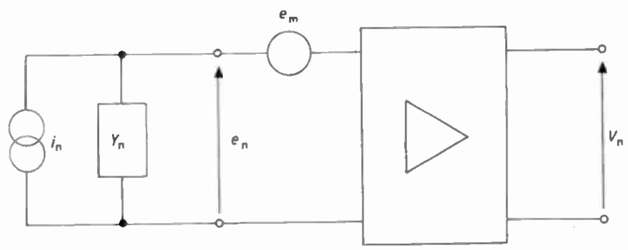


Fig. 3. Equivalent representation of noise measuring channel and input circuit

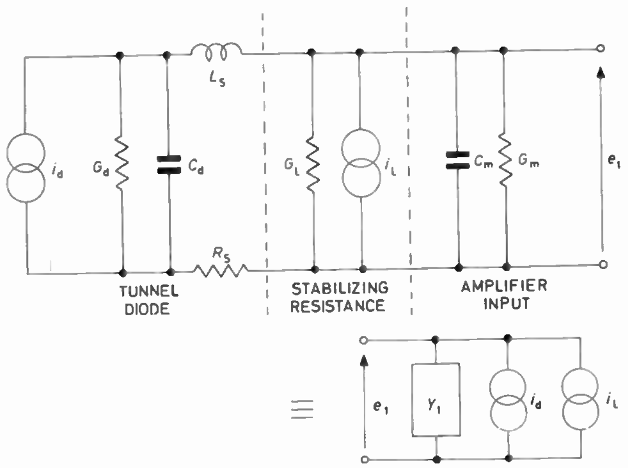


Fig. 4. Detailed input circuit showing the two significant noise sources

Y_n = total input circuit admittance including the amplifier input admittance;

i_n = noise current generator which includes the thermal noise in the stabilizing resistor;

e_m = the noise voltage due to the measuring channel referred to the input circuit.

If the effective noise bandwidth of the channel is Δf c/s, then $e_{m(mn)}{}^2 = 4kTR_m\Delta f$ where R_m is the equivalent noise resistance of the channel.

To obtain I_{eq} for the diode at a particular bias voltage, two measurements are made. The diode is first biased to the required voltage and the output voltage noted, $v_n = v_1$. Then the diode bias is reduced to zero and the output voltage again noted, $v_n = v_2$. Then since the gain of the channel is constant:

$$v_{1(mn)}{}^2/v_{2(mn)}{}^2 = \frac{e_{m(mn)}{}^2 + e_{1(mn)}{}^2}{e_{m(mn)}{}^2 + e_{2(mn)}{}^2}$$

giving:

$$e_{1(mn)}{}^2 = (e_{m(mn)}{}^2 + e_{2(mn)}{}^2)\left[v_{1(mn)}{}^2/v_{2(mn)}{}^2 - \frac{e_{m(mn)}{}^2}{e_{m(mn)}{}^2 + e_{2(mn)}{}^2}\right] \quad \ldots\ldots\ldots\ldots (1)$$

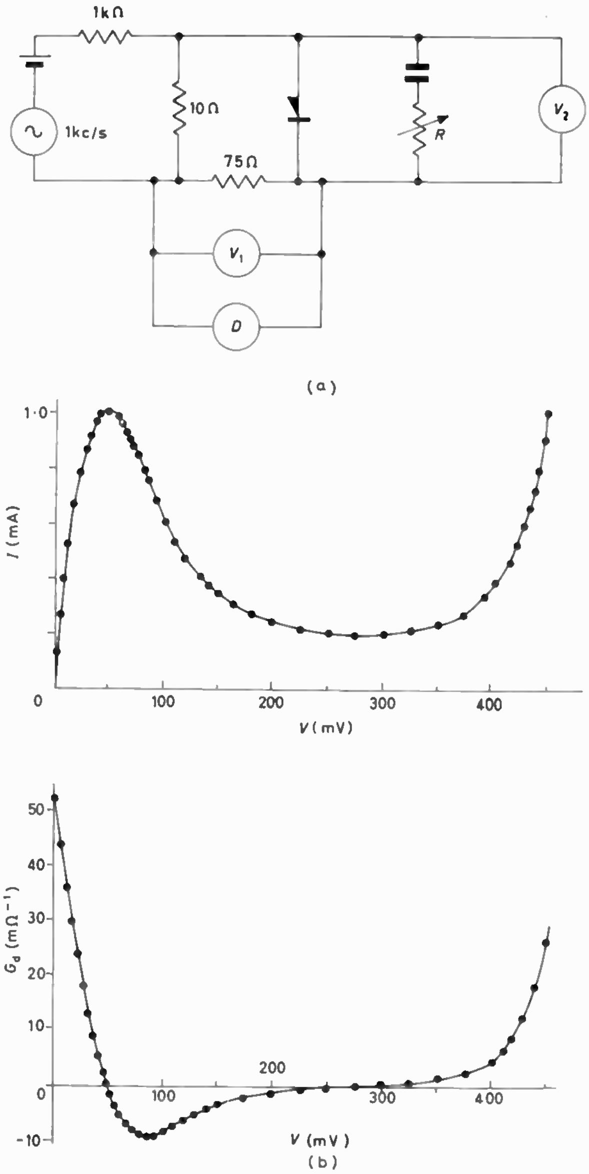


Fig. 5(a). Circuit used for simultaneous measurements of static characteristic and negative slope resistance
(b). Static characteristic and slope conductance

It is convenient to express $e_{1(mn)}{}^2$ in terms of the thermal noise in a resistance R where:

$$e_{(mn)}{}^2 = 4kTR\,\Delta f \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

giving

$$R = (R_m + R_2)\left[v_{1(mn)}{}^2/v_{2(mn)}{}^2 - \frac{R_m}{R_m + R_2}\right]$$

where R_2 is the input circuit resistance for zero bias on the diode.

A second expression for $e_{1(mn)}{}^2$ can be obtained from the detailed input circuit (Fig. 4). i_L is the thermal noise generated in G_L:

$$i_{L(mn)}{}^2 = 4kTG_L\,\Delta f$$

The noise contributions of R_s and G_m are negligible. The equivalent circuit gives

$$e_{1(mn)}{}^2 = \frac{i_{d(mn)}{}^2 + i_{L(mn)}{}^2}{|Y_1|^2}$$

therefore

$$i_{d(mn)}{}^2 = e_{1(mn)}{}^2\,|Y_1|^2 - i_{L(mn)}{}^2 \quad \ldots\ldots\ldots \quad (3)$$

Writing $i_{d(mn)}{}^2 = 2qI_{eq}\Delta f$ and inserting equations (2) and (3) gives

$$I_{eq} = 2kT[R|Y_1|^2 - G_L]/q \text{ amperes.}$$

The channel output is rectified and averaged, so that if all the noise sources have the same amplitude distribution the reading θ is proportional to the r.m.s. total noise voltage. Then:

$$v_{1(mn)}{}^2/v_{2(mn)}{}^2 = (\theta_1/\theta_2)^2 \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

Experimental Arrangements

The measurements were made on JK19B tunnel diodes, which are 1mA peak-current germanium diodes made by S.T.C.

Accurate measurements of the d.c. characteristics and slope conductances were made prior to the noise measurements (Fig. 5(a) and (b)). The circuit enabled simultaneous measurements of voltage, current and negative slope resistance to be made. V_2 measured the diode voltage, V_1 measured the diode current (from the voltage across the 75Ω resistor) and D was a tuned detector which registered a null when R was set to the modulus of the negative slope resistance. The positive conductance regions were measured using a bridge method.

The noise measuring channel is shown in Fig. 6. The stabilizing resistance was a 75Ω composition resistor connected to the diode by short leads to minimize the series inductance. The moving-coil instrument, used as an external meter for the valve-voltmeter, had a sufficiently long time-constant to enable the reading to be averaged by eye.

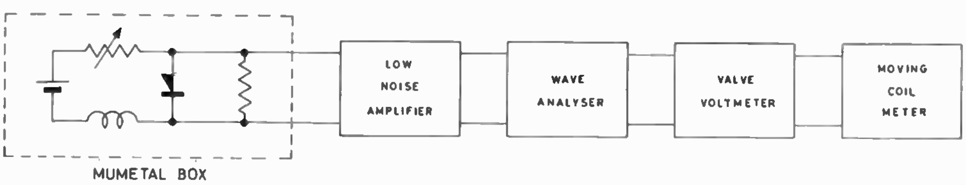
The equivalent noise resistance of the channel was found at the frequency of measurement by connecting a resistance box at the input and noting the intercept of a graph showing the mean-square output voltage against resistance setting.

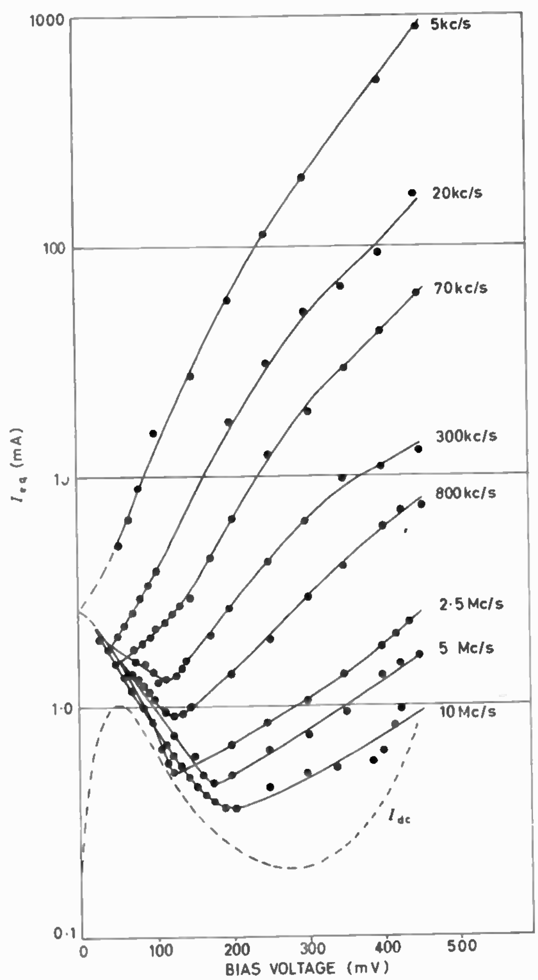
Results and Discussion

The results can be presented either as I_{eq}/bias voltage curves for a particular frequency or as spectra for selected bias values (Fig. 7(a) and (b)).

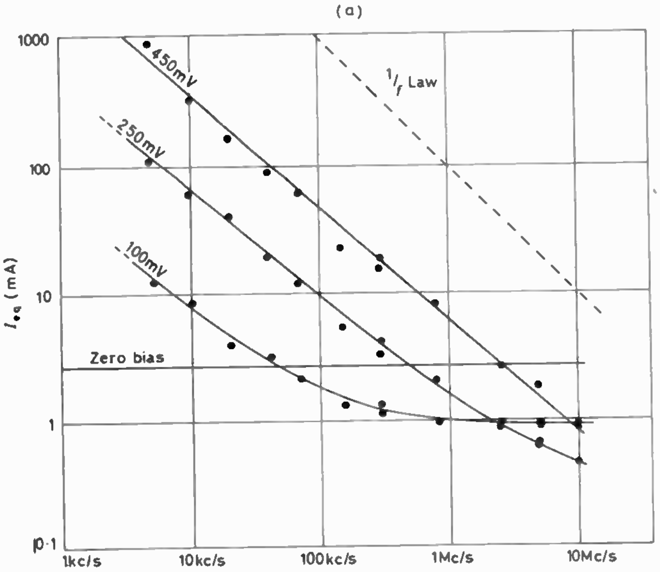
Considering first the I_{eq}/bias voltage curves, for bias voltages $>$ 50mV the noise current is much larger than that corresponding to the full shot noise of the d.c., and increases almost exponentially with bias voltage for the lower frequencies. At the higher frequencies it approximates to the d.c. characteristic, differing most from it in the valley region and below 50mV. It appears that the curves would cross the d.c. characteristic (implying less than full shot noise) above 450mV, but the uncertainty in the measurements is high in this region owing to the very high conductance. Also, it is not certain that the conductance values measured at low frequencies apply at 10Mc/s in this region. The diode is sensitive to ambient temperature changes above about 400mV and these could affect the conductance appreciably.

Fig. 6. Tunnel diode polarizing circuit and block diagram of the noise measuring channel





(a)



(b)

Fig. 7(a). Ieq/bias voltage curves for selected frequencies

(b). Noise spectra for selected bias voltages

The curves from 20kc/s to 10Mc/s converge at 50mV and below to a value of about 2·5mA at zero bias. This can be obtained theoretically by using the fact that the noise is wholly thermal at zero bias. Thus:

$$i_{d(mn)}^2 = 4kTG_o\,\Delta f = 2qI_{eq}\,\Delta f$$

giving $I_{eq} = 2kTG_o/q$, where G_o = the zero-bias conductance = $52\text{m}\Omega^{-1}$ for the diodes used. This gives I_{eq} = 2·6mA. The I_{eq} curve would not be expected to follow I_{dc} in this part of the characteristic since I_{dc} is the resultant of two opposite, uncorrelated currents (the Esaki and Zener currents) so that I_{eq} reflects the sum of their mean-square values and I_{dc} is their numerical difference.

The spectra show that I_{eq} (and therefore $i_{d(mn)}^2$) follows a $1/f$ law up to at least 10Mc/s for 450mV bias, and for decreasing bias voltages they show a grading towards the necessarily flat spectrum at zero bias. The convergence of the I_{eq}/bias curves below 50mV implies that the spectrum is flat before zero bias is reached. This agrees with measurements made by Agouridis and van Vliet[4].

Acknowledgements

The author expresses his gratitude to the Principal and Governing Body of Woolwich Polytechnic for providing the research facilities used in this investigation, and his indebtedness to Mr. K. B. Reed for criticisms and suggestions made during the course of the work.

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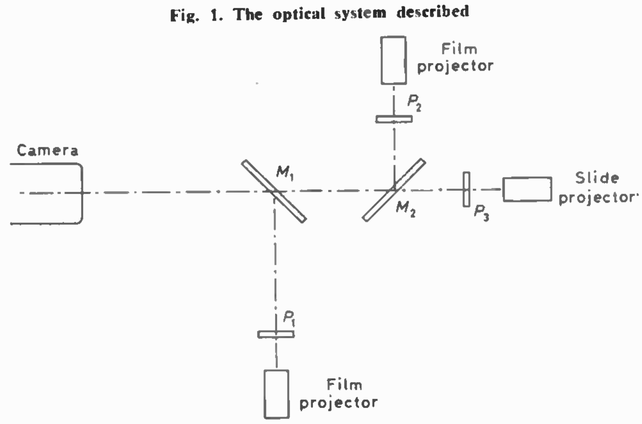
An Optical System Employing a Beam Splitting or Combining Arrangement*

An optical system has been developed which is suitable for use in conjunction with a colour camera for enabling scenes from any one of a slide projector and two film projectors to be projected to the camera. The system (Fig. 1) comprises two partially reflecting members M_1 and M_2 disposed as shown on the axis of the camera and inclined in reverse senses to this axis, with the three projectors arranged as shown, the path lengths from the respective projectors to the camera being the same.

Preferably each partially reflecting member is constituted by a pattern of totally reflecting material, such as in spaced parallel strips, on a transparent substrate, and the members M_1 and M_2 can be formed on interfaces of a prism assembly.

The optical system described has the effect of introducing polarization into the beams of light from the projectors to the camera. Light from the film projectors is polarized to a greater extent in the vertical than in the horizontal direction, whereas light from the slide projector is polarized to a greater extent in the horizontal than in the vertical direction. In order to compensate for the different degrees of polarization the optical system also includes polarizing members, such as pieces of polaroid P_1, P_2 and P_3 as shown each of which is orientated to compensate for the excess polarization in the respective light beams.

Fig. 1. The optical system described



* *A communication from EMI Ltd.*

A Single Stage Amplifier for Repetitive Analogue Computation

By R. E. King*, Ph.D., A.M.I.E.E.

A simple Bootstrap d.c. amplifier having an open loop gain of approximately 2000 is described and an analysis of its frequency response presented. The amplifier is particularly suitable for repetitive analogue computation over the audio range.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

THE need of a simple, reliable, directly-coupled amplifier having a high gain over the operating region has led to the development of a single stage amplifier in which Bootstrap feedback has been employed. The advantages of a single stage configuration with its inherent stability coupled with positive feedback in the form of Bootstrap feedback (which cannot cause instability) yield a compact and efficient unit which is particularly suitable for repetitive analogue computation. Twelve such amplifiers were

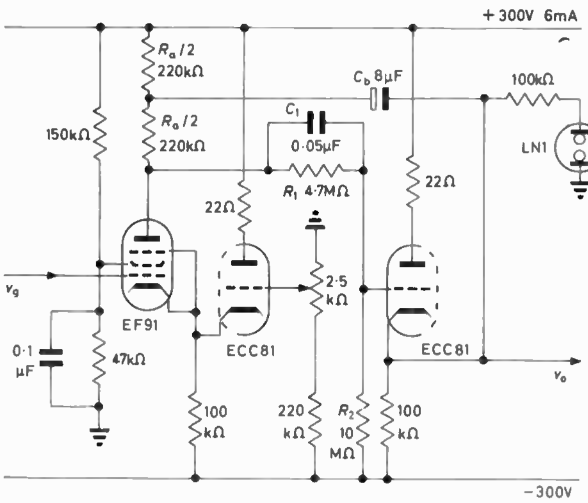


Fig. 1. The Bootstrap amplifier

used in a repetitive analogue computer originally developed for setting up an analogue of a magnetic amplifier[1] and are now used as an instructional unit. Low power consumption and potential compactness of the complete unit leads to the portability necessary for demonstration purposes.

The Amplifier and Circuit Details

Fig. 1 shows the complete Bootstrap amplifier with component values to give an overall bandwidth of from 6c/s to 5kc/s and a gain of 1 800.

The first stage is basically a long-tail pair comprising a starved EF91 coupled with half an ECC81. The starved pentode gives a gain of approximately 300 at zero frequency with an anode load of $\sim$0·5MΩ and a decoupled screen voltage of 50V. With a total cathode (or tail) current of 3mA this long-tail pairing of an EF91 and an ECC81 provides reasonable drift compensation[2] against variations in heater supplies which need not consequently be stabilized. An EF91 is chosen because of its high gain under starved operation. Direct coupling to the output cathode-follower (the other half of the ECC81) is via resistors R_1 and R_2 where R_1 is by-passed by a large 'phase-advancing' capacitor C_1 to ensure unity transmission down to low frequencies. The overall zero frequency gain of the amplifier as such is about 200. Being a single stage amplifier the problems of instability and correct d.c. bias to following stages do not arise. Zero setting is achieved by variation of the grid potential on the triode of the long-tail pair. Stability is considered in detail in the next section.

To increase the gain of an amplifier without recourse to additional stages positive feedback may be employed. Such techniques have been extensively used, a well-known example being the Schmidt trigger circuit. Unfortunately indiscriminate use of positive feedback in amplifiers almost invariably leads to instability if the feedback signal is increased sufficiently. Thus even when such amplifiers are adjusted to be stable and have a high gain, ageing of components and valves may cause large changes in gain and even instability and eventual 'backlash' or hysteresis. Periodic gain adjustment is consequently essential in such cases.

In the Bootstrap positive-feedback configuration however, this situation cannot arise. Consider the mid-frequency range when the inter-stage coupling network (R_1, C_1 and R_2) has unity transmission and the Bootstrap capacitor C_b has negligible reactance. Then the anode current through the pentode is:

$$i_{ap} = (1 - A)\,(2/R_a)\,v_a = (v_a + \mu_p v_g)\,(1/r_{ap})$$

where A is the gain of the output cathode-follower and μ_p and r_{ap} are the pentode parameters under starved operation. The mid-frequency gain of the amplifier is therefore:

$$G(mf) = v_o/v_g = \frac{-\mu_p A}{1 + (1 - A)\,(2r_{ap}/R_a)}$$

Since $A < 1$, $G(mf)$ is always finite, approaching μ_p and is stable whatever the values of the circuit and valve parameters. A more complete analysis indicating the be-

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haviour of the circuit at the low end of the frequency spectrum is presented in the next section.

Bootstrap feedback may be applied over a certain frequency band as in this application[3] by the use of a feedback capacitor C_b or when the full range down to zero frequency is required by replacing the capacitor C_b with a low current neon stabilizer[4].

The amplifier is capable of a maximum output voltage swing of ± 70V at ± 2mA, overload being indicated by a neon LN1.

Low Frequency Performance of the Bootstrap Amplifier

For the sake of simplicity the stray capacitances across the pentode anode load and the loading effect of the inter-

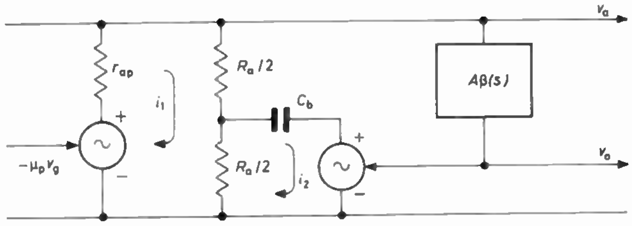

Fig. 2. Equivalent circuit of amplifier

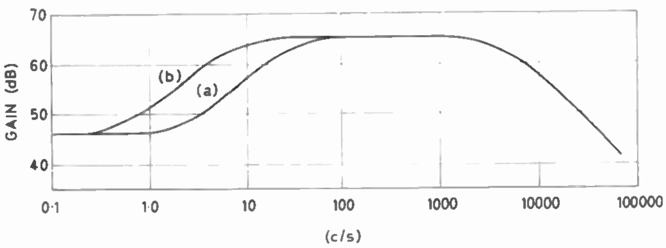

Fig. 3. Frequency response curves
(a) $C_b = 0{\cdot}5\mu F$
(b) $C_b = 8\mu F$

stage coupling network on the pentode input stage are neglected. The small signal equivalent voltage generator circuit is shown in Fig. 2. Here $\beta(s)$ is the inter-stage coupling network transfer function given by:

$$\beta(s) = \frac{\alpha(1 + T_1 s)}{1 + \alpha T_1 s}$$

where:

$\alpha = R_2/(R_1 + R_2)$. $T_1 = R_1 C_1$ and s is the Laplace operator.

The mesh currents i_1 and i_2 can be written in matrix form:

$$\begin{bmatrix} -\mu_p v_g \\ -\beta A v_a \end{bmatrix} = \begin{bmatrix} r_{ap} + R_a & -R_a/2 \\ -R_a/2 & R_a/2 + 1/C_b s \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix}$$

Now:

$$-v_a = \mu_p v_g + i_1 r_{ap}$$

which on substituting above gives:

$$\mu_p v_g \begin{bmatrix} -1 \\ A \end{bmatrix} = \begin{bmatrix} r_{ap} + R_a & -R_a/2 \\ -R_a/2 - A r_{ap} & R_a/2 + 1/C_b s \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix}$$

whence:

$$i_1 = \mu_p v_g \begin{vmatrix} -1 & -R_a/2 \\ A & R_a/2 + 1/C_b s \end{vmatrix} \Delta s$$

$$= -\mu_p v_g \{R_a(1 - A)/2 + (1/C_b s)\}/\Delta(s)$$

and

$$\Delta(s) = \begin{vmatrix} r_{ap} + R_a & -R_a/2 \\ r_{ap}(1 - A) + R_a/2 & 1/C_b s \end{vmatrix}$$

$$= \{(R_a/2) + r_{ap}(1 - A)\}(R_a/2) + \frac{r_{ap} + R_a}{C_b s}$$

$$\simeq \frac{(r_{ap} + R_a)}{C_b s}(1 + T_2 s)$$

where

$$T_2 = R_a^2 C_b / 4(r_{ap} + R_a)$$

The pentode anode voltage v_a is thus:

$$v_a = -\mu_p v_g [1 - (r_{ap}/\Delta(s))\{(R_a/2)(1 - A) + (1/C_b s)\}]$$

and

since $v_o = A\beta(s)v_a$, the overall amplifier gain is:

$G(s) = v_o/v_g =$

$$\frac{-\mu_p A\beta(s)}{\Delta(s)}[\Delta(s) - r_{ap}\{(R_a/2)(1 - A) + (1/C_b s)\}]$$

which on simplifying yields:

$$G(s) = G(0)\left[\frac{(1 + T_1 s)(1 + T_4 s)}{(1 + \alpha T_1 s)(1 + T_2 s)}\right]$$

where

$$G(0) = \text{zero frequency gain}$$

$$= \frac{-\mu_p R_a}{r_{ap} + R_a}\alpha A$$

$$T_4 = \frac{T_2 - \gamma T_3}{1 - \gamma}$$

$$\gamma = \frac{r_{ap}}{r_{ap} + R_a}$$

and

$$T_3 = (R_a/2)(1 - A)C_b$$

The mid-frequency gain can thus be derived by allowing $s \to \infty$, i.e.

$$G(mf) = G(0)T_4/\alpha T_2$$

The most significant angular break frequencies are $1/T_1$ at which the gain of the amplifier begins to rise from its zero frequency value $G(0)$ and $1/T_2$ the angular frequency at which the mid-frequency gain $G(mf)$ begins to drop towards the zero frequency gain $G(0)$.

Conclusions

A basically simple and extremely reliable low power consumption single stage Bootstrap amplifier having adequate gain for normal repetitive operation has been described. The gain of the particular configuration given is 1 800 and the effective bandwidth can be designed as required. The high frequency cut-off frequency may be extended by reduction of the anode load resistance of the pentode input stage making high speed repetitive computation possible.

Acknowledgments

The author wishes to acknowledge the help of Messrs. J. T. Connolly and H. M. Toner in constructing and testing the amplifiers as part of an undergraduate project.

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A Transistor Linear Sawtooth Generator with Negligible Step

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

A linear sawtooth generator of well defined slope and negligible initial step is described. The sawtooth is synchronized to an external trigger. By the design of a suitable switching circuit the delay from the application of the trigger pulse to the commencement of the sawtooth is kept to a low value. Positive and negative going sweeps are obtainable.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

FOR some applications a source of inconvenience with many sawtooth generators is that before commencement of the sweep there is an initial step. If the d.c. level of the sweep at any instant of time is important then it is possible to offset the sweep, by a potentiometer arrangement, by an amount equal to the initial step. In complex equipment the setting up procedure should be as simple as possible and the use of a potentiometer is unsatisfactory. Furthermore the effect of temperature change on the potentiometer can reduce the accuracy of the circuit considerably.

In the circuit to be described the initial offset has a measured value of 20mV and the potential from which the sweep commences is well defined.

Principle of Operation

The initial step may be due to the necessity of turning on an amplifier before the sweep can commence or conversely it may be due to the use of a 'bottomed' active element which must come out of saturation before the sweep can commence. A further reason may be due to the use of a high output impedance amplifier. This could be remedied if the circuit was designed around a relatively low output impedance amplifier in which the active elements always conducted in a linear mode yet so arranged that the output was at a defined potential (e.g. earth) before application of the input trigger.

Fig. 1 shows an idealized arrangement. Normally the switches S_1 and S_2 are closed, hence the output is clamped at earth. On application of the input trigger the switches are opened and the sweep commences as a Miller integrator. Three points are to be satisfied for the system to be accurate.

(1) If the switches are not perfect then their offsets should cancel.

(2) There should be as small a delay as possible in opening the switches after application of the input trigger.

(3) The amplifier gain and bandwidth should be such as to allow the slope to be determined by the CR time-constant and V_1.

Practical Arrangements

Fig. 2 shows the circuit of a sweep generator in which the amplifier would always be conducting and the offset very small. Initially neglect the presence of MR_1. With switch S_1 closed R_1 is connected to the supply voltage V and the circuit acts as a Miller integrator. R_2 is simply an additional load on the amplifier. When S_1 is opened the output returns to earth, due to the balanced input stage used, on an exponential of form:

$$V_o = V_I\,e^{-t/CR} \quad \dots\dots\dots\dots\dots\dots \quad (1)$$

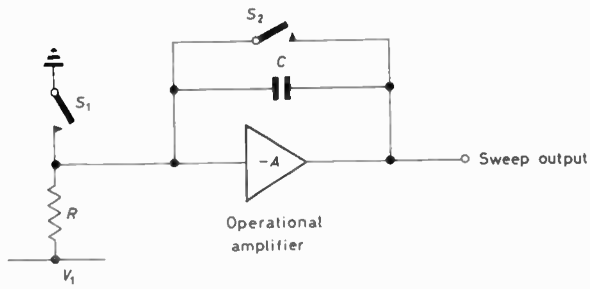


Fig. 1. Idealized arrangement for setting the starting level of a sweep generator

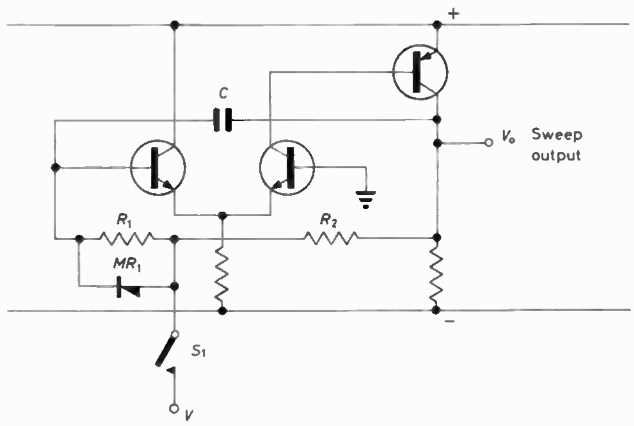


Fig. 2. Practical circuit for a sweep generator in which the amplifier is always conducting and the starting level defined at earth potential

where V_o = output voltage at any instant

V_I = output voltage due to Miller action at the instant of opening S_1.

C = feedback capacitor

$R = R_1 + R_2$

Two main sources of error are apparent in the above.

(1) Any offset voltage due to S_1 would affect the rate of change of output voltage slightly.

(2) Examination of equation (1) shows that ideally the output never reaches zero volts. However if CR was made equal to $\left[\frac{\text{Reset Time}}{8}\right]$ the error becomes approximately 0·035 per cent of V_I which would be satisfactory for many cases.

* *Ferranti Ltd.*

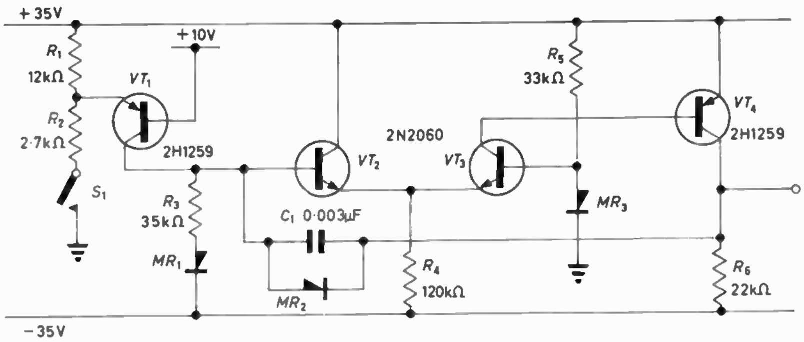


Fig. 3. Second form of practical sweep generator in which the amplifier is always conducting and the starting level defined at earth potential

It will be observed that the transistor switch VT_1 is not saturated, hence the delay in turning off VT_1 will be less than in circuits using saturated switches.

Design Points

In order to satisfy the criterion that the sweep shall commence from earth potential then the voltage drop across MR_3 plus the base emitter diode drop of VT_3 shall balance the voltage drop across MR_2 plus the base emitter diode of VT_2.

A further limitation of the arrangement shown in Fig. 2 is that R_1 appears in both the charging and resetting equations of the sweep generator which imposes a limitation on the value of the voltage V. See Appendix (2) for justification of this statement.

However R_1 can be eliminated from the resetting time-constant by the addition of a further diode and the voltage to which R_1 is switched can be selected at will. See MR_1 in Fig. 2. Fig. 2 assumes a positive going sweep, hence V is a negative voltage. For a negative going sweep MR_1 would be reversed and V would be a positive voltage.

A second solution to the basic problem is shown in Fig. 3. This will be considered in more detail since experimental data is available. Fig. 3 is for a positive going sweep.

With S_1 open current flows through R_1 and VT_1 and attempts to raise the virtual earth (base of VT_2) to +10V. If this occurred the output transistor VT_4 would turn off and the output would drop towards −35V. Clearly this is impossible due to MR_2. The actual mechanism is that MR_2 represents 100 per cent negative feedback around a linear amplifier. The base of VT_3 is a diode voltage drop above earth and with the balanced circuit used the base of VT_2 is at the same potential. The voltage drop across MR_2 brings the output back to earth potential. The accuracy of this balancing arrangement will be considered in a little more detail in the next section.

When S_1 is closed the junction of R_2 and R_1 falls below +10V thus turning off VT_1. Hence the integrator runs up in normal Miller fashion until S_1 is opened again and the sweep returns to earth potential.

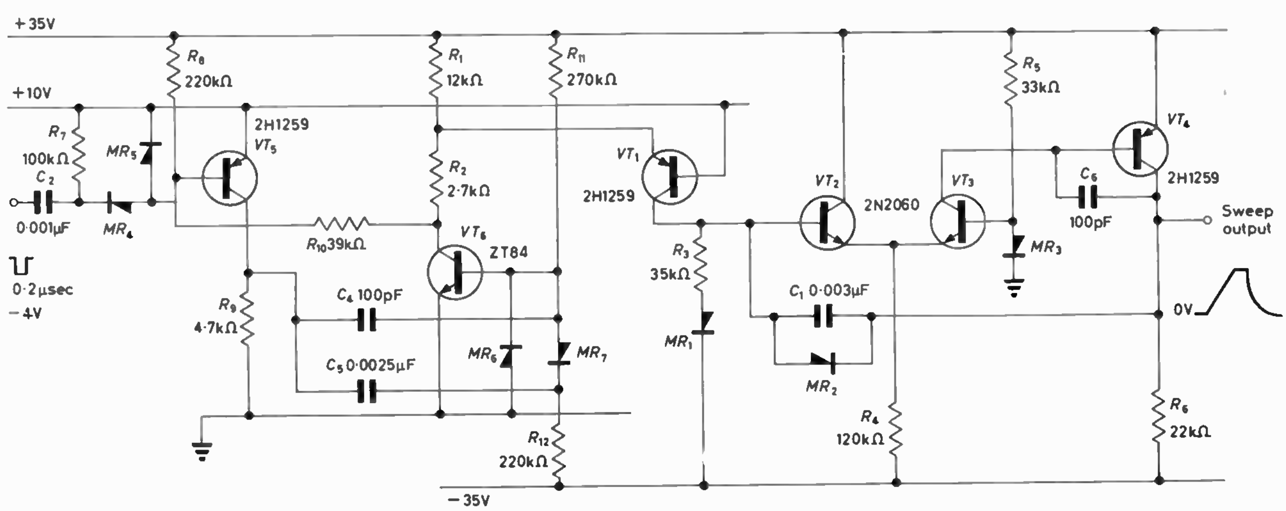
Several manufacturers now produce matched transistors in one heading and the voltage drop across the two base emitter junctions can be made equal to within 5mV if each half has the same current. The change of differential voltage with temperature is in the order of $10\mu V/°C$. Similarly matched diodes are available. In fact MR_1, MR_2 and MR_3 are three diodes from a matched quad, all in one heading. The differential offset voltage of the type used, has a maximum value of 10mV, over the temperature range −55°C to 100°C assuming equal currents through the diodes. Hence using such components the initial offset of the sweep with respect to earth may be as low as 15mV.

In Fig. 3, MR_1, MR_2 and MR_3 each have approximately 1mA flowing through them before commencement of the sweep, determined as follows. When S_1 is open, the current through $VT_1 = (35-10)/12000 = 2mA$. The base of VT_2 is approximately at earth potential, hence the current through $MR_1 = 1mA$. The rest of the current from VT_1 must go through MR_2 (neglecting base current). Hence the current through MR_2 is also approximately 1mA. The current through MR_3 may be controlled by R_5. Hence for approximately 1mA, $R_5 = 33k\Omega$.

The base of VT_2 sits at a diode voltage drop above earth due to the feedback action and MR_3. Now even though the differential offset of MR_2 and MR_3 can be specified accurately the absolute voltage drop cannot. Hence it would be difficult to define exactly the voltage drop across R_3 which determines the slope. By the addition of MR_1 this unknown is removed.

To replace S_1 a monostable multivibrator may be used. As an example assume that a positive going sweep was

Fig. 4. Complete positive going sweep generator



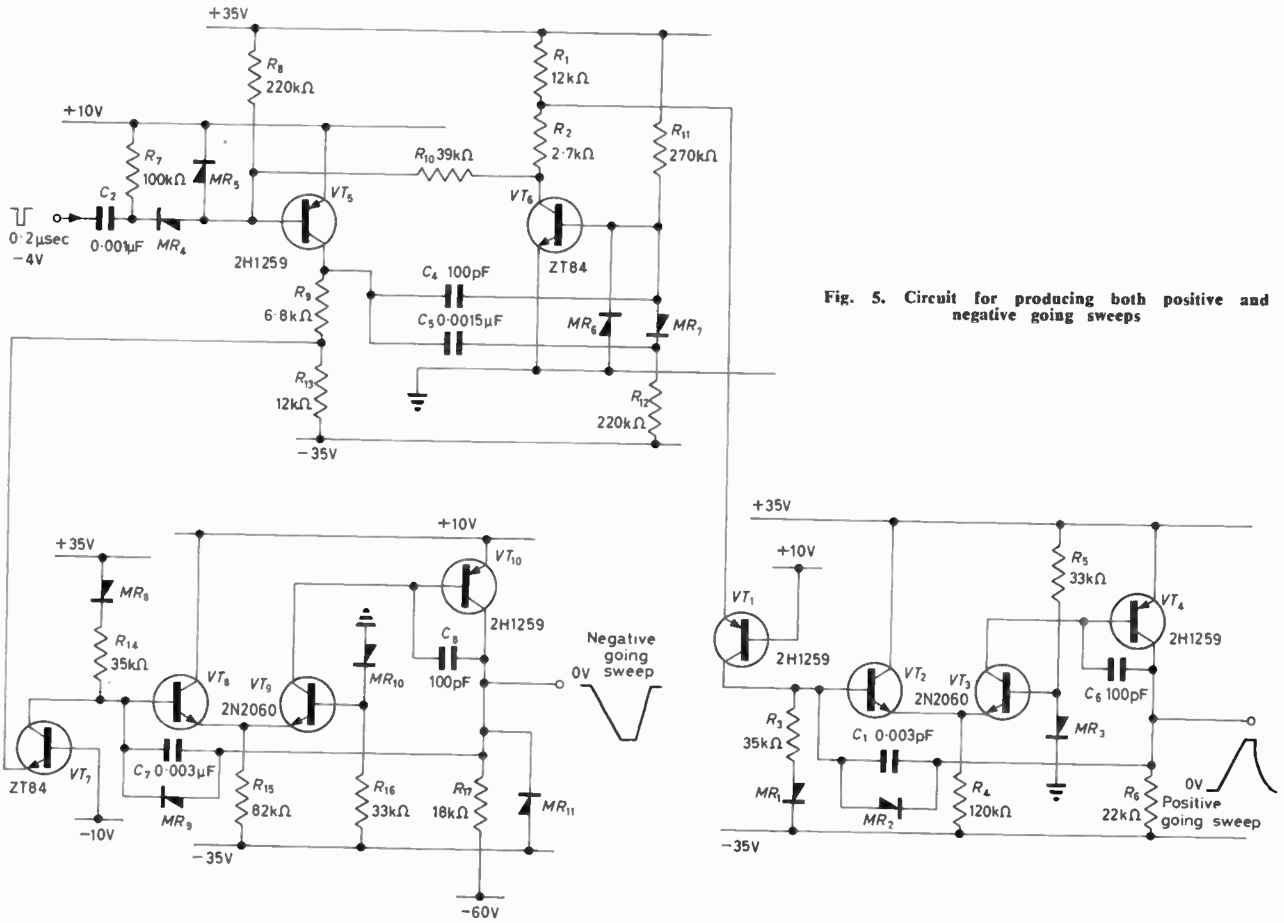


Fig. 5. Circuit for producing both positive and negative going sweeps

required to be produced from a negative input trigger. Fig. 4 shows the circuit used. VT_5 and VT_6 and their associated circuits constitute the monostable multivibrator in place of S_1 in Fig. 3. C_6 is needed to stabilize the amplifier.

Now transistors may be turned on faster than they can be turned off, hence the monostable driving the transistor switch VT_1 was designed around a circuit in which both transistors were normally off. In this way the delay from the application of the trigger to the commencement of the sweep was kept as low as possible. The period of the monostable was calculated to be such as to allow the sweep to run up for the desired time. After which the monostable returns to its quiescent condition, thus allowing the sweep to reset back to earth potential before the application of the next trigger.

Operation of Monostable in Fig. 4

On the application of a negative input trigger VT_5 turns on and the positive going edge produced at its collector is passed through C_5 and turns off MR_7. Hence VT_6 turns on due to current flow through R_{11}. C_4 simply helps to speed up the turn on of VT_6 in allowing the positive edge from VT_5 collector to bypass MR_7 at the instant of turn on of VT_5. VT_5 and VT_6 stay on for a time determined by C_5R_{12}. When VT_6 turns on the junction of R_1 and R_2 falls below +10V thus turning off VT_1. Hence the integrator runs up in normal Miller fashion until the monostable period is complete on which VT_1 conducts again and the sweep returns to earth potential.

Results

As an example the following conditions were designed for.

Assume the sweep was required to change by 30V in 90μsec at a repetition frequency of 3kc/s (333μsec period).

Therefore $(dv/dt) = \frac{1}{3}\text{V}/\mu\text{sec}$

$\therefore\ i = C_1\,(dv/dt)$

$$\therefore\ 35/R_3 = \frac{C_1 \times 10^6}{3}$$

Convenient values of C_1 and R_3 are:

$C_1 = 3000\text{pF}$

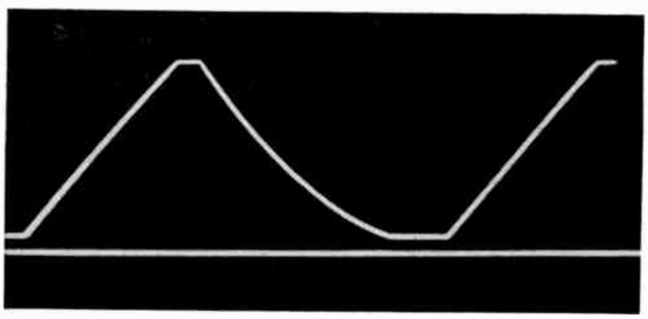
$R_3 = \ \ 35\text{k}\Omega$

Precision components were used.

The monostable was designed for a period greater than 90μsec but short enough to allow it to recover before application of the next trigger.

The measured offset voltage was 20mV above earth which was slightly greater than predicted. The excess offset being due most likely to the fact that VT_2 and VT_3 did

Fig. 6. Positive sweep



not have equal currents through them. This could be remedied by suitable redesign of the amplifier.

The delay from commencement of the sweep after application of the trigger was measured to be 50nsec.

Measurements on sweep linearity and slope gave results consistently better than 0·6 per cent over a temperature range from $-25°$C to $+70°$C.

Negative Going Sweep

Fig. 4 may be extended to give a negative going sweep starting from earth potential by the addition of a further amplifier and associated components. The reversal of sweep polarity requires the amendments indicated in Fig. 5.

Figs. 6 and 7 show the sweep waveforms obtained and Fig. 8 shows the commencement of the negative sweep with respect to a 0·2μsec input trigger.

Further Development

Since the main body of work connected with this article was completed it has been shown that it is possible to make the sweep circuit described self-gating. This had a two-fold effect; first, a saving in components and secondly a speeding up of the commencement of the sweep with respect to the trigger. However, the basic idea remained unaltered.

Acknowledgments

The author wishes to thank Mr. B. E. Pitches whose original suggestions helped to make the system possible, Mr. J. W. Naples for valuable assistance with the experimental work and Ferranti Ltd for permission to publish this article.

Appendix

(1) Verification of Equation (1)

The system is as shown in Fig. 9.

Assume that at the instant of opening the switch S the output is at a voltage V_I. Then for an amplifier of sufficiently high gain:

$$V_o/R = -C\,(dV_o/dt)$$

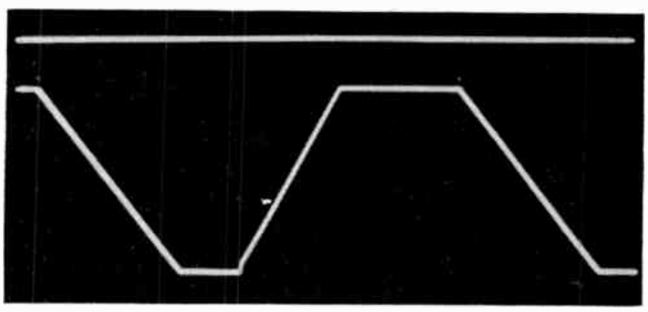
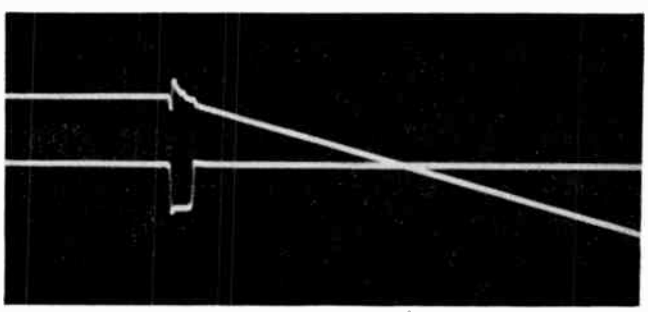


Fig. 7 (above). Negative sweep

Fig. 8 (below). Start of negative going sweep with respect to a 0·2μsec input trigger



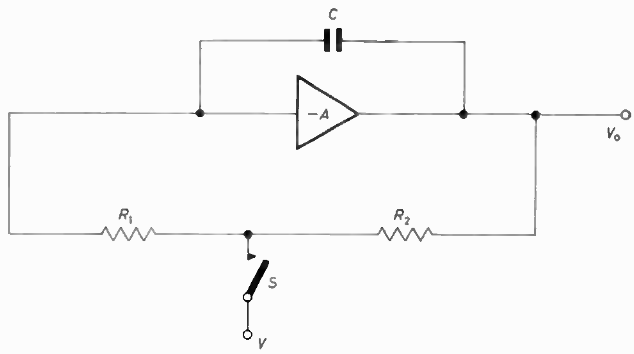


Fig. 9. Simplified form of Fig. 2

where $R = R_1 + R_2$

The Laplace transform of $V_o = V_o'(S)$

The Laplace transform of $dV_o/dt = SV_o'(S) - V_I$

$$\therefore\ V_o'(S)/R = -C\,(SV_o'(S) - V_I)$$

$$V_o'(S) = \frac{V_I}{(1/CR) + S}$$

Taking inverse transforms:

$$V_o(t) = V_I\, e^{-t/CR}$$

(2) Analysis of Fig. 2 for a Typical Case

As in the main text assume a sweep rate of $\frac{1}{3}$V/μsec was desired to last for 90μsec with a repetition frequency of 3kc/s.

Hence reset time = 243μsec.

For approximately 0·035 per cent error, $C(R_1 + R_2) = \frac{\text{Reset Time}}{8}$

$$C(R_1 + R_2) = 30\mu sec \quad \ldots\ldots\ldots\ldots \quad (2)$$

Charging equation is:

$$dV_o/dt = -V/CR_1$$

$$10^6/3 = -V/CR_1$$

$$CR_1 = -3\text{V}\ 10^{-6} \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

Assuming $R_1 = R_2$ equation (2) gives:

$$2CR_1 = 30\mu\text{sec}$$

$$CR_1 = 15\mu\text{sec} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Combining equations (3) and (4):

$$V = -5\text{V}$$

This is rather a low voltage and any offset due to the switch S_1 in Fig. 2 would cause some error.

Now consider Fig. 2 with the addition of MR_1. The reset time-constant becomes R_2C assuming the forward impedance of MR_1 to be small compared with R_2:

$$R_2C = 30 \times 10^{-6} \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

and as before:

$$dV_o/dt = 10^6/3 = -V/CR_1 \quad \ldots\ldots\ldots\ldots \quad (6)$$

It is now possible to specify the voltage V without reference to both equations.

Say $V = -35$V

$CR_1 = 105 \times 10^{-6}$

Say $R_1 = 35\text{k}\Omega$

$C = 3000$pF

and $R_2 = \dfrac{30 \times 10^{-6}}{3000 \times 10^{-12}}$

$= 10\text{k}\Omega$

A Linear Frequency to Voltage Convertor

By W. P. O'Grady*

A semiconductor circuit which converts frequency to voltage in a linear fashion is discussed. Methods for improving the phase characteristic between input frequency and output voltage are considered as well as ways of reducing the ripple voltage superimposed upon the output voltage.

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

MANY physical phenomena are best observed by the change in frequency they produce and since this change is often required in the form of a varying voltage the need for a linear frequency to voltage convertor arises. A non-linear convertor has been described by Earnshaw[1] and he has suggested a linear circuit using valves. The transistor replacement of the linear circuit has been described briefly by Burton and Willis[2]. The basic circuit is shown in Fig. 1.

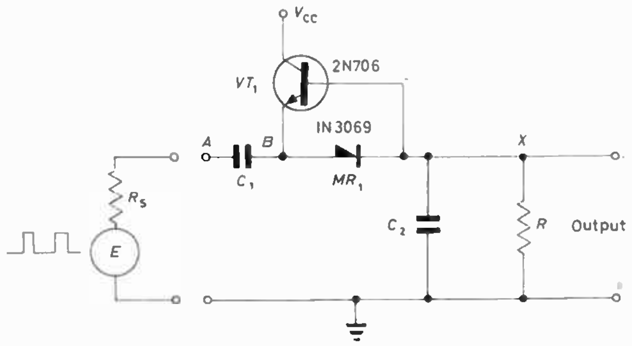


Fig. 1. Basic frequency to voltage convertor

If a frequency f in the form of a pulse train is applied at the input terminals then the voltage at the output is given by the equation:

$$V = EfRC_1$$

Basic Circuit and Theory

Consider a frequency in the form of a pulse train applied at the input terminals of Fig. 1. When the point A is taken positive with respect to earth C_1 and C_2 are charged in series through MR_1 while VT_1 is reversed biased and the potential at X becomes $EC_1/(C_1+C_2)$ when the forward conduction voltage of the diode is neglected. If R is considered as an infinite resistance then the charge on C_2 does not leak away. When A is returned to earth VT_1 becomes forward biased and C_1 discharges through it to high tension and charges up in the opposite direction until the potential at B becomes equal to that at X when neglecting the 'cut on' voltage of the transistor. The condition that now exists is that shown in Fig. 2 where the voltage on C_2, $EC_1/(C_1+C_2)$, is equal and opposite to that on C_1

Since these voltages balance one another out any further voltage inserted into the circuit at A will be divided in the same proportions as previously between C_1 and C_2 and will add linearly to the voltages already across them. Thus when A is again raised to the potential E the potential at X becomes $2EC_1/(C_1+C_2)$ and the voltage across C_1 becomes $EC_1/(C_1+C_2) - EC_2/(E_1+C_2)$. However as before when A is returned to earth the potential at B becomes equal to that at X after C_1 has discharged. Thus for each pulse applied to the input a voltage $EC_1/(C_1+C_2)$ is added to C_2. If R is now considered as being of finite value then when a pulse is applied to A the voltage waveform at X is given by the equation:

$$V_x = V_{op} \exp\left[-\frac{t}{R(C_1+C_2)}\right]$$

(See Appendix (1))

and this continues to be the equation of the waveform when A is returned to earth again as this operation places C_1 in parallel with C_2 through MR_1 provided that the discharge time of C_1 through VT_1 is a lot less than the repetition period of the pulse chain. Thus in one cycle

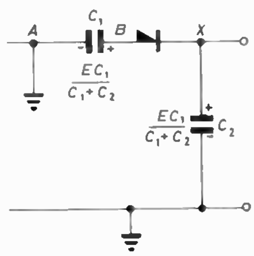


Fig. 2. Circuit condition at beginning of second input pulse

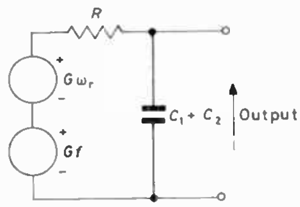


Fig. 3. Basic convertor together with RC filter

the voltage across C_2 will decrease by an amount

$$V_{op}\left[1-\exp\left[-\frac{T}{R(C_1+C_2)}\right]\right]$$

When this value is equal to the voltage added during each cycle the output reaches an equilibrium level, i.e., when:

$$V_{op}\left\{1-\exp\left[-\frac{T}{R(C_1+C_2)}\right]\right\} = \frac{EC_1}{C_1+C_2}$$

$$\text{i.e. } V_{op} = \frac{EC_1/(C_1+C_2)}{1-\exp\left[-\frac{T}{R(C_1+C_2)}\right]}$$

SYMBOLS

V_{cc}	=	Supply voltage
E	=	Amplitude of input pulses
V	=	Average value of output voltage
V_{op}	=	Peak value of output voltage
f	=	Repetition frequency of input pulses
T_p	=	Time-constant of convertor
V_r	=	Peak value of ripple voltage on output waveform
G	=	Gain of convertor
V_{be}	=	'Cut on' voltage of transistor VT_1
V_D	=	Foreward conduction voltage of diode MR_1
V_z	=	Zener voltage
V_{sat}	=	Saturation voltage of transistor VT_2
ω_r	=	Angular frequency of the ripple voltage

* *Cambridge University.*

The d.c. component of the output is the average value of the waveform shown in Fig. 1.

i.e. $V = 1/T \int_0^T V_{op} \exp\left[-(t/T_p)\,dt\right]$ where $R(C_1+C_2) = T_p$

i.e. $V = (V_{op}T_p/T)\,(1 - \exp[-T/T_p]) = EC_1Rf.$

The peak value of the ripple voltage superimposed upon the d.c. level is $V_r = EC_1/(C_1+C_2)$. As shown in Appendix (1) and in Appendix (2) the time-constant of the circuit is $T_p = R(C_1+C_2)$. Thus the three basic equations for the circuit may be written as:

$$\text{Gain} = V/f = EC_1R \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$\text{Ripple voltage } V_r = \frac{EC_1}{C_1 + C_2} \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$\text{Time-constant } T_p = R(C_1 + C_2) \quad \ldots\ldots\ldots \quad (3)$$

It is interesting to note that

$$G = T_pV_r \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Equations (1), (2) and (3) can be represented by the equivalent circuit shown in Fig. 3 when $T_p \gg 1/\omega_r$.

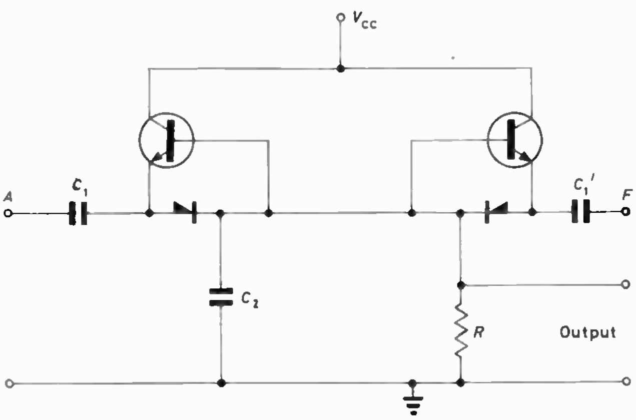


Fig. 4. Double input convertor

This circuit consists of a voltage source Gf in series with a noise generator $G\omega_r$ together with an RC network. The output across $C_1 + C_2$ is $Gf = EC_1Rf$ together with a superimposed ripple voltage $G\omega_r.\ 1/\sqrt{(1+\omega_r^2T_p^2)} \approx G/T_p = V_r$ when $T_p \gg 1/\omega_r$. A filter may now be designed to reduce the ripple voltage further using standard filter theory provided the network consisting of R and C_1+C_2 is considered as an integral part of the filter.

When $T_p \gg 1/\omega_r$ the ripple voltage is triangular in form so the amplitude of the fundamental component is $2V_r/\pi$.

A further advantage can be obtained by placing two convertors in parallel with R and C_2 common to both as shown in Fig. 4, and with the inputs at A and F 180° out of phase.

The output voltages due to the pulse trains applied at A and F for $C_1' = C_1$ are shown in Fig. 5(a) and Fig. 5(b) respectively. The sum of the two voltages as seen at the output is shown in Fig. 5(c). The peak value of the ripple voltage has not changed from $V_r = EC_1/(C_1+C_2)$ but its frequency has doubled, making it easier to attenuate it in a filter network. If $C_1' = C_1$, the gain of the convertor is doubled, i.e., $G = 2EC_1R$ but the time-constant has only been increased to $T_p = R(2C_1+C_2)$.

Other Circuit Considerations

When a voltage E is applied to A in Fig. 1, C_1 and C_2 charge up in series through the source impedance R_s. If the input is a square wave the voltage placed across the capacitors will vary with frequency unless $R_sC_1C_2/(C_1+C_2) \ll 1/2f$, thus introducing a non-linearity into the gain characteristic. The square wave is more desirable than the pulse train in order to get two waveforms 180° out of phase.

In deriving equation (2) the forward conduction voltage (V_D) of the diode and the 'cut on' voltage (V_{be}) of the transistor were neglected. When taking account of these

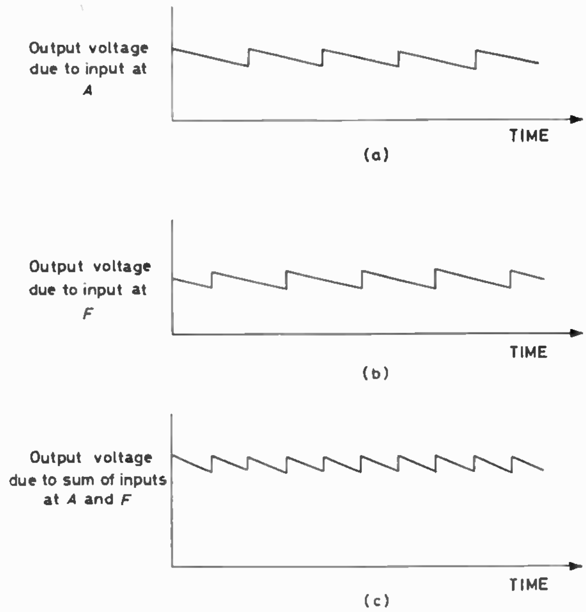


Fig. 5. Output waveforms of double pump

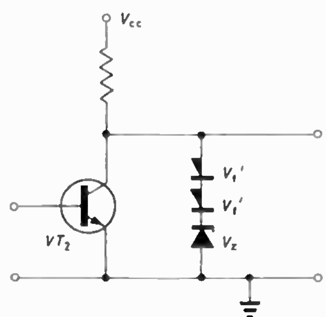


Fig. 6. Generator of pulses whose amplitude varies with temperature

the voltage applied to the circuit during each cycle is $E - V_D - V_{be}$ so the gain equation becomes $G = (E - V_D - V_{be})\ RC_1$. As both V_D and V_{be} have negative temperature coefficients, the gain of the circuit will have a positive temperature coefficient. The best method of compensation is by supplying the input from a saturating amplifier followed by a clipping circuit using two diodes and a Zener diode with a zero temperature coefficient. This circuit is shown in Fig. 6.

If suitable square waves are applied to the base of VT_2 the output will be a square wave of amplitude $V_Z + 2V_f' - V_{sat}$. When using silicon planar semiconductor devices throughout the drift in the diodes will largely cancel out the drift of V_D and V_{be}. The drift of V_{sat} is negigible. The percentage drift of the output is kept low by making E large. It is interesting to note that with this circuit the voltage aims at V_{cc} and therefore reaches the value of $V_z + 2V_f'$ exactly over a wide frequency, range thus overcoming the difficulty discussed above.

Conclusion

The basic design of the circuit is one of compromise betwen gain, ripple voltage and phase lag between output and input. The larger the value of E the smaller will be the drift of output voltage with temperature but the more difficult it will be to ensure a low value of R_s so this aspect of design is a compromise between drift and linearity of the output.

Additional control of the output can be obtained by varying R, C or E.

Acknowledgment

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Appendix

(1) Consider a step voltage E applied to the circuit shown in Fig. 7 which is the same as that in Fig. 1 when MR_1 is forward biased and VT_1 is "cut off". Using Kirchoffs law for loops (1) and (2) gives:

$$e(t) = (1/C_1)\int i_1 dt + (1/C_2)\int (i_1 - i_2)\, dt$$

and $(1/C_2)\int (i_1 - i_2)\, dt = Ri_2$

Writing these equations using Laplace transforms gives:

$$E/P = (I_1/C_1P) + \frac{I_1 - I_2}{C_2P}$$

$$\frac{I_1 - I_2}{C_2P} = RI_2$$

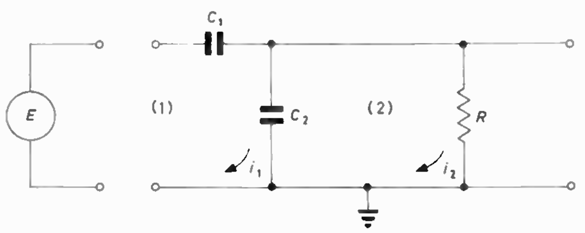


Fig. 7. Convertor equivalent circuit

where $L[e(t)] = L[E] = E/P$
$L[i_1] = I_1 \quad L[i_2] = I_2$

Solving for I_2 gives:

$$I_2 = \frac{(E/R)\ C_1/(C_1 + C_2)}{P + (1/(C_1 + C_2))}$$

$$\therefore\ i_2 = (E/R)\frac{C_1}{C_1 + C_2}\exp\left[-\frac{t}{R(C_1 + C_2)}\right]$$

the output voltage V_x is Ri_2

$$= \frac{EC_1}{C_1 + C_2}\exp\left[-\frac{t}{R(C_1 + C_2)}\right]$$

so the output has a decay time-constant of $R(C_1+C_2) = T_p$.

(2) Consider a step function of frequency f applied to the circuit in Fig. 1. On the first positive going edge of the voltage applied to A the voltage across C_2 assumes the value $EC_1/(C_1+C_2)$. On the second positive going edge it will have decayed to $EC_1/(C_1+C_2)\exp(-t/T_p)$ and will be increased to:

$$\frac{EC_1}{C_1 + C_2} + \frac{EC_1}{C_1 + C_2}\exp(-T/T_p)$$

$$= \frac{EC_1}{C_1 + C_2}\left[1 + \exp(-T/T_p)\right] = V_r(1 + y)$$

where $y = \exp(-T/T_p)$.

At the n^{th} positive going edge the voltage at the output will be

$$V_r(1 + y + y^2 + \dots y^n) = V_r\frac{(1 - y^n)}{1 - y}$$

The number of positive going edges required for this voltage to reach $(1 - (1/e))$ of its final peak value V_{op} is given by the equation:

$$\frac{V_r(1 - y^n)}{1 - y} = \frac{(1 - (1/e))\frac{EC_1}{C_1 + C_2}}{1 - \exp(-T/T_p)} = \frac{(1 - (1/e))\,V_r}{1 - y}$$

i.e. $1 - y^n = 1 - (1/e)$

or $n = \frac{\log(1/e)}{\log y} = T_p/T$.

The time taken for the output to reach $(1 - (1/e))$ of its final value is $nT = T_p$. Thus the time-constant of the circuit is T_p.

REFERENCES

1. Earnshaw, J. B. The Diode Pump Integrator. *Electronic Engng.* *28*, 26 (1956).
2. Burton, P. L., Willis, J. Unusual Transistor Circuits. *Wireless World.* *64*, 107 (1958).

A Transistorized Echo Sounder

Latest addition to the Kelvin Hughes range of echo sounders is the MS.39, a transistorized recording type instrument which has been designed specifically for fishermen and the owners of small boats.

Basically the MS.39 equipment consists of two units only, the recorder and the transducer. The straight-line recorder is watertight, and is available with single or dual scales providing ranges of 0 to 95 fathoms or 0 to 380 fathoms respectively. Continuous phasing allows the recorder to be set to any selected position of the total range scale. The full depth range is covered by one narrow transparent scale which does not obscure the echo trace. Reliability is assured by the use of transistors and printed circuits.

Depth soundings and fish echoes are presented to the observer on a truly orientated chart on 6in wide dry paper viewed through a window 12½in wide by 7½in deep. The instrument is available with Kelvin Hughes 'white line' record chart which enables echoes from fish on or near the seabed to be easily distinguished from the seabed echo itself.

The transducer is incorporated in a streamlined moulded fairing 8½in long by 3½in wide and weighing only 6lb.

Maximum power consumption is 40W, which makes the MS.39 suitable for craft with limited battery capacity.

A suitable headline transducer can be supplied to operate with this equipment.

The recorder unit

A 50kc/s Cold-Cathode Decade Counter

By R. D. Ryan*

A simple npn multivibrator provides an economical driving circuit for a 50kc/s cold-cathode gas-filled decade counting tube. D.C. restoring diodes on the guide rails ensure constant stepping pulse levels

(Voir page 799 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 806)

IN many counting applications the cold-cathode tube provides the simplest and most economical form of decade counter. These tubes are generally limited to maximum frequencies of approximately 10kc/s. However, the Z505S has recently become available with a frequency range up to 50kc/s, when driven by short 6μsec stepping pulses. This gives a decade dead time of 20μsec which is considerably less than that of Geiger tubes often used with such counters.

Transistor drive circuits are desirable to match the compactness and low power requirements of the Z505S. The driving circuit must provide negative pulses of about 100V with microsecond rise and fall times. This suggests the use of an npn switching transistor with a suitable collector base voltage rating. The planar type 2N1990 fulfills these requirements and has been used in the multivibrator circuit described below. This has the advantage of not requiring a large standby current or transformer coupling as is necessary with pnp transistors.

The driving circuits for cold-cathode decade tubes given in data sheets and in the literature show simple *RC* coupling to the guide rails. With irregular input pulse trains, arising for example, in nuclear counting, such coupling will cause the guide rail voltage level to shift from the selected supply voltage. Since the reliability of stepping the discharge from one cathode to the next, particularly at high stepping speeds, depends critically on the guide voltage; it is desirable that d.c. restoring diodes be connected to the guide rails to ensure constant stepping pulse levels.

Circuit Operation

The above principles have been given practical form in the circuit of Fig. 1. VT_1 and VT_2 form a monostable multivibrator triggered by positive input pulses applied through MR_1 to the base of VT_1. The 7μsec pulse duration is determined by the time-constant C_3R_{10}. VT_1 collector, normally held at 90V by MR_3, is switched to 0 volts during the pulse. This negative pulse provides the drive pulse to V_1 guide rail *A*.

At the end of the pulse, when VT_1 is switched off, MR_4 disconnects the collector from C_3 and C_5 to give a more rapid return to the 90V level. Recharging of C_5 generates a positive driving pulse at VT_3 base, switching it on. The resulting negative pulse at VT_3 collector drives V_1 guide rail *B*. Diodes MR_2, MR_6 and MR_7 ensure that the emitter base voltage ratings of the transistors are not exceeded. VT_2 may be either a 2N1990 or a similar transistor with a lower V_{CBO} rating, e.g. 2N2270.

MR_9 and MR_{10} are d.c. restoring diodes which fix the positive level of the driving pulses at 45V.

The K_0 cathode output pulse decreases in amplitude from 20V to 10V as the frequency is increased to 50kc/s. However, it is still adequate to drive a similar second decade counter stage; since the maximum frequency of this stage will be 5kc/s the values of C_3 and C_5 may be increased to give longer guide rail pulses.

The total power consumption for the 50kc/s stage is 1·5W. For the second and following stages, the 475V supply currents to VT_1, VT_3 and V_1 can be reduced without affecting performance. Making R_4, $R_{14} = 2M\Omega$ and $R_{17} = 680k\Omega$ reduces the power per stage to 0·4W.

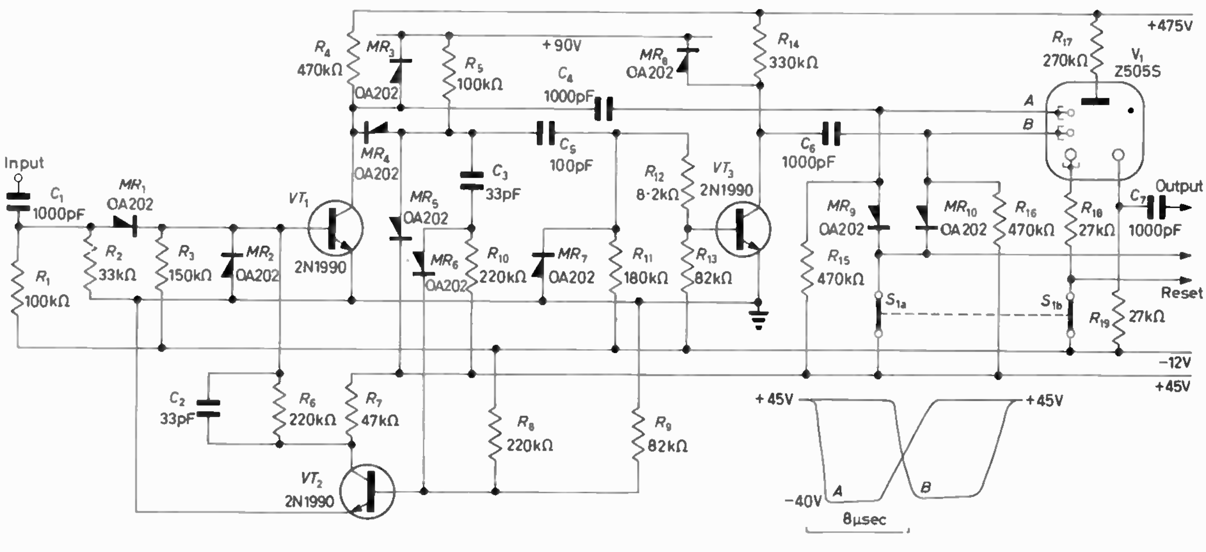


Fig. 1. The counter circuit

* *Australian Atomic Energy Research Establishment, formerly University of New South Wales.*

Short News Items

The Scientific Instrument Manufacturers' Association has been accepted to Associate Membership of the European Committee of Optical and Precision Mechanical Instrument Manufacturers.

This is an international organization which acts in an advisory capacity to the Common Market Commission. It exists to discuss matters of common interest in six specialized committees, as follows:

Committee 1. Ophthalmic Instruments.

Committee 2. Optical Instruments.

Committee 3. Photographic, Cinematographic Instruments and Projectors.

Committee 4. Precision Mechanical Instruments with five sections.

Committee 5. Measuring, Regulation and Control Instruments.

Committee 6. Medical and Surgical Instruments.

None of these committees can deal with electrical or electronic instruments, since these come within the field of other Trade Federations in Germany and France.

The governing body is the Comité Général which consists of four presidents, one each from France, Belgium, Germany and Italy and the chairman and vice-chairman of each of the specialized committees.

SIMA has applied and been accepted as Associate Member in respect of Committees 2, 4 and 5. The Association has also introduced and agreed to act as liaison for the representatives of other Trade Associations who also cover these several sectors.

One of the benefits accruing from membership of the Committee is the exchange of statistical and other trade information. SIMA will be providing similar information regarding the appropriate sector of British industry.

The International Federation of Automatic Control is organizing a symposium on Automatic Control in the Peaceful Uses of Space. The symposium will be held at Stavanger, Norway, from 21 to 24 June 1965. The following topics will be discussed: Injection into space; attitude stabilization; remote control of space vehicles; problems of manned systems; ground systems; reports on control of specific systems; advanced components for space vehicles and instrumentation control problems; digital and analogue computers, as used in automatic control systems in space vehicles or ground stations; future control problems.

The official languages of the Conference will be Russian and English.

Further details may be obtained from Lars Monrad-Krohn, P.O. Box 66, Kjeller, Norway, to whom offers of papers with abstracts should be submitted as soon as possible.

The Tenth International JUREMA Exhibition and Seminar is to be held next year at Zagreb, Yugoslavia.

Further details of the Exhibition and Seminar may be obtained from Jurema, Zagreb, POB. 123, Yugoslavia.

The International Computation Centre (ICC), Rome, announces its first Regional Seminar which will be held at Eindhoven, Holland, from 23 November to 3 December 1964.

The aim of the Seminar is to spread knowledge in automatic data processing and its problems among the heads of central and local governments and other public organization departments.

Further details may be obtained from The International Computation Centre, 23 Viale della Civilta del Lavoro, Rome (EUR), Italy, or The Netherlands Automatic Information Research Centre, 6 Stadhouderskade, Amsterdam, Holland.

STD subscribers in Liverpool, Manchester, Birmingham, Edinburgh and Glasgow are now able to dial direct to the Netherlands, France, Belgium, Western Germany and Switzerland.

Pye Laboratories Ltd of Cambridge has received an order from France for the installation of nuclear reactor television cameras for installation in the reactors of Electricité de France at Chinon in the Loire Valley.

Specially designed cameras, capable of withstanding high temperatures, radiation levels and pressures, will be used to carry out inspection and recovery operations in the heart of the reactors.

Remotely controlled mechanical grabs will be fitted to the cameras to enable debris to be removed from the control and fuel rod channels. Pye have supplied similar installations for nuclear reactors in Italy, Denmark and Japan.

The Ministry of Aviation has placed an order with Decca Radar Ltd for a new Airfield Surface Movement Indication radar for installation at London Airport.

This new radar, now under development, will be capable of continuous movement presentation. Operating on the 8mm wavelength, with an aerial rotation rate approaching 1 000rev/min the equipment is superior to any previous a.s.m.i.

Decca Radar Ltd developed and manufactured the first a.s.m.i. radar system for civil aviation use, which was installed at London Airport early in 1955. This radar, which is still in use, operates in the Q-band (8mm) and provides high definition and clarity on a conventional radar display at a data rate of 20 per minute.

But with comparatively slow data rates, targets can move too far before new positional information is available. The new Decca A.S.M.I.3 equipment now ordered as a replacement for the existing equipment is due to come into operation in about two years time.

Negotiations are in hand for installation of this new radar set at other major airports in Europe.

Aveley Electric Ltd has fitted an automatic immediate replay device to its portable language laboratory.

In a standard language laboratory the student switches his recorder on and hears the master track which contains phrases in the language being taught. The phrases having gaps between them allows the student to repeat the phrase he hears. The students rendering of the phrase is recorded on the bottom track. To hear his recording the student has to operate either a lever or press a button or switch and cause the recorder to rewind to the point at which the master phrase was heard. He then releases the lever or switch and the recorder is put into replay condition and the student hears again the master phrase followed by his rendering of the phrase. He can thus hear his mistakes and effect an improvement.

With the system introduced by Aveley Electric Ltd the student hears the phrase from the master track, speaks his rseponse into his microphone and as soon as he has finished speaking he hears his response replayed to him.

The student has no mechanical operations to perform. This saves the time spent rewinding the machine, also the student does not need to hear the master phrase again.

The Royal Aircraft Establishment, Farnborough, has placed an order valued at £40 000 with English Electric-Leo Computers Ltd for four automatic data loggers, each of which can continuously monitor and record temperatures and strains at 768 different points on an airframe under test.

The electronic loggers will provide high speed information during extensive ground tests of current and future aircraft, probably including the Concord and TSR-2. Temperature and stress readings from aircraft in test rigs will be recorded by the loggers on punched paper tape ready for computer analysis.

This order follows similar pioneering work by the company in co-operation with R.A.E. in the production and installation of equipment to simulate the kinetic heating effect of air friction in high speed flight.

Two of these kinetic heating rigs have been used by the R.A.E. since March 1961. The data loggers now ordered may be linked to these heating rigs or to 'torture chamber' strain rigs in which complete airframes are subjected to flexing and twisting forces greatly in excess of those which they will meet in flight.

Two British companies, Associated Electrical Industries and Elliott Automation are members of an international consortium which has been formed to compete for a £110M contract for a new NATO defence network against air attack. The other three in the new consortium are Litton Industries Inc. of America, leaders of the consortium, CSF Compagnie Generale de Telegraphie sans Fil of France, and International Telephone and Telegraph Co., Europe.

The proposed NATO network to be known as NADGE (Nato Air Defence Ground Environment) will provide instant response against attack by any of the most advanced supersonic aircraft. It will defend the entire NATO front, with a chain of defence stations stretching through nine countries from Norway to Turkey. Contracts are expected to be awarded mid-1965.

An electro-mechanical page turning device designed to assist physically disabled people to enjoy the benefits of normal reading has been developed by the Engineering Department of Cambridge University and tested at the National Spinal Injuries Centre, Stoke Mandeville. The device is designed for use with paper-backed books and an able bodied person first has to position the book, but from then on page turning operations are carried out entirely by the reader.

The whole unit is mounted on a mobile tubular-steel frame that is positioned so that the book is held face-down over the reader. Height and angle are both adjustable and a reading light is incorporated. Page turning is derived from normal reading actions, the book being mounted with its pages bent back along a curved metal guide and the splayed edges held firmly in position by a sprung metal 'finger'. A remote switching unit containing two micro-switch buttons is placed in any convenient position for the patient and provides the means of initiating and controlling page turnover. Pressure on one button causes the 'finger' to retract slightly so that a single page is released, to 'flip' over and come to rest against the perspex window. Pressure on the second button causes the window to move aside momentarily, allowing the page to fall forward slightly, so that, as the window returns to its original position, the page is automatically turned over.

The Compagnie Generale de Telegraphie Sans Fil (CSF) and the General Dynamics Corporation are to form a new French company in which CSF will hold a major interest.

The new company, which will design and manufacture satellite tracking equipment, is to be known as Societe D'Equipements Spatiaux et Astronautiques (SESTRO) and will be located at Corbeville, near Paris.

SESTRO will combine the expanding activities of the CSF Space Division and the CSF experience in electronics with the technological capability in space systems of General Dynamics, which designed and built the AZUSA missile tracking system and GLOTRAC tracking network used in United States missile and space programmes.

The new Severn Bridge, to be opened towards the end of next year, is to have a toll collecting system which will include an ARCH 1000 computing system, supplied by Elliott Traffic Automation, a member of the Elliott-Automation Group, and designed to a Ministry of Transport specification. There will be ten traffic channels on the bridge with toll gates at the Bristol end. Normally four channels will permanently carry traffic in each direction, the direction on the remaining two being changed to accommodate the changing flow. All ten channels will be in operation 24 hours a day.

Tolls will vary according to the type of vehicle and the number of axles. Each collector will have an identifying key which he will insert into the computer console at the beginning of his tour of duty. Thereafter all the tolls he charges will be recorded by type and size of vehicle against his code number so that a printed record of all vehicles using the bridge, their direction of travel and the tolls paid on each channel will be instantly available to the management, showing also how much money each collector has taken in his tour of duty.

Ferranti Ltd has been awarded the U.K. 3 satellite solar cell contract by the Ministry of Aviation for the supply of 38 000 MS20 silicon solar cells having a minimum conversion efficiency in outer space of 10·5 per cent and 13·5 per cent at the earth's surface. The cells will be mounted in large arrays on the paddles of the U.K. 3.

A new submarine telephone cable system between the United Kingdom and Denmark has recently been brought into operation.

The cable, which contains 120 circuits, runs between Winterton, Norfolk and Maade, near Esbjerg, Denmark, and comprises some 298 nautical miles of 0·62in diameter coaxial cable with 24 submerged repeaters and one submerged equalizer along its length. The laying, which was completed during May of this year, had to be closely controlled to ensure that the last repeater at the U.K. end was correctly positioned in relation to the Winterton repeater station. This was achieved and the repeaters were all laid in positions very close to their planned locations. The new cable is the second in a series of five cables across the North Sea which are planned to be completed by 1966.

The Hawker Siddeley Group is co-operating with the G.P.O. on the extension of the satellite ground station at Goonhilly Down to enable the aerial to be used with the satellite 'Early Bird' after its launching by the American Communications Satellite Corporation in March 1965. The satellite communication system is to be used for transatlantic telephone links to supplement the transatlantic cables.

Hawker Siddeley Group, with Husbands & Co., consulting engineers to the G.P.O., designed and installed the control system for the steerable aerial which is at present being used for experimental work for communications via a transitory satellite. 'Early Bird', however, will be a geo-stationary satellite appearing almost stationary to the aerial and will thus allow 24 hour a day communication over the Atlantic.

The present Goonhilly Down ground station has proved to be extremely reliable during its 18 months of operational service and now that a permanent multi-channel telephone link is to be put into operation, modifications are being incorporated to duplicate essential parts of the ground station in order to ensure the continuity of operation required in a vital communications link.

Ekco Electronics Ltd is to develop the highly specialized airborne weather radar for the prototype Concorde. This type of equipment, for detection of turbulent flying conditions, is already provided by Ekco for many of the world's air lines. However, the radar requirement of the Concorde will be more exacting than those of any aircraft yet designed. A completely new system, using the most advanced techniques, is necessary to provide the increased detection range and to give the very high standard of reliability required for supersonic operation.

A new digital liquid-level measuring system, developed by the Datex Division of Elliott-Automation, makes possible the precise and completely automatic monitoring by a master station of the level of water or other liquid at any number of slave stations which can be many miles away. The system is completely passive until interrogated and can be applied to the level measurement of rivers, reservoirs, dams and tanks.

The system operates over a two-wire line providing an accuracy of 0·1 per cent or better and also a high degree of integrity as it is not affected by interference or any changes in the characteristics of the line.

The basic equipment consists of a simple and robust shaft encoder. No power supplies are necessary at the point of level measurement. The information on the encoder is sampled by a scanner at the slave station and transmitted in the form of digital pulses to the master station. There the pulses are stored and decoded as required to provide a visual display, and/or a printed record of the height of the liquid at the remote monitoring point.

The British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA) state that the total sterling values of their members' sales of valves, tubes and semiconductor devices during the three months ended 30 June 1964 were as follows: Valves and tubes, £12·4M, semiconductor devices, £6·2M.

This brings the total value of sales for the first half of 1964 to £37·1M. The figure for the whole of 1963 was £61·6M.

The British Post Office has placed a contract with The M-O Valve Co. Ltd for high power travelling wave tubes for use in the 'Early Bird' system, the world's first commercial communications satellite.

The tubes are water-cooled C-Band travelling wave amplifiers, giving an operating power output of 10kW at a frequency of 6301Mc/s and will be used as transmitter amplifiers at Goonhilly Downs Station.

The tube uses a coupled-cavity slow wave structure of a type which gives high gain per unit length, good power handling capacity and freedom from unwanted oscillations. The tuning range of the r.f. structure is over 225Mc/s and the small signal bandwidth better than 30Mc/s (2dB points).

Focusing is achieved by means of rugged copper foil coils wound directly on to the tube body. This technique produces very good alignment between magnetic field and electron beam, thus reducing interception of the beam on the structure to a minimum. It also results in a much more compact assembly than could be obtained with a conventional solenoid, since the solenoid does not have to be large enough to pass over the waveguide outlets and also the good thermal contact between the coils and the body, which is water-cooled, avoids the necessity of having cooling fins or a cooling jacket on the magnet.

Space Environment Simulators is the subject of a one-day symposium to be held at Northampton College of Advanced Technology, St. John Street, London, W.1, on 17 November 1964. This symposium is sponsored jointly by the British Interplanetary Society and the Society of Environmental Engineers. Enrolment forms are available from the British Interplanetary Society, 12 Bessborough Gardens, London, S.W.1.

A new all-electronic desk calculator has been designed by Mullard Ltd to demonstrate the suitability of a range of cold cathode tubes and semiconductors to this type of application. Although a commercial machine will not be produced by Mullard, the circuits employed will be made available to manufacturers.

The machine, which uses both transistors and cold cathode tubes, employs the decimal counting system throughout. It has an information store comprising three registers each with twelve decades, but could be adapted for two or four registers of different size with only minor modification.

One of the main objects of the investigation has been to reduce the cost of building such machines while maintaining a high reliability. Considerable effort has been devoted to the choice of components, and also to investigating new circuit techniques. This has resulted in a new 'dynamic display system' which has considerably reduced costs. With the circuit techniques and arithmetical approach employed it has been possible to obtain, using twelve decades, a similar performance to a '10 × 10 × 20' mechanical machine. A floating and completely automatic decimal point is provided on all calculations.

In choosing a keyboard the advantages and disadvantages of both the full keyboard (100 number keys for 10 decades) and the simplified keyboard (10 number keys) have been considered. This machine has been designed with a simplified keyboard in order to preserve the advantage of simple operation which is one of the main features of electronic control. Only minor changes would be required to incorporate a full keyboard.

All the normal functions of multiplication, division, addition, subtraction and also transfer and back transfer between registers are provided for. The contents of the registers are displayed on numerical indicator tubes and gas diodes are used for decimal point indication. There are twelve number tubes corresponding to the twelve decimal counters of the registers.

Dorval Airport, Montreal, Canada, a major North American and international air centre is the seventh major airport to install the Solartron Transistorized Air Traffic Control Simulator for the training of future air traffic controllers.

The two target, transistorized radar simulator, designed and manufactured by the Military Systems & Simulation Division of the Solartron Electronic Group Ltd, Farnborough, Hampshire, has been ordered by the Canadian Department of Transport. The two synthetic aircraft targets may be realistically controlled during both s.r.e. and precision approach radar working over a radar range of 0 to 40 nautical miles, with responses simulating the returns from the runway and offset markers included for p.a.r. simulation. In addition, the simulator may be locked to the live radar equipment for both s.r.e. and p.a.r. operation.

Dawe Instruments Ltd, Acton, London, W.3, has sponsored the founding of a Lectureship (to be known as The Dawe Lectureship) at Southampton University's Institute of Sound and Vibration Research, which will foster research and teaching in the field of noise measurement, under the Directorship of Professor E. J. Richards. Mr. P. L. Tanner has been appointed to the position of Dawe lecturer.

LETTERS TO THE EDITOR

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

Infinite Array of 1Ω Resistors

DEAR SIR,—We are extremely thankful to Mr. Scraton for his interesting letter (June 1964, page 412) wherein a general solution to the above problem for the case of a square mesh in two, as well as in three dimensions has been quoted with a number of interesting observations.

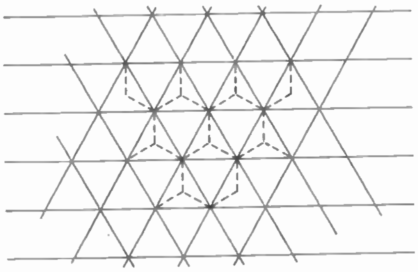


Fig. 1. Conversion to hexagonal meshes

It is, however, felt that in the above cases, the solution has been considerably simplified as the network conformed to an orthogonal geometrical frame. In the following, an attempt has been made to generalize the solution to include other possible configurations as well, with the only imposition that there should be a translational symmetry, i.e. all the nodes are equivalent.

First, it follows from topological considerations, that the regular figures which can fill a plane completely, are finite in number, viz. 'square', 'equilateral triangle' and 'regular hexagon'. The case of the square mesh has already been dealt with and the other two cases will be discussed here. Further, it is observed (Fig. 1) that by simple delta-star transformation, the triangular meshes can be converted into hexagonal meshes and hence it is sufficient to consider the triangular mesh structure only, to complete the generality.

A general node in the case of an infinite triangular mesh structure (Fig. 2) can be designated by the double suffix (m, n) referred to oblique axes chosen as shown in the figure.

We are interested in the resistance $(R_{m,n})$ between node $(0, 0)$ and the general node (m, n). The nodal equation for the general node (m, n) is

$$(V_{m+1,n} - V_{m,n}) + (V_{m,n+1} - V_{m,n}) + (V_{m-1,n} - V_{m,n}) - (V_{m,n-1} - V_{m,n}) + (V_{m-1,n+1} - V_{m,n}) + (V_{m+1,n-1} - V_{m,n}) = I_{m,n}$$

where $I_{m,n}$ is the current input at the node (m,n).

The solution is a difference equation and we are interested in the solution only for integral values of m and n. Following the normal approach ('Operational Calculus' by Van der Pol and Bremmer, page 365) we get:

$$R_{m,n} = 1/4\pi^2 \int_{-\pi}^{+\pi}\int_{-\pi}^{+\pi} \frac{1 - \cos(mp+nq)}{1 - \cos p - \cos q - \cos(p-q)}\, dp\, dq$$

This integral differs significantly from that met with in the case of square mesh structure, in that it contains a mixed term $(\cos(p-q))$ in the denominator. In general this must be evaluated numerically except for the case of adjacent nodes where it yields:

$$R_{10} = R_{0,1} = R_{-1,1} = R_{-1,0} = R_{0,-1} = R_{1,-1} = \tfrac{1}{3}\Omega$$

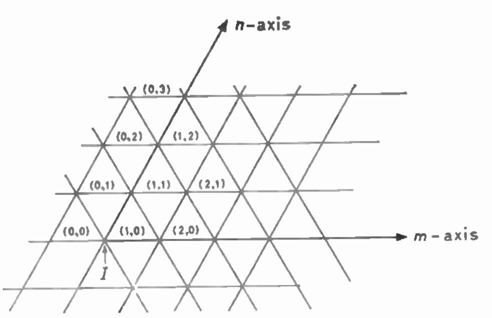
It is readily seen that the integral is divergent for $|m|+|n|$ tending to infinity, i.e.

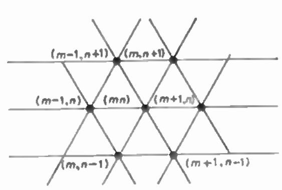
$$R_{m,n} \to \infty$$
$$|m| + |n| \to \infty$$

Yours faithfully,

A. PRABHAKAR and
N. KUMAR
Institute of Armament Technology,
India.

Fig. 1. Infinite triangular mesh





The Correspondent replies:

DEAR SIR,—The mesh considered above is a particular case of a far more general type of mesh which can be dealt with by this method. The triangular mesh is, of course, exactly equivalent to a square mesh with an additional resistor along one diagonal of each square (Fig. 1). A rather more general result may be obtained by including both diagonals, as shown in

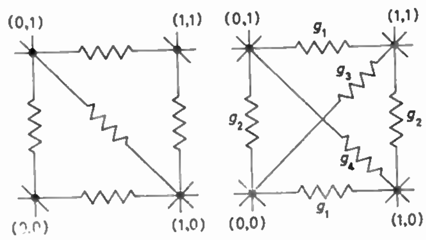


Fig. 1 (left). Square mesh with one diagonal
Fig. 2 (right). Square mesh with two diagonals

Fig. 2. It is not necessary for every resistor to be of equal value as long as all parallel resistors are of the same value. If the *conductances* of the resistors are as indicated in Fig. 2, then the resistance between the nodes (0,0) and (m,n) is:

$$1/4\pi^2 \int_{-\pi}^{\pi}\int_{-\pi}^{\pi} \frac{1 - \cos(m\theta + n\phi)}{g_1(1-\cos\theta) + g_2(1-\cos\phi) + g_3(1-\cos(\theta+\phi)) + g_4(1-\cos(\theta-\phi))}\, d\theta\, d\phi$$

The result of Messrs. Prabhakar & Kumar is given immediately by putting $g_1 = g_2 = g_4 = 1$ and $g_3 = 0$.

Furthermore, if an additional set of resistors of conductance g is included so as to connect each node (x,y) with the node $(x+a, y+b)$, it is merely necessary to add an extra term $g[1-\cos(a\theta + b\phi)]$ into the denominator of the integrand. In this way the complexity of the mesh can be increased indefinitely.

Yours faithfully,

R. E. SCRATON,
Northampton College, London.

P.S.—The denominator of the main equation given in Messrs. Prabhakar and Kumar's letter should read "$3-\cos p -$ etc., *not* $1 - \cos p -$ etc."

BOOK REVIEWS

Sampling System Theory and its Application Vols. 1 and 2

By Ya. Z. Tsypkin, 742 pp. Med. 8vo. Pergamon Press. 1964. Price £5.

THIS English translation of a book written by such a well-known author should arouse interest, if only because it gives the Russian viewpoint on this important topic. The book appears to lose little by being translated, as the text is very readable.

The basic approach is conventional in that the analogy is stressed between sampled systems and continuous systems.

Vol. 1 comprises four chapters dealing with descriptions of pulse systems, mathematical techniques for analysis, fundamentals of open-loop systems (including those with variable parameters) and typical examples of such systems.

Vol. 2 consists of two chapters on the fundamentals of closed loop systems with again some typical examples. The use of a digital computer as an element of a control system is briefly considered.

Perhaps the best feature is the detailed consideration given to the examples of the various types of systems, and this aspect should prove to be of interest to the control engineer. Another good feature is the extensive bibliography which covers Western and Russian work in this field.

If criticism is warranted. it would be that more attention could have been paid to variable rate sampling and sensitivity analysis of systems subject to parameter variation.

Finally the book is intended for readers with a basic knowledge of the theory of continuous systems.

R. J. A. PAUL.

An Introduction to Electronics

By B. V. Rollin. 216 pp. Med. 8vo. Clarendon Press, Oxford. 1964. Price 30s.

THE author of this book is a lecturer in physics at the University of Oxford, and presumably the text is related to lectures given to undergraduates reading honours physics at that university. The preface states that the book is intended for the use of students of physics and engineering in universities and technical colleges from first year undergraduate to first year graduate level. It is difficult to see how one volume could cater for the needs of such a wide range of readers and, in fact, the book does not seem to be suitable either for students of engineering or for those on technical college courses.

The first chapter gives a brief survey of linear circuit theory; resonant circuits are discussed without consideration of fractional detuning, and coupled circuits are covered in one page. The same chapter also deals with transmission lines, filters, waveguides and even stub matching. A short chapter on components is followed by an introduction to vacuum tubes. Miller effect in the triode is not mentioned and valve characteristics are poorly drawn. Chapter 4 is on vacuum tube amplifiers and oscillators; the gain expression for the triode is derived from the differential relations, and not from the equivalent circuit, and the effect of coupling and decoupling capacitors on frequency response in multi-stage amplifiers is not explained. A chapter on rectifiers and filters completes the introductory material on thermionic valves.

Chapter 6 deals with semiconductors; here the treatment is much more advanced than in the previous part of the book and most of this chapter is "starred" in order to indicate that it is only suitable for final year and graduate students. Semiconductor devices and circuits are in Chapter 7. The manufacture of the junction transistor is described but typical characteristics are not given. The chapter contains a number of typical circuits, these do not have component values and are described very briefly. A short chapter on noise is followed in Chapter 10 by a "starred" discussion of high frequency valves and techniques including the travelling wave amplifier and the maser. Receivers, optical devices and aerials are then discussed leading to a final chapter on laboratory experiments.

The book covers a bewildering range of topics and shows large variations in the level of treatment. The elementary material is already available in well established texts and the more advanced topics would. in any case, require further reading. The book does not appear to be suited for teaching purposes.

V. H. ATTREE

Frequency Filter Methods

By T. Laurent. 320 pp. Royal 8vo. J. Wiley & Sons Ltd. 1964. Price 132s.

IN the Foreword to this book Dr. Scowen describes it as a 'technical autobiography' since it presents the contributions made by Professor Laurent to the theory of wave filters over a period of about 30 years. The author is best known for his use of frequency transformations in filter design and for his invention of the template method. The first two-thirds of his book is used to explain frequency transformations and then to use them in a wide variety of ways, including in the use of templates and the design of the zigzag band-pass filter he invented. In this section Professor Laurent also expounds his method of 'echostant matching', in which the matching of a ladder filter to the terminating resistances is designed to maximize and hold constant the image return loss and hence to keep the echo constant and small.

Although this material is derived from the papers published by the author the book is in no sense a simple compendium of these. The argument flows smoothly and logically and the reader will inevitably be impressed by the power of the methods so carefully yet succinctly explained.

In the remaining section the author deals with transients in filters, the design of attenuation and phase equalizers, frequency separators, the effect of inserting a ring modulator between filters and a number of other problems.

The book is well-produced and can be recommended to all filter designers and to those interested in the more practical aspects of network theory.

J. T. ALLANSON.

Square-Loop Ferrite Core Switching

By P. A. Neeteson. 185 pp Med. 8vo. Philips Technical Library. 1964. Price 47s. 6d.

DR. NEETSON in his third book on non-linear components in pulse circuits deals with the square loop ferrite toroid. This component has been in use in digital circuits for over a decade and its applications continue to grow. Accordingly it is timely that a book dealing with the practical aspects of circuit design using these components should appear. In this book a comprehensive treatment of the problem of practical design has been made.

The fundamental aspects of magnetization reversal are introduced in terms of the reversible and irreversible switching processes and these dynamic processes are related to the static hysteresis loop. A necessarily approximate model for irreversible switching is presented upon which the analyses for various load conditions are made. Equations are derived in terms of toroid parameters for the switching voltage as a function of time and flux for open-circuit, resistive and inductive loads. The very important case of a toroid loaded by other toroids is also treated. A chapter is spent in relating the calculated and measured values of peaking voltage and time. Quantitative agreement is good for the resistive load but poorer for the other cases. Nevertheless, the model is useful in predicting qualitatively the nature of the switching waveform.

The practical cases of the design of switch core drive cricuits for coincident current and word organized memory systems are given in detail and also a comprehensive treatment of the ferrite core shift register. Two chapters are spent extending the simple model to account for the practical limitations of large outer to inner diameter ratio of the toroid and the effect of drive pulses with finite rise times. Switching by a voltage source and energy dissipation in a switch core are also dealt with. The book concludes with a chapter on dynamic hysteresis loops.

Although it does not present a complete solution to the complex problems of designing circuits using square loop ferrite toroids, the book affords a valuable guide and as such should be read by anyone concerned with the design and use of square loop ferrite circuits.

D. C. STAPLETON.

Electrical Circuits with Variable Parameters including Pulsed Control Systems

By V. A. Taft, translated by F. Immirzi, translation edited by R. C. Glass. 109 pp. Demy 8vo. Pergamon Press. 1964. Price 50s.

THERE is no doubt that systems with variable parameters are becoming increasingly important. A book, setting out between two covers methods enabling the practising engineer to evaluate the properties of such circuits would be most welcome. Judging by the list of contents and relying on the great reputation of the author, V. A. Taft has written just such a book for the Russian reading engineer. The hopes of the mere English reading engineer are raised by the present title. If he is an accomplished mathematician he will be able to read and understand the formulae in spite of the fact that even they are presented in a form differing in some minor details from usual English practice. The practical man who hopes to learn to handle circuits with variable parameters from this book will soon find that the translation has been contrived with great ingenuity apparently for the sole purpose of obscuring the connexions between the mathematical steps. However, given time, dedication and an utter indifference to the accepted conventions of the English language an industrious student can undoubtedly work through the 106 pages closely printed on beautiful paper. It would appear that English is not the translator's mother tongue, but this fact, though deprecated by professional translators, would have been of little importance if the "Translation Editor" had done some editing. However, it is realized that the translation of highly specialized texts is no easy matter and the fees which teams of expert translators and editors would have to be paid might easily make a technical book wholly uneconomical, particularly in view of the comparatively small editions to be expected. Once it is generally recognized that technical translation requires not only a thorough command of the languages involved but also detailed understanding of the subject matter the Universities might consider this type of work as suitable tasks for post graduate students qualifying for a master's or even a doctor's degree.

K. L. SELIG

Elements of Feedback and Control

By A. M. Hardie. 344 pp. Demy 8vo. Oxford University Press. 1964. Price 55s.

THE book, written primarily for students working for the H.N.C. or Part III of the I.E.E. examination, is claimed in the Preface to be "intermediate in treatment between the very elementary and the over-advanced". In fact, the author has chosen to avoid the Laplace transform and to omit the Routh-Hurwitz criteria, the inverse Nyquist locus, root-locus methods and simple non-linear system—with no great loss for an introductory test but falling far short of advanced level. About one-fifth of the book is devoted to negative feedback amplifiers and the remainder to the performance, stability and means of improving the characteristics of some simpler linear systems.

The author has clearly drawn heavily on considerable teaching experience and has appreciated many of the difficulties students have with feedback and control systems. His emphasis on the dimensional checking of equations, his efforts to bring out the common features of servomechanisms and feedback amplifiers and his convincing treatment of Nyquist's criterion are commendable examples of this.

Quite rightly, a considerable proportion of the book is devoted to the properties of systems whose behaviour is described by the second order differential equation with constant coefficients. Unfortunately, the phase characteristic for such a system which is given in Fig. 15(b) incorrectly implies a discontinuity of 180° at the undamped frequency. This is not supported by the correct polar plot of Fig. 19 but the error is propagated in a worked example at a later stage (Ex. 2 p. 64). Numerous worked examples and a good supply of exercises, with answers, at the end of each chapter are a valuable feature of the book.

As a class textbook some lecturers might prefer a more concise treatment covering more ground but the book has much to recommend it to those who have to obtain a basic understanding of the topic mainly by private study.

R. L. FORD

Nonlinear-Magnetic Control Devices

By W. A. Geyger. 406 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price £5

This book is concerned with the non-linearity of the magnetic saturation characteristic of 'soft', or easily magnetized, ferro-magnetic materials and saturation phenomena is utilized to obtain some useful results.

In addition to magnetic amplifiers, the book discusses many other non-linear-magnetic devices, such as ferroresonant circuits, magnetic frequency multipliers, switching-transistor magnetic-core multivibrators, flux-gate magnetometers, and combinations of magnetic-core arrangements with various types of semiconductor devices (transistors, silicon-controlled rectifiers, etc.).

Gas Discharge Tubes

By H. L. Van der Horst. 318 pp. Med. 8vo. Philips Technical Library. 1964. Price 57s. 6d.

The principles of action of gas discharge devices and their construction are fully dealt with in this book and there are chapters on hot and cold cathode types, thyratrons and mercury cathode tudes. Other sections deal with special developments in the field and photocells with gas amplification.

Filament Winding (Its Development, Manufacture, Applications and Design)

By D. V. Rosato and C. S. Grove. 360 pp. Med. 8vo. J. Wiley & Sons. 1964. Price 113s.

It is claimed that this is the first book to be devoted in its entirety to the technique of filament winding. Numerous windings for aerospace, hydrospace, and commercial industries are discussed in detail. In each application the winding patterns are based on fibre geometry, fundamental stresses for various shapes and fittings, and the properties of the composite materials. The author lists and evaluates specifications, test methods, and quality controls as well as critical cost analysis.

Micropower Electronics

Edited by E. Keonjian. 216 pp. Royal 8vo. Pergamon Press. 1964. Price 84s.

This book contains a series of eight lectures organized by the Avionics panel of the Advisory Group for Aeronautical Research and Development (A.G.A.R.D.).

These lectures dealt with the general subject of reducing power dissipation in micro-electronic devices; and were presented in France, Germany, Italy and the United Kingdom.

Selected Papers on Semiconductor Microwave Electronics

Edited by Somner N. Levine and R. R. Kurzrok. 297 pp. Med. 8vo. Constable. 1964. Price 18s.

In this book are 26 selected papers dealing with use of the p-n junction to achieve amplification and frequency conversion of microwave frequencies. These papers have been previously published in the Proceedings of the I.R.E., Journal of Applied Physics, the Bell System Technical Journal and elsewhere.

Pressure Measurement in Vacuum Systems

By J. H. Leck. 221 pp. Demy 8vo. 2nd Edition. Chapman & Hall Ltd. 1964. Price 45s.

This edition although it follows the same approach and layout of the earlier edition has been revised and considerably increased in length to take account of new developments in the subject. The material is presented in seven independent chapters, the first five of which describe the different methods of pressure measurement in common use. The final chapters are devoted to gauge calibration and to residual gas analysis. The chapter, new this edition, entitled "The Analysis of Residual Gases in Vacuum Systems" deals exclusively with the use of mass spectrometers as vacuum analysers.

ELECTRONIC EQUIPMENT

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

(Voir page 793 pour la traduction en français; Deutsche Übersetzung Seite 800)

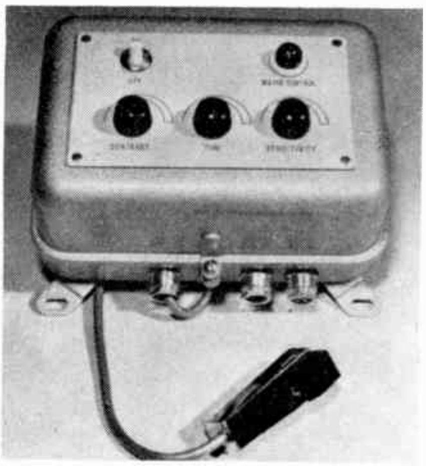
PHOTO-ELECTRIC DETECTOR

Simmonds Relays Ltd, Edinburgh Place, Temple Fields, Harlow, Essex

(Illustrated below)

Latest Visolux photo-electric device from Simmonds Relays is the RL.1, a photo-electric reflecting light gate for scanning printed edge marks on paper, plastic and other moving reeled materials without making physical contact. Equipped with an impulse amplifier of high sensitivity the RL.1 can, by its scanning action, bring other operations into play such as subsequent guillotining.

The RL.1 equipment consists of a compact scanning head containing a projector lamp and receiver, the whole unit measuring only 65mm long × 18mm across and at its deepest 30mm. The scanning head is joined to the main amplifier by a 2m cable lead.



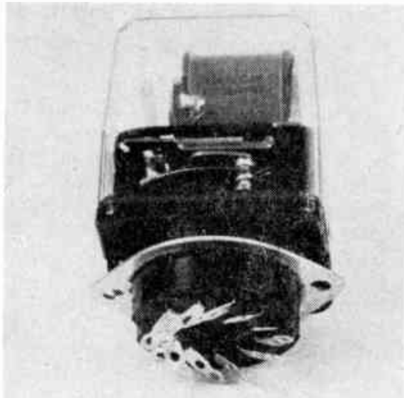
The impulse amplifier switch is fully transistorized and contains a built-in a.c. mains pack for 240V, 50c/s and in addition to on/off switch and mains warning light, is fitted with three controls for adjusting for sensitivity contrast and time delay. The fall off delay of the output relay is adjustable between 40 and 150msec. The amplifier is fitted with three fixing lugs and measures 190mm × 140mm. The relay switch is fitted with changeover contacts rated at 2A non-inductive 240V a.c.

EE 75 751 *for further details*

MINIATURE RELAY

B & R Relays Ltd, Temple Fields, Harlow, Essex

(Illustrated above right)



The D05/P-D55/P is a miniature plug-in relay with three changeover contacts, and it is designed for mounting on an 11 pin international base. An unusual design feature is the three peripherally arranged contacts, each contained in individual pockets, the latter giving improved insulation and preventing flash-over between contacts. Follow through is achieved by a rolling action of each contact on making. This results in a longer contact life due to reduction in mechanical wear.

The three changeover contacts are of silver, and each is rated at 6A 250V a.c. 30V d.c. Nominal coil power required is 2W or 4VA and the maximum d.c. coil resistance is 10 200Ω. The D05 is suitable for operation from voltages up to 170V d.c. and the D55 suitable for voltages up to 350V a.c. Weighing only 3¼ oz the relay can be fitted with a vacuum impregnated coil for tropical and high humidity conditions. It is supplied complete with a snap-on cover made from Makrolon, and the pin connexion diagram is engraved on the top of the cover.

The D05/P-D55/P is supplied in only one form and pin connexions cannot be varied. It measures 2 5/16in high × 2in × 1 9/16in across the base.

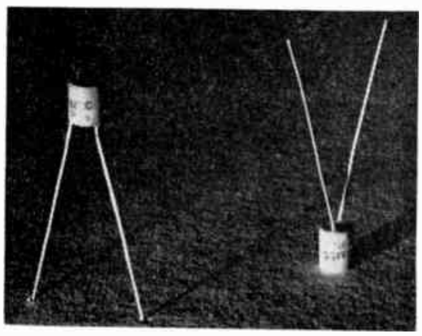
EE 75 752 *for further details*

WIRE WOUND RESISTORS

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

(Illustrated below)

Type P8S is a new addition to the range of encapsulated precision wire-wound resistors manufactured by Miniature Electronic Components Ltd.



It is 5/16in in height × 9/32in diameter with printed circuit leads to 0·2in spacing. This is the smallest resistor in the range, rated at 0·2W at 85°C. Temperature range is −65°C to +150°C.

The range includes 16 types with axial or printed circuit leads, or radial lugs. Resistance values up to 4·5MΩ are available, standard tolerances being 1 per cent 0·25 per cent and 0·1 per cent. Full Type Approval to DEF.5113 has been granted for types P12, P34 and P56.

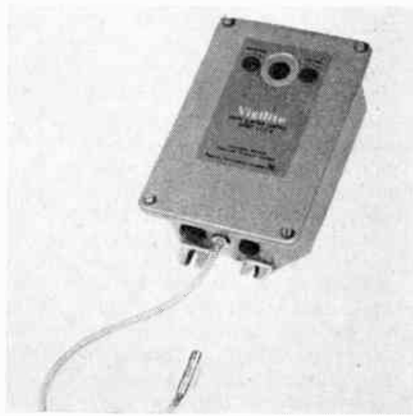
EE 75 753 *for further details*

PHOTO-ELECTRIC SWITCH

Lancashire Dynamo Electronic Products, Rugeley, Staffordshire

(Illustrated below)

A new photo-electric switching unit series LPC.2, the 'Vigilite,' has been produced by Lancashire Dynamo Electronic Products (M.I. Group).

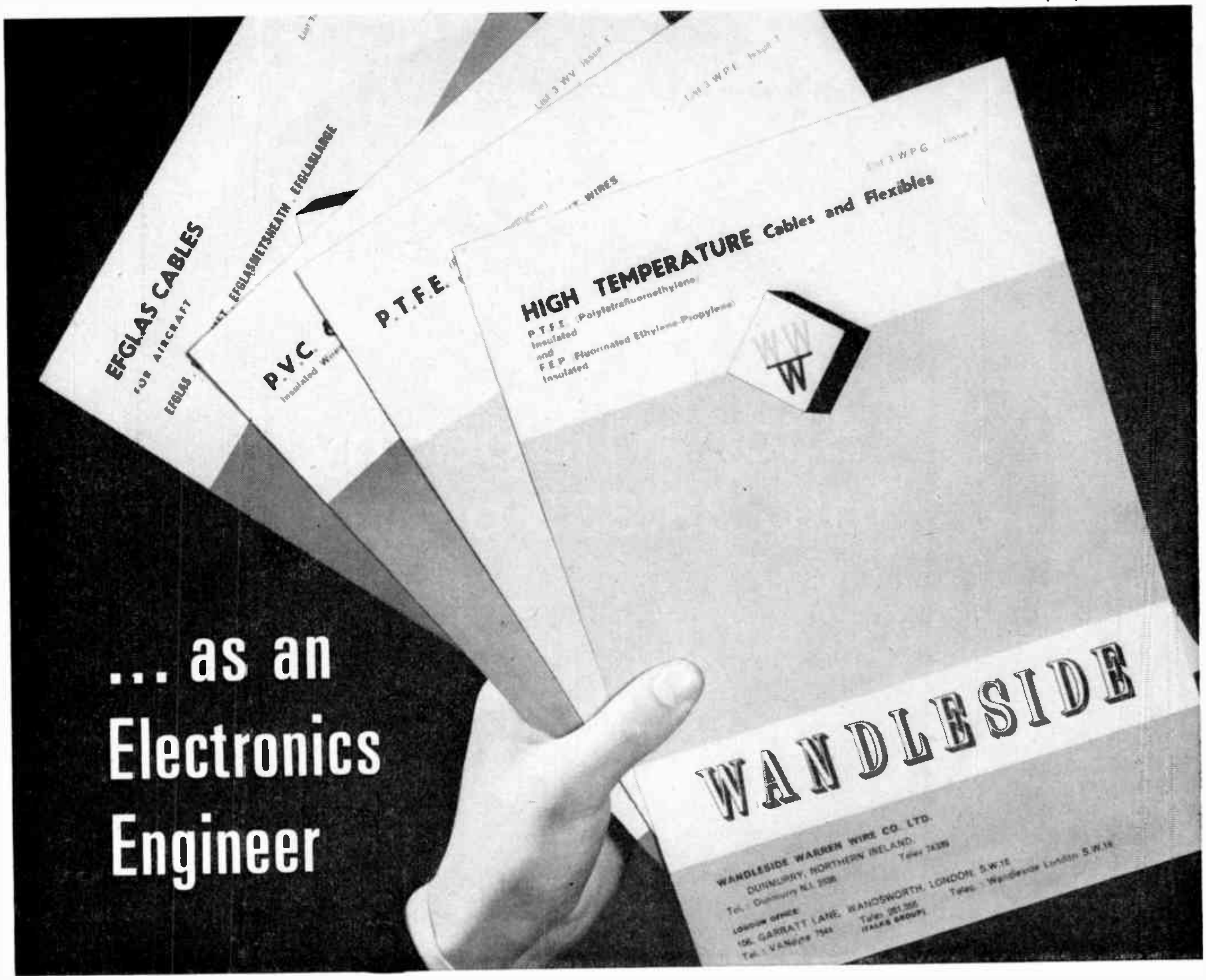


The unit, which is available in two forms—one with two remote heads and one with built-in cell and a single remote head—is suitable for indoor or outdoor use and controls all forms of counting, sorting, batching and detecting operations at a maximum rate of five operations per second.

The circuit is of fully transistorized printed card design, with the various components mounted directly on the card itself. The printed board, 3/32in thick, fits within the control case and is retained by screws top and bottom.

The case is of cast iron construction with a detachable front cover having a weatherproof sealing strip around the edges. Overall dimensions being 6¼in wide, 8¼in high and 3 1/16in deep. Provision is made for wall mounting by four projecting lugs.

The internal relay is fitted with one pair of normally open and one pair of normally closed contacts, which may be used for switching purposes, the rating being 5A, 230V or 1A at 440V a.c., non-inductive.



YOU SHOULD READ THESE BROCHURES

A COMPREHENSIVE RANGE OF CONNECTING WIRES FOR ELECTRONICS AND AIRCRAFT USE

HIGH TEMPERATURE Cables and Flexibles, P.T.F.E. (Polytetrafluoroethylene) Insulated and F.E.P. (Fluorinated Ethylene-Propylene) Insulated.

P.T.F.E. (Polytetrafluoroethylene) Insulated Equipment Wires to Specification E.L. 1930.

P.V.C. & POLYTHENE Insulated Wires and Flexibles.

EFGLAS CABLES For Aircraft, Efglas, Efglasmet, Efglasmetsheath, Efglaslarge.

LATEST INFORMATION!

P.T.F.E. Insulated Wires and Flexibles can now be supplied in **etched** form for those manufacturers having difficulties in bonding or potting.

WANDLESIDE WARREN WIRE COMPANY

In Association with **WANDLESIDE CABLE WORKS LIMITED**

Wandleside Warren Wire Co. Ltd., Dunmurry, N.Ireland. Phone: Dunmurry 2536. Telex: 74339. Grams: Derriaghy, Dunmurry, N.Ireland.
London Office: 106 Garratt Lane, S.W.18. Phone: Vandyke 7544. Telex: 261355. Grams: Wandleside, London, S.W.18.

By appropriate positioning of a pair of internal plug links, the unit can be arranged so that the relay is de-energized when light is established or when it is interrupted. In this manner either 'fail to light' or 'fail to dark' operation is afforded at the option of the user.

Preset sensitivity and differential controls are also provided.

The unit will operate from a 110 or 250V 50c/s single phase supplies.

EE 75 754 *for further details*

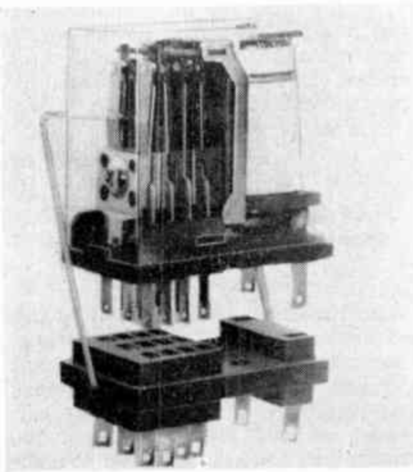
MINIATURE RELAY

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustrated below)

These low cost, miniature relays are available in two or four pole changeover versions with either gold-plated or silver contacts. They are supplied with dust-proof transparent covers and are mounted on plug-in bases.

The contacts are rated at 1A at 100V a.c. or 24V d.c. with coil voltages up to 85V d.c. The contact pressure is better than 10g.



EE 75 755 *for further details*

NOR SIMULATOR

Brensal Electronics Ltd, Charles Street, Bristol 1

(Illustrated above right)

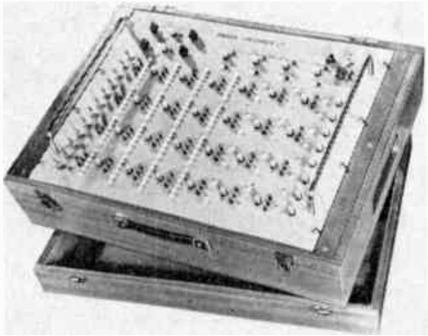
This self-contained unit is intended for use both in industry and technical colleges, where the demonstration or test of simple logic functions is required. All signal interconnexions are made on the front panel patch-board by means of the patch-cords which are provided with the equipment. In this way systems can easily be wired direct from logic diagrams.

The modules provided are as follows:

24 NOR elements 16 with 6 inputs and 8 with 2 inputs, all elements have 3 output sockets which can be used to drive 6 inputs if required.

2 timers variable delay over the range 0 to 60sec, two types of output are available depending on the input socket used.

4 binary counters connexions to input, set, reset and both Q and Q' outputs



are readily available on the patch-board.

12 toggle switches for use as input devices—these switches select either 0V or 24V as input signals to NOR elements.

8 output units provided to simulate output conditions, each stage having one input only.

Signal lamps are provided on all modules, the lamp being illuminated when output is at 0V.

Limited power is available for driving units external to the simulator.

The equipment operates from 200 to 250V a.c. mains. The simulator is mounted in a wooden cabinet with detachable lid, overall dimensions 26½in × 19in × 7½in.

EE 75 756 *for further details*

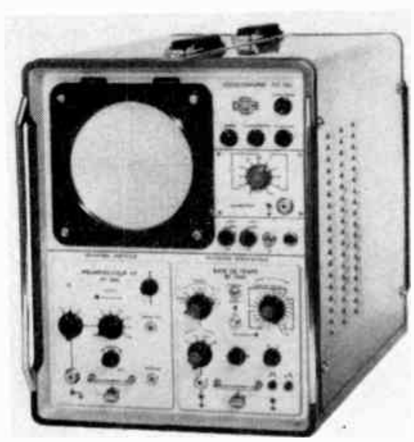
LARGE SCREEN OSCILLOSCOPE

Distributed by: Claude Lyons Ltd, Hoddesdon, Hertfordshire

(Illustrated below)

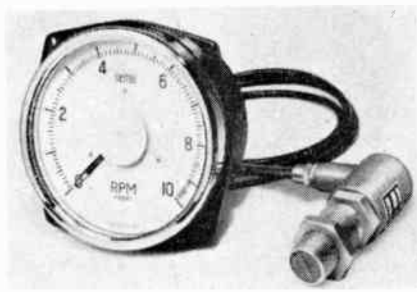
The gap between the small screen measuring oscilloscope and the television tube display oscilloscope is now filled by a range of five instruments manufactured by Constructions Radioelectriques et Electroniques du Centre (C.R.C.) and available through Claude Lyons Ltd, which provide all the features expected in a modern high-performance measuring oscilloscope and employ a specially developed 180mm (7in) screen cathode-ray tube to provide twice the effective display area of a 5in model.

A recent addition to the range is the type OC 746 large screen X-Y oscilloscope. This instrument has identical X and Y channels, each designed to accept any one of a variety of plug-in units. As an X-Y oscilloscope, two identical amplifier plug-ins (single trace, dual trace, differential or low level) are used. With a single amplifier plug-in and the time base unit type BT 7461 in the X channel, a versatile conventional oscilloscope results.



With the HF 5661 (single trace) and BT 7461 plug-ins, the OC 746 provides vertical sensitivity of 50mV/cm to 20V/cm from d.c. to 1Mc/s. Time-base range is 0·5μsec/cm to 2·5sec/cm with sweep expansion and full sync and trigger facilities. A voltage calibrator (in the main chassis) provides an accurate 1kc/s square wave at levels from 0·5mV to 100V. The BF 5662 plug-in provides differential input and sensitivities from 1mV/cm to 50V/cm, and the low-level unit BF 5672 sensitivities from 100μV/cm to 50V/cm, with differential input, while the dual-trace unit type CE 5673 provides chopped or alternate sweep presentations with sensitivity from 50mV/cm to 20V/cm.

EE 75 757 *for further details*



ELECTRONIC TACHOMETER

S. Smith & Sons (England) Ltd, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated above)

A new 6in industrial electronic tachometer is announced by Smiths Industrial Division. Two standard models with scale range 0 to 10 000rev/min and 0 to 15 000rev/min are available, but indicators with scale ranges of up to 1 000 000rev/min can be made to order.

This tachometer does not require mechanical coupling with any moving part and is ideal for applications where no mechanical take-off is available.

The instrument has a 270° circular scale moving coil indicator mounted in a case which also incorporates the transistorized drive circuit. Ferrous lobes on a rotating part of a machine passing in close proximity to a separately mounted magnetic perception head generate impulses which the transistorized circuit converts into an output proportional to their frequency. Thus the position of the pointer is dependent on the speed of the shaft to which the lobes are attached. This tachometer requires a 24V d.c. electricity supply, but as the maximum current consumption is only 100mA, primary cells may be used.

The ferrous lobes may be cast,

machined or brazed into a shaft, pulley, spur gear or flywheel. Perception heads are available for use with lobes mounted around the periphery or an alternative head can sense the rotation of a slotted bolt screwed into the centre of an exposed shaft end.

Where multiple indication is required any number of indicators may be operated from a single perception head. Multiple perception head installations with a single indicator are also possible using low contact resistance switches.

EE 75 758 *for further details*

TIME AND EVENTS RECORDERS

Bowmar Instrument Ltd, Sutherland Road, London, E.17

(Illustrated below)

Bowmar Instrument Ltd announce the production in this country of their elapsed time indicator type 1440 and events indicator type 1989.

High readability, four drum digital

read-out and miniature size make these units suitable for airborne equipment applications. The display is engraved with $\frac{1}{8}$in high white numerals on black drums, giving a total time of 9 999 hours. Overall diameter is 0·67in, length 1·8in and weight 1·8oz. Various types of mountings can be supplied including panel sealing versions. Power consumption is 1·1W on 115V 400c/s single phase supply. Invertors are available to enable the units to work from 28V d.c. and are approximately 1in × 1in × 1$\frac{1}{2}$in in size.

The events indicator is housed in a similar case and is capable of 9 999 counts. Input is either 24 to 28V d.c. at 2W or 115V 400c/s at 2W. Maximum counting rate: 10 per second with pulse length of 50msec.

Both the above units are non resettable and are designed to meet the Specification MIL-M-7793C.

EE 75 759 *for further details*

OVERLOAD PROTECTION THERMISTOR

Standard Telephones & Cables Ltd, Connaught House, 63 Aldwych, London, W.C.2

The first of a new STC range of positive temperature coefficient thermistors is now available. This is the PTC 120, which has been specifically designed for the protection of motors, generators, transformers and other equipment against excessive temperatures which might cause damage to the winding insulation. Nominal theshold temperature of the device is 120°C.

The PTC 120 thermistor element is only 0·186in in diameter. The outer casing of this small device is tumbler de-burred so that it can be included in windings without any risk of removal of enamel insulation from the coils.

The PTC leads are arranged as a twisted pair and the standard length is 24in. Other lengths can be supplied, and, for three-phase working, multiple arrangements of the PTC 120 are available.

EE 75 760 *for further details*

MINIATURE SOLDERING IRON

Antex Ltd, Grosvenor House, Croydon, Surrey

(Illustrated below)

The latest addition to the Antex range is the model C240N. This is a miniature mains voltage soldering iron weighing between 2 and 3oz. It is fitted with a 'Ferraclad' bit which is claimed to last at least five times longer than ordinary

nickel plated bits and to provide greater heat capacity and retention. The bit can be replaced easily without any danger of damaging the iron. The tool is robustly constructed and is suitable for continuous use.

EE 75 761 *for further details*

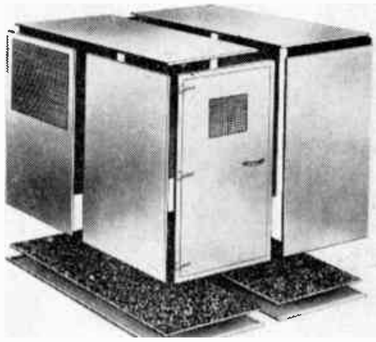
SHIELDED ROOM KITS

Belling & Lee Ltd, Great Cambridge Road, Enfield, Middlesex

(Illustrated below)

Now available from Belling-Lee in addition to its range of screened compartments is the "100" series of modular enclosures, offering all the features normally found only in custom-built installations.

The new series is based on a range of interchangeable metal framed subunits, which can be quickly assembled to provide a shielded room with attenuation of up to 100dB at frequencies between 1 and 500Mc/s.



Features of the series include special conducting gaskets for efficient r.f. bonding of the panels, and a new design of push-fitting door which is self-bonding and holding.

Modular window panels can be supplied in metal mesh, or in a honeycomb form comprising a layer of r.f. shielding ducts. This operates on the principle of 'waveguides below cut-off', giving a high degree of attenuation to r.f. below the cut-off frequency, but allowing good visibility in the forward direction and a free passage of ventilating air.

A heavy-duty feed-through panel, operating in conjunction with appropriate filters, accommodates electricity power supplies and telephone and other services.

A major advantage of the new series is flexibility of room size and layout, as enclosures can easily be modified or, extended at any time. They can also be taken down and stored compactly or transported to another site. The material is supplied in kit form, with the subunits, linoleum-tiled wooden floor panels, and all nuts, bolts and r.f. gasketing required. The unit panels are 7ft 4in high × 3ft 8in wide, and the doors and services feed-through panels are all of this size module. Half modules, work benches, etc., are available for specialized applications, and there is a comprehensive range of electrical filters.

EE 75 762 *for further details*

SELF-BALANCING RECORDER

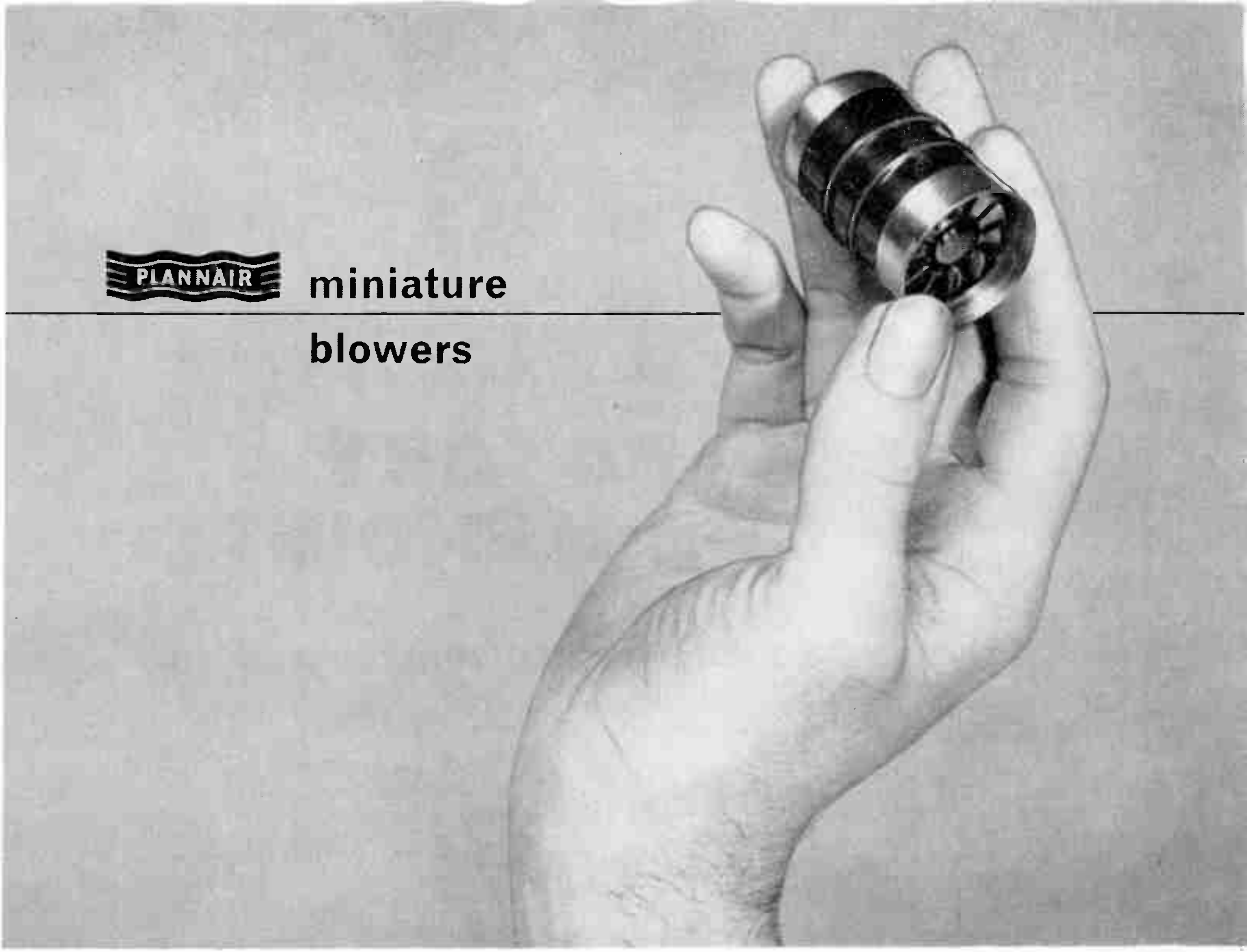
Distributed by: Wessex Electronics Ltd, Midsomer Norton, Somerset

(Illustrated below)

The Latronics Recordette-4 is a high performance, self-balancing potentiometer recorder, designed for versatile operation. This instrument offers small size, portability, a wide range of chart speeds, ink and inkless recording and many other features including accuracy $\frac{1}{2}$ per cent of span, with sensitivity of $\frac{1}{4}$ per cent of span, full servo performance $\frac{3}{4}$sec full-scale balancing time.

The type 1 and 2 input units provide an electrical span adjustable between 10 and 100mV full scale. Type 3 input unit provides an adjustable span between 10mV and 100V. Type 4 input unit is designed for thermocouple inputs, providing direct recording of temperature.

The Recordette-4 has been designed for reliable, long term, operation. Direct access to chart drive, the pen, the input



EE 75 146 *for further details*

Designing new equipment?

you'll need this guide to aero-thermal control

With increasing miniaturisation, temperature control is becoming more and more critical. To meet this need Plannair has designed a whole range of miniature blowers offering the best combination of size, weight and performance.
Sixteen of these blowers (including the tiny $2\frac{1}{2}$ oz Thimble blower, which is only 1.85 in. long and 1.13 in. dia.)— out of a total range of over 1,000 designs — have been featured in a new publication, 'Plannair Miniature Blowers'. Full dimensions, and operating and performance data are given. Send for your copy by using the reader reply service or this coupon.
Note: Air movement and temperature control problems are often more than a matter of just fitting a blower. Plannair's aero-thermal control specialists will be pleased to come and discuss your specific problem with you right from the planning stage. Many manufacturers already take advantage of this service and PLAN WITH PLANNAIR.

To Plannair Ltd. Windfield House, Leatherhead, Surrey. Telephone: Leatherhead 5341-50

Please send me a copy of 'Plannair Miniature Blowers'

Name ______________________

Position ______________________

Company ______________________

Address ______________________

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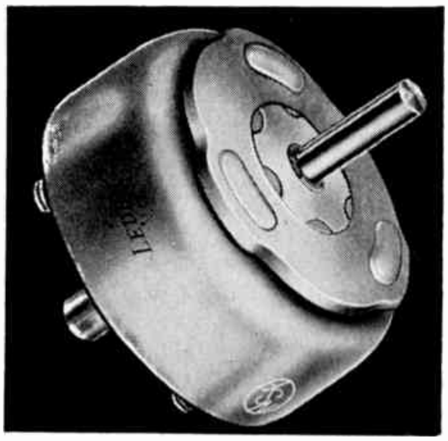
PLANNAIR specialists in aero-thermal control

PLA 15

EE 75 147 *for further details*

LEDEX



ROTARY SOLENOIDS

A solution to many remote control problems

OTHER NSF PRODUCTS

- 'Oak' MULTI-BANK ROTARY WAFER SWITCHES
- 'Cutler-Hammer' APPLIANCE SWITCHES
  Licensees of Cutler-Hammer Inc. Milwaukee, U.S.A.
- 'SS-Union' MINIATURE HERMETICALLY SEALED RELAYS
  Licensees of Westinghouse Brake & Signal Co. Ltd., London
- 'Centralab' SMALL CARBON TRACK POTENTIOMETERS
  Licensees of Centralab Inc., Milwaukee, U.S.A.

- Frictionless, snap action
- Up to 98 lb/in. torque
- Rotary strokes of 25° to 95°, clockwise or anti-clockwise
- Substantially linear and level work-to-rotation curve

Enquiries invited for Solenoids with special torque-to-stroke characteristics.

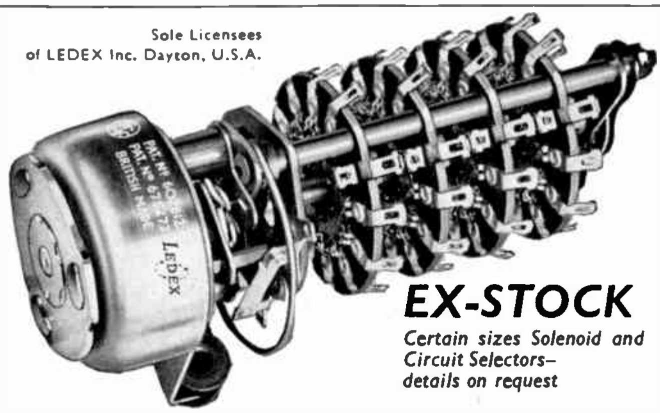
LEDEX

CIRCUIT SELECTORS

—combinations of Solenoid and Rotary Wafer Switch sections, permitting a versatility of complicated circuit switching with innumerable applications.

TECHNICAL BROCHURE ON REQUEST

Sole Licensees of LEDEX Inc. Dayton, U.S.A.



EX-STOCK

Certain sizes Solenoid and Circuit Selectors– details on request

NSF LIMITED

31-32, ALFRED PLACE, LONDON, W.C.1

Telephone: LANgham 9561 Telegrams: ENESEF TELEX LONDON Telex: 21907

WORKS: KEIGHLEY, YORKSHIRE

Telephone: Keighley 4221 Telegrams: ENESEF TELEX KEIGHLEY Telex: 51270

A MEMBER OF THE Simms GROUP OF COMPANIES unit and all controls makes the operating procedure extremely straightforward.

Chart speds from $\frac{1}{4}$in/h to 48in/min are available, the 4in paper may be fed on to the take-up roll or may be allowed to feed out through a slot at the bottom of the door for instant reference and tear-off. The panel-mounted version requires a panel cut out of $6\frac{3}{4}$in × $8\frac{3}{4}$in.

EE 75 763 *for further details*

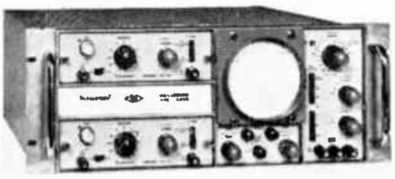
LABORATORY OSCILLOSCOPE

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(Illustrated below)

Telequipment Ltd has introduced the D43R, a 19in rack mounting oscilloscope occupying only 7in of vertical panel space, and having a depth of 16in.

The D43R, designed for a wide range of industrial, laboratory and medical electronic applications, is available with a choice of five Telequipment plug-in amplifiers, two of which are new.



With a 4in flat-faced p.d.a. tube operated at 4kV, the D43R has a timebase providing 18 pre-set calibrated sweep speeds from 0·5sec/cm to 1μsec/cm. Rise time is 23nsec, input impedance 1MΩ, and the unit operates on 100 to 240V (50 to 100c/s) power supplies.

Plug-in amplifiers available are types A, B and C (general purpose, differential and ultra high gain), in addition to the new units, D and G. Type D, an envelope monitor, is a tuned amplifier with four switched ranges from 2·5 to 32Mc/s. When tuned to resonance, sensitivity is approximately 1V/cm, modulation frequency is available as a sync output, and input impedance is 50Ω.

Type G is a combined wideband and differential amplifier, the differential inputs having a rejection greater than 200 : 1 from d.c. to 1Mc/s (sine wave input), reducing to not less than 30 : 1 at 12Mc/s. Maximum input voltage 10V peak-to-peak (on 10mV/cm range).

EE 75 764 *for further details*

DIGITAL VOLTMETER-RATIOMETER

Digital Instruments Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated above right)

Digital Measurements Ltd has recently added the DM2022 digital voltmeter to its wide range of digital instrumentation. This new instrument has the accuracy and reliability of the DM2020 voltmeter but has a longer scale (39999) and built-in ratiometer facilities.

The new instrument has an accuracy of 0·0025 per cent f.s.d. ±0·01 per cent of reading and the high resolution of 1 part in 40 000. There are five ranges covering from 0 to 2kV; the sensitivity on the lowest range being 10μV. The display can be scaled externally to give direct readings in lb/in², °C, etc from voltage analogue inputs. The input impedance is greater than 25 000MΩ on the two lowest ranges and 10MΩ on the higher ranges. The input can be isolated from earth to reject common mode voltages present at the signal source.

The ratiometer facility is an important feature of the instrument in that it enables comparison measurements to be made between voltage ratios and external references, e.g., potentiometer calibration, analogue computer measurements, etc.

The DM2022 incorporates the wide range of facilities featured in the earlier DM instruments. These include full accuracy maximum and minimum operating modes and decade outputs in any one of six codes to drive printers, punches, etc.

EE 75 765 *for further details*

PROXIMITY SWITCHES

W. H. Sanders (Electronics) Ltd, Gunnels Wood Road, Stevenage, Hertfordshire

(Illustrated below)

These units have been supplied to the U.K.A.E.A. for some time and are now available to industry generally. The switch action relies upon an external ferrous material to complete a magnetic circuit and can be used to detect the presence of a workpiece or machine part without contact being made. Such a requirement exists where the use of conventional switches is impractical, for example when corrosive or inflammable vapour, dust or liquid is present or where radio-active materials are being handled.

The proximity switch operates when ferrous material is within 0·5in of the pick-up head, the actual distance at which the switch operates can be adjusted by means of a sensitivity potentiometer. The pick-up head is completely impregnated and encapsulated in epoxy-resin and is unaffected by the conditions met within most industrial processes.

The two versions of the proximity switch available are: Type FSPS/042 and the smaller model PDM/3173, which is suitable for use with a bistable transistor amplifier. The smaller unit is designed for assembly within existing transistor logic systems, but can be used separately as a normal switching element.

EE 75 766 *for further details*

WAVEFORM GENERATOR

Servomex Controls Ltd, Crowborough, Sussex

(Illustrated below)

As a development of the LF 51 low frequency waveform generator, Servomex Controls Ltd has designed a new instrument which is capable of producing 90 distinct waveforms in the audio and servo frequency ranges and yet is small and light enough to be carried in one hand. This notable advance in design of this type of instrument has been achieved by the extensive use of transistors and printed circuits.

The new generator, type LF 141, has a frequency range of 0·002c/s to 2kc/s and produces all the 90 waveforms without any ancillary equipment whatever. The output voltage is zero to ten volts peak in three ranges with maximum peak voltage of 10V, 1V and 0·1V.

The instrument has a direct frequency dial and the output voltage can be either balanced about the zero line or polarized. This produces additional waveforms and some delayed waveforms. A special feature is that it can be keyed; the application of an external paralysing voltage will stop the main generator, allowing one or more cycles to appear.

A further feature of this instrument is the completely separate auxiliary oscillator for triggering the main set, by means of which a large variety of new waveforms can be produced. Because the auxiliary oscillator generates a ramp as

well as the sharp triggering pulse, it is possible to synchronize an oscilloscope which has adequate d.c. trigger level controls to see the start of the main wave. This is equivalent to having a low frequency delay line built into the c.r.o., with a maximum delay of several seconds.

The basic integrator is of a new design which eliminates the delay usually experienced with this type of instrument after applying a trigger pulse. The use of a variable phase unit will produce an improved sine wave owing to the absence of flat tops on the triangular wave of the basic generator. With this unit the wave can be started at any required phase angle.

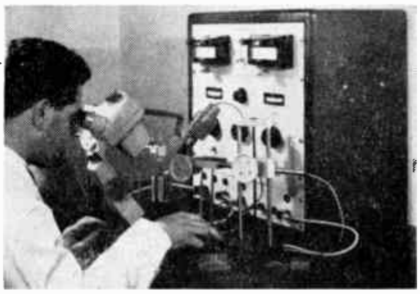
EE 75 767 *for further details*

THERMOCOMPRESSION BONDER

G. V. Planer Ltd, Windmill Road, Sunbury-on-Thames, Middlesex

(Illustrated below)

This equipment, recently introduced by G. V. Planer Ltd, is designed as a versatile tool for the production of thermocompression bonds in research and manufacture. It is particularly suited for bonding leads to thin film microcircuits, as well as to semiconductor materials and devices, etc.



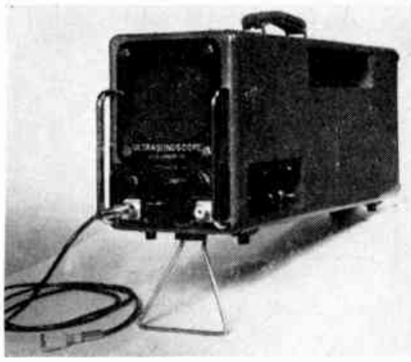
Special features of the apparatus are the new precision micromanipulators with *X*, *Y* and *Z* motion and special double-stem 'chessman' controls, independent thermostatic control of the bonding chisel and substage, and the flexible configuration of micromanipulator and substage positions on the work desk.

Two micro-manipulators are comprised in the equipment, one carrying the thermostatically controlled bonding chisel with variable load, the other the compressed-air assisted wire feed appliance. The latter is designed to accept standard spools, being suitable for wires of between 0·0005in and 0·01in in diameter. A wire cutter is mounted at the nozzle of the feed mechanism. The bonding chisel and substage can be maintained at temperatures of up to 600°C, with provision for the issue of inert gas over the working area, if required.

A stereozoom binocular microscope with continuously variable magnification of 7 to 30 or 14 to 60 and adjustable illumination is incorporated.

The operating components and the control console are mounted on a steel desk with a surface area of 2ft 6in × 4ft.

EE 75 768 *for further details*



ULTRASONIC FLAW DETECTOR

Ultrasonoscope Co. (London) Ltd, Sudbourne Road, Brixton Hill, London, S.W.2

(Illustrated above)

This new ultrasonic flaw detector, known as the Mark 5, is fully transistorized and has a 5in. c.r.t. which provides for accurate thickness measurement and viewing under site condition.

The switched frequencies are 5, 2½ and 1¼Mc/s and the gain is controlled with a calibrated attenuator. The fastest time-base speed is 2μsec and this means that ¼in of steel may be expanded across the full screen. The time-base has a continuously variable delay and the maximum range is 20ft in steel. The pulse repetition frequency is variable from 50 to 1 000c/s for the elimination of 'ghost echoes'. Scatter from the structure of a metal can be 'grass-cut' with a reject control with switched steps. Either rectified or unrectified display may be selected and the unrectified presentation has proved essential for a number of difficult problems.

The most important controls are on the front panel and the remainder are grouped in order of use along one side. The re-chargeable batteries have a life of 7 hours and are instantly interchangeable. The Mark V can, if preferred, be operated on a normal mains supply by using a special mains pack in the battery compartment. The overall dimensions are 9¼in × 7in × 21½in and the weight is 28lb including batteries.

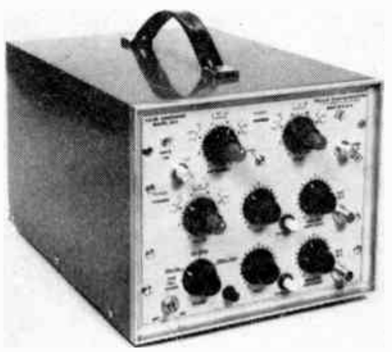
EE 75 769 *for further details*

PULSE GENERATOR

Distributed by: B. & K. Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Illustrated below)

The Texas Instruments model 6613 general purpose pulse generator provides coincident positive and negative output pulses at a rate determined by an internal clock generator, an external signal, or from a manual single-shot switch. Pulse width is variable up to 90 per cent duty cycle. Delay may be adjusted for output before or after sync output. Output amplitude may be varied from zero to 10V with output overload protection provided. Printed circuit board construction is used throughout, and the complete unit measures only 8½in × 8½in × 12in and weighs 10lb.



Clock pulse repetition frequency can be varied in 6 decade stepped ranges from 15c/s to 15Mc/s and amplitude from 0 to 10V into 50Ω. Rise and fall times can be varied from 10nsec to 10msec in 3 stepped ranges. Delay can be varied from 30nsec to 30msec in 6 decade stepped ranges. Width can be varied from 30nsec to 30msec in 6 decade stepped ranges. Sync output is 3V positive, rise and fall times 15nsec and duration 50 to 80 per cent duty cycle depending on frequency control setting. The external drive input requires a minimum of 2V positive, with 20nsec rise and 30nsec minimum duration.

A single pulse can be provided by a front panel push-button control, and a front panel switch allows gating of output pulses, 'off' at zero volts and 'on' at 7V positive.

EE 75 770 *for further details*

COUNT-DOWN RELAY

Distributed by: D. Robinson & Co. Ltd, 5-7 Church Road, Richmond, Surrey

(Illustrated on page 791)

The Rodene count-down relay is a device for counting any number of a.c. or d.c. pulses. Each incoming pulse energizes a solenoid which indexes a ratchet wheel one position. When the preset number has been reached, a micro-switch is operated which re-directs any further pulse to the separate resetting coil (if fitted) or to the user's external circuit.

Each 'count-down' relay can be set to count any number up to 20. Two or more relays can be cascade connected such that the resetting pulse of relay A operates the stepping coil of relay B to give decade counting.

'Count-down' relays are available for either auto-reset or manual reset applications. Auto-resetting types have a pointer which remains at the preset position during timing. The pointer of the manual reset types indexes at the same rate as the ratchet wheel, thus giving visual indication at all times of the number of pulses received.

Auto-resetting is effected by a single pulse to a separate coil which releases a ratchet and allows a return spring to reset the relay within 50msec.

One or two switch banks can be fitted instead of or in addition to the micro-switch and can be used to initiate a series of external circuits throughout the entire cycle. Any pulse length over 25msec will operate the relay, but as in all relays dissipation is high to give quick operation and, therefore, the coil should not

EE 75 148 *for further details*

NEW

PHILIPS

hf double beam oscilloscope

PM 3230

HIGH SENSITIVITY	LARGE BANDWIDTH
20 mV/div	0 - 10 Mc/s
2 mV/div	0 - 2 Mc/s
(1 division = 8 mm)	

Clear, sharp picture provided by new "side-by-side" double-gun 10 cm (4 in) cathode ray tube with full scale horizontal and vertical deflection of both beams.

Clear, logical lay-out of front plate, resulting in quick and easy operation.

19 calibrated sweep speeds from 0.5 μs/div to 0.5 s/div, sweep expansion up to 5×

Jitter free triggering with simple controls stability being preset, special TV frame position

Largely transistorized

Compact and lightweight (11 kg - 24 lbs)

Full set of accessories available

Complete manual with full service instructions supplied with the instrument

Sold and serviced all over the world

Compartment for mains flex provided

Transparant dust cover with pockets for all accessories provided

Low power consumption (70 W) and wide range of supply frequencies (50 - 400 c/s), making it suitable for use with an inverter

Facility for photographic recording, three preset levels of graticule illumination for simple camera setting

All components easily accessible for maintenance and repair

PHILIPS

PHILIPS electronic measuring instruments

For the U.K.:
The M.E.L. Equipment Company Ltd.,
207 Kings Cross Road, London WC1

be energized continuously. Maximum stepping rate is 5 steps per second and the life of the relay is in the order of 3 million operations. All models are available with 'AMP-Faston' connectors.

EE 75 771 *for further details*

FREQUENCY STANDARD

General Radio Co. (U.K.) Ltd, Marlow Road, Bourne End, Buckinghamshire

(Illustrated below)

The type 1115-B standard-frequency oscillator is the latest in General Radio's series of quartz-crystal frequency standards. The new instrument uses a 5Mc/s, fifth-overtone gettered crystal in a three-stage transistor oscillator. Crystal, oscillator, and a.g.c. circuits are all enclosed in a proportional-control oven. The oscillator frequency is divided to supply outputs at 1Mc/s and 100kc/s as well as at 5Mc/s.

Short-term stability (deviation at 95 per cent confidence limits) is 1 part in 10^{11} (r.m.s.) over a 10sec averaging time, 100 parts in 10^{11} (r.m.s.) over 300μsec. Typical ageing is less than 1 part in 10^{10} per day after one year. Spectral line width is under 0·25c/s at X band. Noise pedestal is 145dB down for a 1c/s bandwidth at the 5Mc/s output.

The new standard uses all silicon solid-state active components, in a package that is specially designed and constructed for mobile and severe-duty applications. A built-in nickel-cadmium-battery supply takes over automatically and operates the standard for 35 hours upon a.c.-main failure.

EE 75 772 *for further details*

CLIMATIC CABINET

Fisons Scientific Apparatus Ltd, Loughborough, Leicestershire

The Weyco Division of Fisons Scientific Apparatus Ltd has recently added a new climatic cabinet to its range of environmental test equipment.

This new cabinet, known as the CM/49, has the following features:-

Temperature Range −25°C to 150°C controlled to ±1°C.

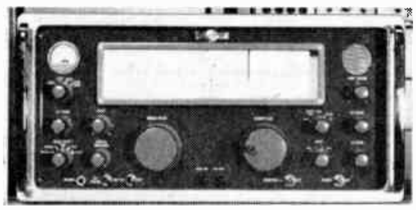
Humidity Range, 12 per cent to maximum controlled to ±1 per cent r.h.

Capacity of Working Chamber 9·4ft³.

A patented airflow system ensures even distribution of conditioned air throughout the working chamber. Humidity is introduced by injection of atomized water. A hermetically sealed refrigeration unit is incorporated which not only enables operation at low temperatures but also makes it possible to achieve humidities lower than ambient. All controls are mounted at the front of the cabinet for ease of operation and access to all electrical components is gained by lifting the hinged control panel. A twin pen recorder to give continuous recording of temperature and humidity is fitted as standard.

The new cabinet will meet most of the needs of environmental engineers and quality control personnel in the packaging, pharmaceutical paint and chemical industries.

EE 75 773 *for further details*



HIGH STABILITY H.F. RECEIVER

The Marconi Co. Ltd, Chelmsford, Essex

(Illustrated above)

A high stability, general purpose, h.f. receiver, type H2301, covering the frequency band 500kc/s to 30·5Mc/s is now available from The Marconi Co. Ltd. This economic and compact receiver is suitable for reception of a.m., c.w. and s.s.b. signals, and with the addition of a special f.s.k. adaptor, type H5011, it can be used with single channel teleprinter circuits.

The frequency range is covered in thirty steps of 1Mc/s with an overlap of 100kc/s and the calibration accuracy is within 1kc/s. The dial calibration is linear and presented in such a manner that the frequency is given by combining the readings on separate 'Mc/s' and 'kc/s' scales. Calibration can be checked at any time with a built-in 100kc/s crystal oscillator.

A high selectivity is achieved by means of a double conversion technique, employed with crystal control of the first local oscillator to give an exceptionally high stability for this class of receiver.

The second i.f. operates at 500kc/s and a cathode-follower provides an output at this frequency for connexion to ancillary equipment. Five positions of selectivity are available, two of which employ band-pass filters. An a.f. filter is fitted for selective c.w. reception.

All panel controls are conveniently positioned for ease of operation and the receiver incorporates a built-in loudspeaker. Separate audio outputs are available with independent gain controls.

High-quality components are used throughout the receiver which is of rugged construction, being made up of six sub-chassis and two double screened boxes bolted together. It is suitable for either rack or bench mounting and will operate from any normal a.c. mains supply.

The f.s.k. adaptor is fully transistorized and caters for shifts of from 200 to 850c/s, at telegraph speeds of up to 100 bauds.

The Marconi H2301 receiver was originally an Eddystone receiver, but has been modified to meet a Marconi specification for s.s.b. and f.s.k. operation.

EE 75 774 *for further details*

MULTI-RANGE METER

Oxley Developments Co. Ltd, Priory Park Ulverston, Lancashire

(Illustrated below)

Recently introduced to the British market by Oxley Developments Co. Ltd, the 'MONOC' is a pocket-sized multi-range meter of unusual design.

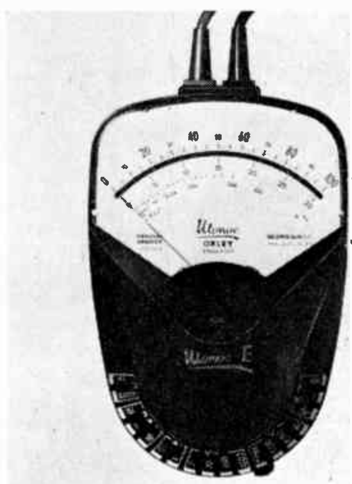
With a d.c. accuracy of 1·5 per cent and an a.c. accuracy 2·5 per cent it covers ranges up to 1000V d.c. and a.c. Current ranges of up to 1A d.c. and 10A a.c. are also provided. An ohmeter that requires no adjustment will make measurements up to 2MΩ and there are many accessories for extending the ranges.

Due to the original design it has been possible to provide a mirror scale instrument with a 3·6in scale in a case only 6in by 4in by 2in (approximately).

The movement, which is well protected by both diodes and fuses, is practically indestructible and due to its careful balancing the instrument can be used in any position.

An elastic bracelet allows full advantage to be taken of the light design when the instrument can be used on the wrist away from a bench or other flat surface, and leaving both hands free.

The single range selector switch allows easy operation with one hand, which adds to the convenience of the instrument.



EE 75 775 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

Unless otherwise stated, all meetings will be held at Savoy Place, commencing at 5.30 p.m.

Ordinary Meetings

Date: 5 November Time: 5.30 p.m.
Lecture: Mercury Arc Valves for H.V. D.C. Transmission
By U: Lamm

I.E.E./R.Ae.S. London Joint Group on the Applications of Electricity in Aircraft

Date: 24 November Time: 6 p.m.
Discussion: Unconventional Methods of Generating Electrical Power in Aircraft of the Future
Opened by: M. W. Thring

Electronics Division

Date: 2 November
Lecture: N.A.S.A. Research and Technology Programmes in Space Navigation
(Joint meeting with Royal Aeronautical Society)
Date: 3 November
Lecture: The Problem of Vestigial-Sideband Transmission and Reception
By: W. Wharton and B. J. Rogers
Date: 4 November
Lecture: A Push-Pull Parallel Aerial System for Navigation and Application to Communication Problems with Simple Methods of Side-Lobe Cancellation
By: E. O. Willoughby
Date: 9 November
Discussion: Submarine Telephone Cables—Protection, Damage and Repair
Opened by: R. J. Halsey and N. E. Holmblad
Date: 12 November
Lecture: Pressure Sensitive Effects in Semiconductors
By: K. Preece, P. Lundberg, P. Selway and V.G. Tull
Date: 18 November
Discussion: What is Measurements?
(Joint meeting with the Science and General Division)
Opened by: L. Finkelstein. G. G. Gouriet, P. Vigoureux and A. C. Lynch
Date: 23 November
Lecture: Loop Gain and Return Difference of Transistor Amplifiers
By: R. F. Hoskins
Date: 25 November
Discussion: Symbols for Logic Circuits
(Joint meeting with the I.E.R.E. Computer Group at the London School of Hygiene and Tropical Medicine)
Opened by: J. A. Lawrence and J. H. Smith
Date: 30 November Time: 2.30 p.m.
Discussion: Lasers
(Joint meeting with the I.E.R.E. Medical and Biological Group)

Power Division

Date: 2 November
Lecture: Modern Introduction to the General Theory of Electrical Machines
By: A. J. Ellison
Date: 4 November
Discussion: The Training of the Industrial Engineer
Opened by: C. J. Foster
Date: 12 November
Discussion: Current Balance Earth-Leakage Circuit-Breakers—Design Features and Future Employment
Opened by: E. J. Sutton
Date: 13 November
Lecture: Electromagnetic Theory of Electrical Machines
By: S. A. Nasar
Date: 17 November
Lecture: Adjustable Frequency Invertors and their Applications to Variable-Speed Drives
By: D. A. Bradley. C. D. Clarke, R. M. Davis and D. A. Jones
Date: 19 to 20 November
Conference: Commutation in Rotating Machines (All wishing to attend must register: forms available on application)
Date: 25 November
Lecture: Marine Electric Propulsion
By: D. St. J. Seigne

Science and General Division

Date: 2 November
Lecture: Modern Introduction to the General Theory of Electrical Machines
By: A. J. Ellison
Date: 4 November
Discussion: The Training of the Industrial Engineer
Opened by: C. J. Foster
Date: 10 November
Lecture: Computer Aided Study of Character Recognition
By: J. A. Weaver
Date: 18 November
Discussion: What is Measurement?
Opened by: L. Finkelstein, G. G. Gouriet, P. Vigoureux and A. C. Lynch
Date: 26 November
Colloquium: Plasma Jets

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All meetings at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, unless otherwise stated.

Television Group

Date: 4 November Time: 6 p.m.
Lecture: U.H.F. Television Reception over Long Distance Paths
By: B. W. Osborne

Electro-acoustics Group

Date: 11 November Time: 6 p.m.
Papers on: Electro-mechanical Filters
Contributions by: M. Borner and A. Russen

Radar Group

Date: 18 November Time: 6 p.m.
Lecture: An Air Surveillance Radar System
By: R. L. Burr and J. Flounders

Joint I.E.R.E.-I.E.E. Computer Groups

Date: 25 November Time: 6 p.m.
Discussion: Logic Circuit Symbols
Opened by; J. A. Lawrence and J. H. Smith

Joint I.E.R.E.-I.E.E. Medical Electronics Group

Date: 30 November Time: 5.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.1.
Discussion: Medicial Applications of Lasers

Southern Section

Date: 26 November Time: 7.30 p.m.
Held at: Basingstoke Technical College, Basingstoke
Lecture: Colour Television Transmission Systems
By: W. Wharton
Date: 4 November Time: 6.30 p.m.
Held at: Highbury Technical College, Cosham, Portsmouth
Lecture: The Application of Radar and other Electronic Techniques to Meteorology
By: W. A. Grinstead and A. P. Tuthill

South Western Section

Date: 18 November Time: 7 p.m.
Held at: The University of Bristol Engineering Laboratories, Bristol
Lecture: Quality and Reliability
(Joint meeting with the Institution of Production Engineers)
By: D. J. Hewitt
Date: 19 November Time: 7 p.m.
Held at: Plymouth College of Technology, Plymouth
Lecture: Ballistic Missile Early Warning System (Joint meeting with the Western Centre of the I.E.E.)
By: B. Batt

South Wales Section

Date: 11 November Time: 6.30 p.m.
Held at: The Welsh College of Advanced Technology, Cardiff
Lecture: Automobile Electronics
By: R. A. Evans

Scottish Section

Date: 11 November Time: 7 p.m.
Held at: The Department of Natural Philosophy, The University, Drummond Street, Edinburgh
Lecture: Signal Processing Filters and Networks (Joint meeting with Scottish Centre, Electronics and Measurements Section, I.E.E.)
By: D. J. H. MacLean
Date: 12 November Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow
Lecture: Signal Processing Filter and Networks (Joint meeting with Scottish Centre, Electronics and Measurement Section I.E.E.)
By: D. J. H. MacLean

South Midland Section

Date: 26 November Time: 7 p.m.
Held at: B.B.C. Club, Evesham
Lecture: U.H.F. Broadcasting and B.B.C.2
By: D. B. Weigall

East Midland Section

Date: 12 November Time: 6.30 p.m.
Held at: University of Leicester, Leicester
Lecture: Education and Training for Professional Radio and Electronic Engineers
By: A. J. Kenward

North Western Section

Date: 4 November Time: 6 p.m.
Held at: Reynold Building. Manchester College of Science and Technology, Manchester
Lecture: Television Receiving Aerials at U.H.F. (Joint meeting with the North Western Centre of the Electronics Section of the I.E.E.)
By: C. F. Whitbread

Yorkshire Section

Date: 4 November Time: 6.30 p.m.
Held at: Department of Electrical Engineering, University of Sheffield, Sheffield
Lecture: Environmental Testing of Electronic Equipment
(Joint meeting with Sheffield Sub-centre, Electronics Section of the I.E.E.)

THE RADAR AND ELECTRONICS ASSOCIATION

Date: 12 November Time: 7 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2
Lecture: Televising the Tokyo Olympics
By: L. F. Matthews

THE TELEVISION SOCIETY

Date: 6 November Time: 7 p.m.
Held at: The Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3
Lecture: Factors Affecting the Acceptability of Colour Reproduction
By: R. W. G. Hunt
Date: 19 November Time: 7 p.m.
Held at: The Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3
Lecture: A British Video Tape Recorder
By: J. L. E. Baldwin

PUBLICATIONS RECEIVED

FLOWMETER CALIBRATION — Rotameter Manufacturing Co. Ltd (a member of the Elliott-Automation Group) of Purley Way, Croydon, have produced a 20-page booklet on a new method of calibrating their 'Metric' Series Rotameters without the customer needing to return the instrument to the factory. Copies of the booklet, known as publication RP.3000, are available free of charge from the Publications Department, Rotameter Manufacturing Co. Ltd, 330 Purley Way, Croydon, Surrey.

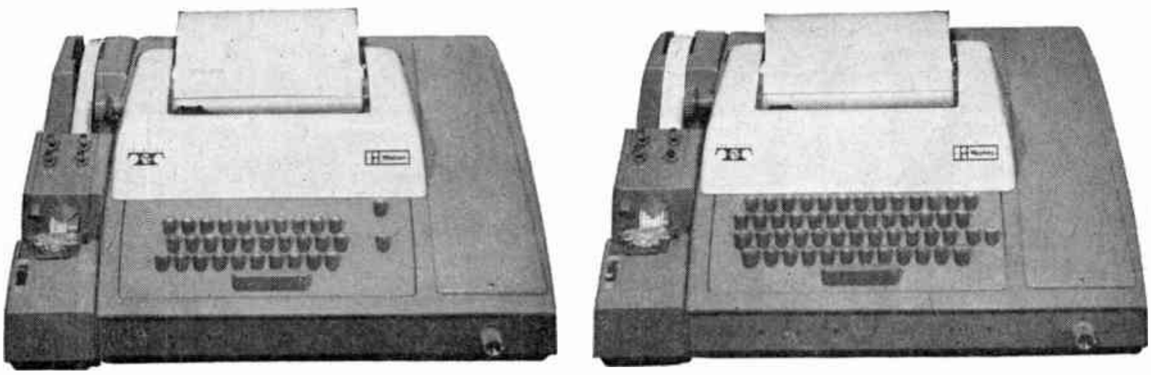
SMALL SIGNAL CHARACTERISTICS OF STC SILICON PLANAR TRANSISTORS is the title of a 38-page Application Report from STC. The report presents the data needed for the design of small signal amplifiers. The introduction presents the arguments for and against the basing of transistor circuit design on h or y parameters and finds in favour of the latter. Copies of this booklet, reference MK/178, are available from the STC Semiconductor Division (Transistors), Footscray, Sidcup, Kent.

SELECTION GUIDE FOR PRECISION RESISTORS. The 150 variations of Muirhead's range of precision wire wound resistors are given in a new publication entitled 'Precision Resistors'. A section is devoted to the selection of resistors by Value, Time-Constant, Accuracy, Rating, Temperature Coefficient, Temperature Range, Wire, Winding, Mounting and Lead Arrangements. Accuracies range from 1 per cent to 0·02 per cent, power ratings from 0·5W to 2·0W, and temperature coefficients can be as low as 10 parts in 10^6/°C. Copies of the brochure are available from Muirhead & Co. Ltd, Beckenham, Kent.

THE MAZDA VALVE DATA BOOKLET for 1964/5 has recently been published. The presentation continues in the handy pocket size which will remain open on the bench at any particular page. The booklet contains 160 pages giving abridged data on 259 current and obsolescent valves and picture tubes. A complete list of obsolete Mazda types and their applications is included for reference. The Equivalents List contains over 1200 fully cross-indexed types. Copies may be obtained by post, free of charge, by writing to: Thorn-AEI Radio Valves & Tubes Ltd, Publicity Department, 155 Charing Cross Road, London, W.C.2.

WAYCOM LTD have recently produced a pamphlet giving details of their range of heat sinks and transistor coolers available in the United Kingdom. The range includes transistor coolers for TO-5 and TO-18 types, and for the TO-5 the company are able to offer two alternatives: (a) The standard TO-5 cooler; (b) the types WA.201 and WA.202 for printed circuit boards. Further information, copies of the pamphlet and price list are available from the Publicity Manager, Waycom Ltd, Capacity House, Rothsay Street, Tower Bridge Road, London, S.E.1.

EE 75 150 *for further details*



These two Teletype page printers offer several unique advantages. They're available immediately from Westrex

The design refinements on Teletype Model 32 and 33 are just part of the overall picture of Westrex leadership. Model 33 works on an 11 unit 8 level code, Model 32 on a 7.5. unit 5 level code.
Unique Westrex features offered on both models are 100% accurate error correction or addition of information into tapes, combined or independent carriage return and line feed, and actuation of auxiliary functions by switch operation of character keys.

In addition the Model 32 offers a low paper alarm, coded keyboard, and has GPO approval for usage on external lines.
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WESTREX COMPANY LIMITED/152 COLES GREEN ROAD/LONDON NW2/GLADSTONE 5401

EE 75 151 *for further details*


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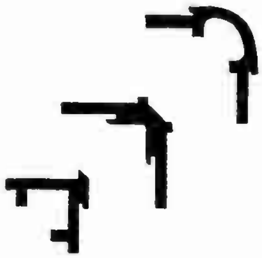

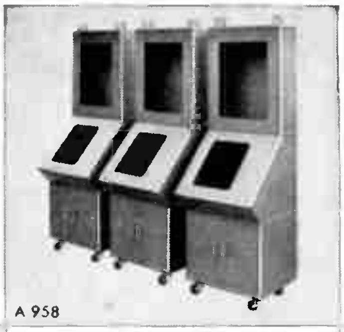
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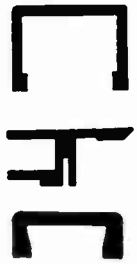

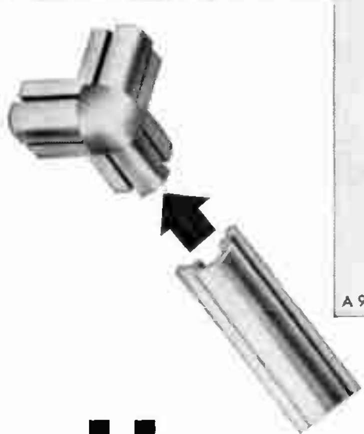

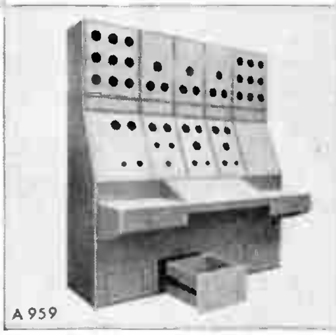
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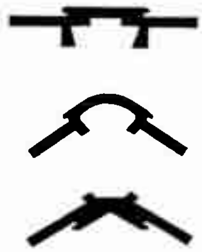

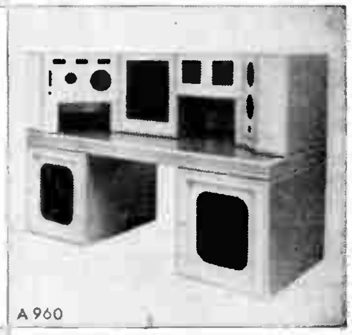
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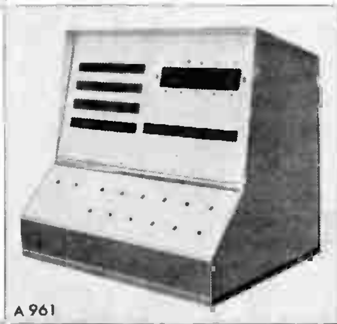
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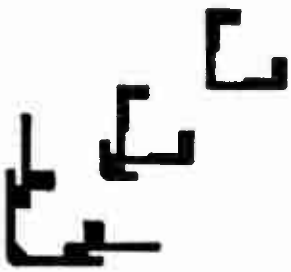


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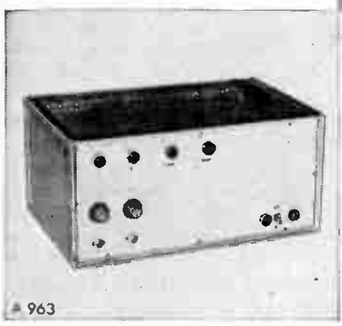
imlok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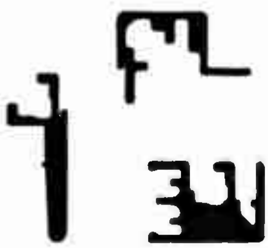

Imlok is a simple system of corner connectors and interlocking extrusions designed for making cases and cabinets of virtually any shape or size, in any quantity—cheaply and quickly. You can buy the individual components and build your own structures (choose from 48 different connectors, 60 extrusions and a whole host of accessories) or you can have a fully finished cabinet built by Imhofs. Send today for full details.

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IMLOK

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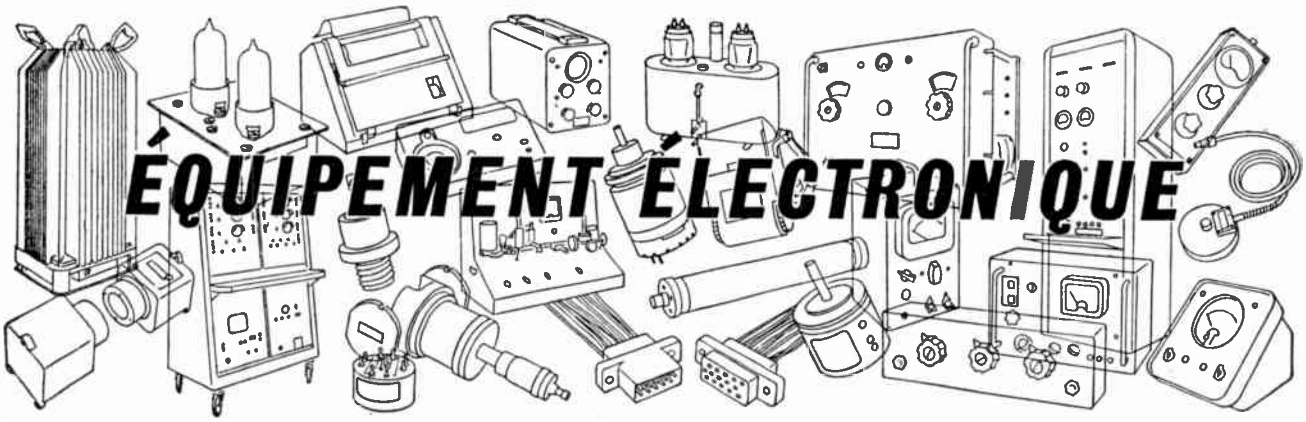
W2



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 786 à 791

DÉTECTEUR PHOTOÉLECTRIQUE

Simmonds Relays Ltd, Edinburgh Place, Temple Fields, Harlow, Essex

(Illustration à la page 786)

Le plus récent des dispositifs photoélectriques réalisé par la société Simmonds Relays est le modèle Visolux RL.1. Il s'agit d'un déclencheur photoélectrique réfléchissant la lumière et prévu pour l'analyse sans contact physique des marques sur les bords de papier, de matières plastiques et d'autres matières à rouleaux mobiles. Le détecteur RL.1 qui est équipé d'un amplificateur d'impulsions d'une grande sensibilité peut, par son action d'analyse, effectuer d'autres opérations telles que la guillotine subséquente.

L'équipement RL.1 se compose d'une tête d'analyse compacte, contenant une lampe de projection et un récepteur, l'ensemble de l'équipement ne mesurant que 65 mm de long × 18 mm de large et 30 mm de profondeur. La tête d'analyse est reliée à l'amplificateur de secteur par un câble de 2 m.

Le commutateur de l'amplificateur d'impulsion est entièrement transistorisé et comprend un bloc d'alimentation secteur incorporé pour tension alternative de 240 V, 50 Hz. En plus du commutateur à arrêt/marche et d'une lampe témoin secteur, il est muni de trois commandes pour le réglage du contraste de sensibilité et de l'action différée. La durée de chute du relais de sortie est réglable entre 40 et 150 msec. L'amplificateur est muni de trois pattes de fixation et mesure 190 mm × 140 mm. Le commutateur du relais est muni de contacts de permutation d'une puissance nominale non-inductive de 2 A, 240 V c.a.

EE 75 751 *pour plus amples renseignements*

RELAIS MINIATURE

B & R Relays Ltd, Temple Fields, Harlow, Essex

(Illustration à la page 786)

Le relais miniature interchangeable D05/P-D55/P comporte trois contacts de permutation et il est prévu pour montage sur base internationale à 11 broches. Il se distingue en particulier par trois contacts disposés périphériquement, chacun de ces contacts étant placé dans une poche individuelle, cette dernière assurant un meilleur isolement et empêchant les décharges entre contacts. La poursuite est réalisée par l'action de roulement de chaque contact à la fermeture. Il en résult une plus longue durée de contact due à la réduction de l'usure mécanique.

Les trois contacts de permutation sont en argent et la puissance nominale de chacun d'eux à 6 A est de 250 V c.a., 30 V c.c. La puissance de bobine nominale exigée est de 2 W ou 4 VA et la résistance de bobine maxima c.c. est de 10 200 Ω. Le D05 peut être utilisé pour des tensions atteignant 170 V c.c. et le D55 peut être utilisé pour des tensions atteignant 350 V c.a. Ne pesant que 106 g le relais peut être muni d'une bobine imprégnée au vide pour les conditions tropicales et d'humidité élevée. Il est fourni complet avec un cache poussière à fixation instantanée en Makrolon. Le diagramme des connexions des broches est gravé sur le haut de la housse.

Le D05/P-D55/P n'est fourni qu'en une seule forme et les connexions de broches ne peuvent pas être changées. Il mesure 5.87 cm de haut × 5,08 cm de large × 3,80 cm de profondeur le long de la base.

EE 75 752 *pour plus amples renseignements*

RÉSISTANCES BOBINÉES

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

(Illustration à la page 786)

La résistance bobinée type P8S est le dernier-né de la série des résistances bobinées de précision encapsulées fabriquées par la société Miniature Electronic Components Ltd. La nouvelle résistance mesure 0,79 cm de hauteur × 0,71 cm de diamètre et comprend les conducteurs de circuit à espacement de 0,53 cm. C'est la plus petite résistance de la série et sa puissance nominale est de 0,2 W à 85°C. La gamme de températures s'étend de −65° C à +150° C.

La gamme comprend 16 types avec câbles axiaux ou conducteurs de circuit imprimé, ou avec pattes radiales. Les valeurs de résistance peuvent atteindre 4,5 MΩ, les tolérances normales étant de 1%, 0,25% et 0,1%. Les types P12 P34 et P56 ont été homologués sous la référence DEF.5113.

EE 75 753 *pour plus amples renseignements*

COMMUTATEUR PHOTOÉLECTRIQUE

Lancashire Dynamo Electronic Products, Rugeley, Staffordshire

(Illustration à la page 786)

Un nouvel élément de commutation photoélectrique, le "Vigilite", série PLC.2, a été réalisé par la Lancashire Dynamo Electronic Products (M.I. Group).

Cet élément est livrable en deux modèles: un modèle avec deux têtes à distance et un modèle avec cellules incorporées et une seule tête à distance. Il est prévu pour l'utilisation intérieure ou extérieure et il commande toutes les formes de comptage, triage, production par fournée et détection à une vitesse maxima de cinq opérations par seconde.

Son circuit imprimé est entièrement transistorisé et les divers composants sont montés directement sur la plaquette de circuit imprimé. Cette dernière, dont l'épaisseur est de 0,23 cm, s'insère dans le coffret de commande dans lequel elle est fixée par des vis à la partie supérieure et à la partie inférieure.

Le coffret en fonte comporte un couvercle frontal amovible, muni d'une bande de scellement étanche autour des bords. Les dimensions hors-tout du coffret sont: 15,55 cm de large, 21,59 cm de haut et 7,77 cm de profondeur. Le montage mural peut s'effectuer à l'aide de quatre pattes en saillie.

Le relais intérieur comprend une paire

de contacts normalement ouverts et une paire de contacts normalement fermés, pouvant être utilisés pour la commutation. La puissance nominale est de 5 A, 230 V ou 1 A à 440 V c.a. non-inductif.

En disposant de manière appropriée une paire de maillons de fiches intérieures, le relais de l'appareil peut être désamorcé lorsque la lumière est faite ou lorsqu'elle est interrompue. Grâce à cette méthode, on peut effectuer la rupture à la lumière ou lorsqu'elle est interrompue, au choix de l'utilisateur.

La sensibilité préréglée et des commandes différentielles sont également prévues.

L'appareil peut être utilisé sur alimentation monophasée de 110 ou de 250 V, 50 Hz.

EE 75 754 *pour plus amples renseignements*

RELAIS MINIATURE

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustration à la page 787)

Ces relais miniature à bas prix sont livrables en version à permutation bipolaire ou quadripolaire avec contacts dorés ou d'argent. Ils sont fournis avec des housses transparentes anti-poussière et sont montés sur des bases à fiches.

La puissance nominale des contacts à 1 A est de 100 V c.a. ou 24 V c.c., avec des tensions de bobine pouvant atteindre 85 V c.c. La pression de contact est supérieure à 10 g.

EE 75 755 *pour plus amples renseignements*

SIMULATEUR "NI"

Brensal Electronics Ltd, Charles Street, Bristol 1

(Illustration à la page 787)

Cet appareil autonome peut être utilisé dans l'industrie ainsi que dans les collèges techniques pour la démonstration ou le contrôle de fonctions logiques simples. Toutes les interconnexions de signaux s'effectuent sur le panneau frontal de préaffichage à l'aide de cordons fournis à cet effet avec l'appareil. Grâce à ce dispositif, des systèmes peuvent être bobinés directement sans difficulté à partir de diagrammes logiques.

L'appareil comporte les modules suivants:

24 éléments NI dont 16 à 6 entrées et 8 à 2 entrées, tous les éléments comportant 3 douilles de sortie pouvant être utilisées pour entraîner 6 entrées si nécessaire.

2 minuteries à retard variable dans la gamme de 0 à 60 sec; deux types de sorties sont prévus en fonction de la douille d'entrée utilisée.

4 connexions de compteurs binaires reliées à l'entrée, au réglage et au réenclenchement. Des sorties Q s'obtiennent aisément sur le panneau de préaffichage.

12 tumblers, pouvant être utilisés comme dispositifs d'entrée. Ces tumblers choisissent des tensions de 0V ou de 24 V pour les signaux d'entrée aux élément NI.

8 éléments de sortie pour simuler les conditions de sortie, chaque étage n'ayant qu'une seule entrée.

Tous les modules comportent des voyants lumineux, ces derniers s'éclairant lorsque la sortie est à 0 V. Un courant limité peut être obenu pour actionner les éléments à l'extérieur du simulateur.

L'appareil fonctionne sur courant secteur alternatif de 200 à 250 V. Le simulateur est monté dans un coffret en bois avec couvercle amovible et ses dimensions hors-tout sont de 67,31 cm × 48,26 cm × 19,05 cm.

EE 75 756 *pour plus amples renseignements*

OSCILLOSCOPE À GRAND ÉCRAN

Distributeurs: Claude Lyons Ltd, Hoddesdon Hertfordshire

(Illustration à la page 787)

Le vide existant, en quelque sorte, entre l'oscilloscope de mesure à petit écran et l'oscilloscope d'affichage à tube de télévision vient d'être comblé par une gamme d'instruments de précision fabriqués par la société Constructions Radio-électriques et Électroniques du Centre (C.R.C.). Ces instruments, fournis en Angleterre par la société Claude Lyons Ltd, offrent tous les avantages d'un oscilloscope de mesure moderne à grande performance et emploient un tube cathodique spécial à écran de 180 mm fournissant une surface d'affichage effective deux fois plus grande que celle d'un modèle de 12,5 cm.

Cette gamme vient d'être complétée par l'oscilloscope X-à à grand écran, type OC 746. Cet instrument comporte des canaux X et Y identiques, dont chacun peut recevoir n'importe lequel d'une variété d'éléments à fiches. Deux fiches d'amplification identiques (trace unique, double trace, niveau bas ou différentiel) sont utilisées comme oscilloscope X-Y. Une seule fiche d'amplificateur et la base de temps BT 7461 dans le canal X produisent un oscilloscope classique des plus souples.

Par l'emploi des fiches HF 5661 (trace unique) et BT 7461, l'oscilloscope OC 746 assure une sensibilité verticale de 50mV/cm à 20V/cm du courant continu à 1 MHz. La gamme de la base de temps va de 0,5 µsec/cm à 2,5sec/cm avec possibilités d'expansion du balayage, de synchronisation totale et de déclenchement. Un calibreur de tension (se trouvant dans le châssis principal) fournit une onde carrée précise de 1 kHZ à des niveau de 0,5 mV à 100 V. La fiche BF 5662 fournit une entrée différentielle et des sensibilités de 1 mV/cm à 50 V/cm, tandis que l'élément à fiches BF 5672 à niveau réduit fournit des sensibilités de 100 µV/cm à 50 V/cm avec entrée différentielle. L'élément à double trace type CE 5673 assure une présentation intermittente ou à balayage à alternance avec une sensibilité de 50 mV/cm à 20 V/cm.

EE 75 757 *pour plus amples renseignements*

TACHYMÈTRE ÉLECTRONIQUE

S. Smith & Sons (England) Ltd, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustration à la page 787)

La société Smiths Industrial Division vient d'annoncer la mise au point d'un nouveau tachymètre électronique industriel de 15 cm. Deux modèles standard avec gamme d'échelle de 0 à 10 000 tours/minute et 0 à 15 000 tours/minute sont prévus, mains des indicateurs avec des gammes d'échelle pouvant atteindre 1 000 000 de tours/minute peuvent être fournis sur commande.

Ce tachymètre n'exige pas de couplage mécanique avec une partie mobile quelconque et il est particulièrement indiqué pour les applications ne comportant aucune prise mécanique.

Il comporte un indicateur à cadre mobile et à échelle circulaire de 270° monté dans un coffret qui comprend également le circuit d'entraînement transistorisé. Les lobes ferreux de la partie rotative d'une machine passant très près d'une tête à perception magnétique, montée séparément, produisent des impulsions que le circuit transistorisé convertit en une sortie proportionnelle à leur fréquence. Ainsi, la position de l'index dépend de la vitesse de l'arbre auquel les lobes sont fixés. Ce tachymètre exige une alimentation en tension continue de 24 V mais sa consommation électrique maxima n'est que de 100 mA. Des cellules primaires peuvent être utilisées.

Les lobes ferreux peuvent être coulés, usinés ou soudés à un arbre, à une engrenage droit ou à un volant. On a prévu des têtes de perception pouvant être utilisées avec des lobes montés autour de la périphérie. En variante, une autre tête peut être utilisée pour percevoir la rotation d'une pièce à fentes vissée au centre de l'extrémité d'un arbre à nu.

Lorsque une indication multiple est exigée, on peut utiliser n'importe quel nombre d'indicateurs à partir d'une seule tête de perception. On peut également obtenir des installations à têtes de perception multiples avec un seul indicateur, à l'aide de commutateurs de résistance à faible contact.

EE 75 758 *pour plus amples renseignements*

ENREGISTREURS DE TEMPS ET DE RÉSULTATS

Bowmar Instrument Ltd, Sutherland Road, London, E.17

(Illustration à la page 788)

La société Bowmar Instrument Ltd vient d'annoncer la production en Angleterre d'un indicateur de temps écoulé, type 1440, et d'un indicateur de résultats, type 1989.

Ces appareils sont particulièrement utiles pour les applications de bord car ils sont très lisibles, puisqu'ils prévoient la lecture numérique à quatre tambours, et de dimensions miniature. L'indication est donnée au moyen de chiffres blancs d'une hauteur de 0,31 cm sur tambour noir, le temps total étant de 9999 heures. Le diamètre hors-tout de l'appareil est de 17 mm, sa longueur est de 46 mm et il pèse 50 g. Divers types de montage peuvent être fournis, y compris les versions à scellement sur panneaux. La consommation électrique est de 1,1 W sur courant monophasé de 115 V, 400 Hz. Grâce à des inverseurs de courant mesurant environ 2,5 cm × 2,5 cm × 3,7 cm, on peut utiliser ces appareils sur tension continue de 28 V.

L'indicateur de résultats est logé dans un coffret analogue, et il peut effectuer des comptages de 9 999 coups. Le courant d'entrée est soit de 24 à 28 V c.c. à 2 W, soit de 115 V, 400 Hz à 2 W. La vitesse de comptage maxima est de 10 coups par seconde, avec durée d'impulsion de 50 msec.

Les deux appareils sont préréglés sans changement possible et ils ont été conçus pour répondre à la spécification MIL-M-7793C.

EE 75 759 *pour plus amples renseignements*

THERMISTANCE DE PROTECTION CONTRE LES SURCHARGES

Standard Telephones & Cables Ltd, Connaught House, 63 Aldwych, London, W.C.2

Le premier modèle d'une nouvelle gamme STC de thermistances à coefficient de température positive vient d'être mis sur le marché. Il s'agit du PTC 120, spécialement conçu pour la protection des moteurs, générateurs, transformateurs et autres appareils contre les températures excessives qui pourraient endommager l'isolement du bobinage. La température nominale de seuil du dispositif est de 120°C.

L'élément de thermistance du PTC 120 ne mesure que 4,75 mm de diamètre. L'enveloppe exterieure de ce petit dispositif est conçue de manière à pouvoir l'insérer dans des enroulements sans le moindre risque d'enlever l'isolement de l'émail des bobines.

Les conducteurs du PTC 120 sont sous forme de paires enroulées et leur longueur normale est de 60 cm. D'autres longueurs peuvent être fournies sur demande. Pour l'utilisation triphasée des versions différentes du PTC 120 peuvent être fournies.

EE 75 760 *pour plus amples renseignements*

FER À SOUDER MINIATURE

Antex Ltd, Grosvenor House, Croydon, Surrey

(Illustration à la page 788)

Le modèle C24ON vient d'être ajouté à la gamme de fers à souder Antex. Il s'agit d'un fer à souder miniature pour tension secteur pesant entre 56,6 gr. et 84,9 gr. Il est muni d'une panne "Ferraclad" dont la durée serait au moins cinq fois plus lonque que celle des pannes nickelées ordinaires et dont la capacité de fournir et de garder la chaleur serait nettement supérieure à celle de ces dernières. La panne peut être remplacée aisément et sans risque d'endommager le fer. Cet outil est de construction fort robuste et il est conçu pour l'usage continu.

EE 75 761 *pour plus amples renseignements*

TROUSSES DE CHAMBRES BLINDÉES

Belling & Lee Ltd, Great Cambridge Road, Enfield, Middlesex

(Illustration à la page 788)

En plus de sa gamme de compartiments à écrans, la société Belling & Lee fournit maintenant la série "100" d'enclos modulaires, offrant toutes les possibilités qu'on trouve normalement dans les installations construites sur commande.

La nouvelle série est basée sur une gamme d'éléments interchangeables à châssis métallique, pouvant être assemblés rapidement pour fournir une chambre blindée avec une atténuation pouvant atteindre 100 dB à des fréquences variant entre 1 et 500 MHz.

La nouvelle série comprend des garnitures des conduction spéciales pour la métallisation efficace HF des panneaux, ainsi qu'un nouveau modèle de porte fixée par pression et à auto-métallisation.

Des panneaux à fenêtres modulaires peuvent être fournis en métal à mailles, ou en nid d'abeilles comprenant une couche de conduits de blindage HF. Ces panneaux fonctionnent suivant le principe des "guides d'ondes au-dessous du point de coupure), donnant un degré élevé d'atténuation à la haute fréquence au-dessous de la fréquence de coupure, mais assurant une bonne visibilité dans la direction avant et un passage libre à l'air de ventilation.

Un panneau d'alimentation de puissance, fonctionnant en liaison avec des filtres appropriés, loge les blocs d'alimentation, le téléphone et d'autres dispositifs.

La nouvelle série se caractérise en particulier par la souplesse qu'elle permet de donner aux dimensions et à la disposition de la chambre, car les enclos peuvent être aisément modifiés ou élargis. Ils peuvent être également démontés et emmagasinés de manière compacte ou transportés vers un autre lieu. Les diverses pièces sont fournies sous forme de grande trousse, avec les sous-éléments, les panneaux de parquets au linoléum ainsi que tous les boulons, écrous et garnitures HF nécessaires. Les panneaux mesurent 2,23 m de hauteur × 1,11 m de largeur et les portes et autres panneaux de transmission sont tous de cette dimension modulaire. Des demi-modules, des bancs de travail, etc. peuvent être fournis pour des applications spéciales. Il y a, en outre, une gamme étendue de filtres électriques.

EE 75 762 *pour plus amples renseignements*

ENREGISTREUR À AUTO-ÉQUILIBRAGE

Distributeurs: Wessex Electronics Ltd, Midsomer Norton, Somerset

(Illustration à la page 788)

L'enregistreur potentiométrique auto-équilibreur Latronics Recordette-4 est un appareil à performance élevée, conçu pour une grande variété d'utilisations. C'est un instrument de format réduit, portatif, prévu pour une gamme étendue de vitesses d'avance du papier, l'enregistrement à l'encre et sans encre, et se caractérisant, en outre, par une précision d'$\frac{1}{2}$% de portée, une sensibilité de $\frac{1}{4}$% de portée, une performance asservie totale et un temps d'équilibrage sur la totalité de l'échelle de $\frac{1}{2}$ de seconde.

Les éléments d'entrée type 1 et 2 fournissent une portée électrique réglable entre 10 et 100mV sur la totalité de l'échelle. L'élément d'entrée type 3 fournit une portée réglable entre 10 mV et 100 V. L'élément d'entrée type 4 a été conçu pour les entrées de thermocouples, assurant l'enregistrement direct de la température.

La Recordette-4 a été étudiée pour assurer un fonctionnement sûr et de longue durée. L'accès direct au mécanisme d'entraînement de la bande, au style et à l'élément d'entrée, ainsi qu'à toutes les commandes rend l'utilisation de l'enregistreur extrêmement aisée.

Des vitesses d'enregistrement de 0,63 cm/h à 121,9 cm/minute sont prévues, la bande de papier de 10 cm peut être fixée sur le rouleau ou passée à travers une ouverture à l'extrémité inférieure de la porte pour les besoins de référence immédiate. Le modèle à montage sur panneau exige un découpage de panneau de 17,14 cm × 22,22 cm.

EE 75 763 *pour plus amples renseignements*

OSCILLOSCOPE DE LABORATOIRE

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(Illustration à la page 789)

La société Telequipment Ltd vient d'introduire sur le marché l'oscilloscope de laboratoire D43R, qui est un appareil pour montage sur bâti de 47,5 cm n'occupant qu'un espace de panneau vertical de 17,5 cm et dont la profondeur est de 40 cm.

Le D43R a été étudié pour une gamme étendue d'applications électroniques, médicales, industrielles et de laboratoire. Il est offert avec un choix de 5 amplificateurs à fiche Telequipment dont deux sont nouveaux.

Il comporte un tube à accélération de post-déviation à face plate de 10 cm, utilisé à 4 kV. Sa base de temps fournit 18 vitesses de balayage calibré préréglées de 0,5 sec/cm à 1μsec/cm. Le temps de montée et de 23nsec. l'impédance d'entrée est de 1 MΩ et l'appareil fonctionne sur alimentation de 100 à 240 V (50 à 100 Hz).

Les amplificateurs à fiches prévus sont les type A, B et C (universel, différentiel et à gain ultra-élevé), en plus des nouveaux éléments D et G. Le contrôleur d'enveloppe type D est un amplificateur accordé avec quatre gammes de commutation de 2,5 à 32 MHz. Lorsque l'appareil est accordé à la résonance, la sensibilité est d'environ

1 V/cm; la fréquence de modulation est fournie sous forme de sortie synchronisée et l'impédance d'entrée est de 50Ω.

Le type V est un amplificateur combiné à large bande et différentiel, les entrées différentielles ayant un rejet supérieur à 200:1, du c.c. à 1MHz (entrée d'onde sinusoïdale), tombant à un minimum de 30:1 à 12MHz. La tension d'entrée maxima est de 10 V de crête à crête (sur gamme de 10mV/cm).

EE 75 764 *pour plus amples renseignements*

VOLTMÈTRE-ENREGISTREUR DE RAPPORT NUMÉRIQUE

Digital Instruments Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustration à la page 789)

La société Digital Measurements Ltd vient d'ajouter le voltmètre numérique DM2022 à sa gamme étendue d'instruments numériques. Ce nouvel appareil a la précision et la fiabilité du voltmètre DM2020 mais comporte une échelle plus longue (39999) ainsi qu'un dispositif incorporé de mesure du rapport.

La précision du nouvel instrument est de 0,0025% sur la totalité de l'échelle à ±0,01% de la lecture et sa résolution élevée est de 1 partie dans 40000. Il comporte 5 gammes s'étendant de 0 à 2 kV, la sensibilité sur la gamme la plus réduite est de 10 μV. L'indication peut être présentée sur échelle extérieure donnant des lectures en lb-in^2, °C, et à partir d'entrées analogiques de tension. L'impedance d'entrée est supérieure à 25000 MΩ sur les deux gammes les plus basses et de 10 MΩ sur les gammes élevées. L'entrée peut être isolée de la masse afin de pouvoir rejeter les tensions de mode commun se trouvant dans la source de signaux.

Le dispositif de mesure du rapport constitue un avantage important, car il permet d'effectuer des mesures de comparaison entre les rapports de tension et les références extérieures, comme par exemple l'étalonnage potentiométrique, les mesures de calculatrice analogique, etc.

Le DM2022 comprend tous les dispositifs figurant dans les instruments DM antérieurs. Ces derniers comportent les modes de fonctionnement maximum et minimum de précision totale, ainsi que des sorties de décades dans n'importe lequel des six codes pour l'entraînement d'imprimeurs, de perforateurs, etc.

EE 75 765 *pour plus amples renseignements*

COMMUTATEURS DE PROXIMITÉ

W. H. Sanders (Electronics) Ltd, Gunnels Wood Road, Stevenage, Hertfordshire

(Illustration à la page 789)

Ces composants sont fournis depuis quelques temps à la United Kingdom Atomic Energy Authority et sont maintenant fournis à l'industrie en général. Le mouvement de commutation est basé sur un matériau ferreux extérieur qui lui permet de compléter un circuit magnétique et il peut être utilisé pour déceler la présence d'une pièce de machine quelconque sans qu'aucun contact ne soit effectué. Cette condition se pose lorsque l'emploi de commutateurs classiques est impraticable, par exemple en cas de présence de vapeur corrosive ou inflammable, de poussière ou de liquide, ou lorsqu'on manipule des matières radioactives.

Le commutateur de proximité est utilisé lorsque le matériau ferreux est à une distance d'environ 1,5 cm de la tête de captage. La distance effective à laquelle le commutateur fonctionne peut être réglée au moyen d'un potentiomètre de sensibilité. La tête de captage est entièrement imprégnée et encapsulée sous résine d'époxyde. Elle est insensible aux conditions qu'on rencontre normalement dans la plupart des procédés industriels.

Les deux versions du commutateur de proximité sont les suivantes: FSPS/042 ainsi que le modèle réduit PDM/3173, pouvant être utilisé avec un amplificateur à transistors bistable. Le modèle PDM/3173 peut être assemblé à l'intérieur de systèmes logiques à transistors, mais il peut également être utilisé séparément comme élément de commutation normale.

EE 75 766 *pour plus amples renseignements*

GÉNÉRATEUR DE FORMES D'ONDES

Servomex Controls Ltd, Crowborough, Sussex

(Illustration à la page 789)

En complément au générateur de formes d'ondes basse fréquence LF 51, la société Servomex Controls Ltd a réalisé un nouvel instrument capable de produire 90 formes d'ondes distinctes dans les gammes de fréquences acoustique et de servo-fréquence, tout en étant suffisamment petit et léger pour pouvoir être porté dans une seule main. Ce progrès notoire dans la conception de ce type d'instrument a été réalisé par l'emploi étendu de transistors et de circuits imprimés.

Le nouveau générateur, type LF 141, a une gamme de fréquence de 0,002 à 2 Hz et il produit toutes les 90 formes d'ondes sans matériel auxiliaire aucun. La tension de sortie va de 0 à 10 V de crête dans trois gammes, avec tensions de pointe maxima de 10 V, 1 V et 0,1 V.

L'instrument comporte un cadran à fréquence directe et la tension de sortie peut être soit équilibrée autour de la ligne de zéro soit polarisée. Ce processus produit des formes d'ondes supplémentaires ainsi que quelques formes d'ondes différées. L'instrument peut également être modulé; l'application d'une tension paralysante extérieure arrête le générateur principal, ce qui donne lieu à l'apparition d'un ou de plusieurs cycles.

L'instrument comporte en outre un oscillateur auxiliaire entièrement à part pour déclencher le générateur principal. On peut ainsi produire une grande variété de nouvelles formes d'ondes. En raison du fait que l'oscillateur auxiliaire produit une rampe ainsi que l'impulsion de déclenchement aigue, on peut synchroniser un oscilloscope comportant des commandes adéquates de niveau de déclenchement de courant continu afin d'observer le début de l'onde principale. On a donc ainsi l'équivalent d'une ligne à retard à basse fréquence incorporée dans l'oscilloscope cathodique, avec un retard maximum de quelques secondes.

L'intégrateur de base est d'une conception nouvelle qui élimine le retard que l'on subit habituellement avec ce type d'instrument, après avoir appliqué une impulsion de déclenchement. L'emploi d'un élément à phases variables produit une onde sinusoïdale améliorée en raison de l'absence de sommets plats, sur l'onde triangulaire du générateur de base. Grâce à cet élément on peut mettre l'onde en train à n'importe quel angle de phase voulu.

EE 75 767 *pour plus amples renseignements*

APPAREIL DE MÉTALLISATION PAR THERMOCOMPRESSION

G. V. Planer Ltd, Windmill Road, Sunbury-on-Thames, Middlesex

(Illustration à la page 790)

Cet appareil, qui vient d'être réalisé par la société G. V. Planer Ltd, constitue un outil d'une grande souplesse d'emploi pour la production d'agglomérants à thermocompression utilisés dans la recherche et la construction. Il est particulièrement indiqué pour la métallisation de conducteurs à des micro-circuits à film mince, ainsi que pour les matériaux et dispositifs semi-conducteurs, etc.

Cet appareil se distingue en particulier par ses nouveaux micro-manipulateurs de précision à mouvement X, Y et Z, ainsi que par des commandes spéciales à double tige, sa commande thermostatique indépendante du ciseau de métallisation et du sous-étage, et par la configuration flexible du micro-manipulateur et des positions des sous-étages sur l'établi.

L'équipement comprend deux micro-manipulateurs, dont l'un porte le ciseau de métallisation à commande thermostatique avec charge variable et l'autre le dispositif d'avance du fil métallique à air comprimé. Ce dernier peut recevoir des bobines standard, étant étudié pour des fils d'un diamètre variant entre 0,013 mm et 0,25 mm. Un couteau à fil métallique est fixé à l'extrémité du mécanisme d'avance. Le ciseau de métallisation et le sous-étage peuvent être maintenus à des températures atteignant 600° C, et permettent l'émission de gaz inertes sur la surface de travail, si nécessaire.

Un microscope binoculaire "stéréo-zoom", avec grossissement à variation continue de 7 à 30 ou de 14 à 60 et éclairage réglable, est incorporé à l'appareil.

Les composants de fonctionnement et le pupitre de commande sont montés sur un meuble en acier avec une surface de 75 cm × 121,9 cm.

EE 75 768 *pour plus amples renseignements*

DÉTECTEUR DE PAILLES À ULTRA-SONS

Ultrasonoscope Co. (London) Ltd, Sudbourne Road, Brixton Hill, London, S.W.2

(Illustration à la page 790)

Ce nouveau détecteur de pailles à ultra-sons, appelé Mark 5, est entièrement transistorisé et comporte un tube cathodique de 12,5 cm qui assure la précision de la mesure d'épaisseur et la vision la plus nette possible.

Les fréquences de commutation sont de 5, 2,5 et 1,1/4 MHz et le gain est commandé par un atténuateur étalonné. La vitesse la plus rapide de la base de temps est de 2 μsec. On peut ainsi étendre l'image de 6 mm d'acier sur toute la largeur de l'écran. La base de temps a un retard à variation continue et la portée maxima est de 6 m dans l'acier. La fréquence de répétition des impulsions varie de 50 à 1000 Hz, afin de pouvoir éliminer les échos fantômes. Les phénomènes de diffusion provenant de la structure d'un métal peuvent être également éliminés à l'aide d'une commande de rejet à degrés de commutation. On peut obtenir une image rectifiée ou non rectifiée, cette dernière s'étant démontrée essentielle pour un certain nombre de problèmes complexes.

Les commandes les plus importantes sont placées sur le panneau frontal, les autres étant groupées par ordre d'emploi sur un des côtés. Les batteries rechargeables ont une durée de 7 heures et sont instantanément interchangeables. Le détecteur Mark 5 peut, le cas échéant, être utilisé sur alimentation secteur normale, par l'emploi d'un bloc d'alimentation secteur spécial dans le compartiment des batteries. Les dimensions hors tout de l'appareil sont de 24,13 cm × 17,78 cm × 54.61 cm et il pèse 12,63 kg, batteries y comprises.

EE 75 769 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS

Distributeurs: B. & K. Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Illustration à la page 790)

Le générateur d'impulsions universel, modèle 6613, construit par la Texas Instruments, fournit des impulsions de sortie positives et négatives coïncidentes à une vitesse déterminée par un générateur à minuterie intérieur, un signal extérieur, ou à partir d'un commutateur manuel monocourse. La largeur de l'impulsion est variable jusqu'à 90° du cycle de régime. Le retard peut être réglé pour la sortie avant ou après la sortie de synchronisation. L'amplitude de la sortie peut être variée de 0 à 10 V avec protection contre les surcharges de sortie. L'appareil est entièrement à circuits imprimés et il mesure dans son ensemble 21.59 cm × 21,59 cm × 30,48 cm et pèse 4,5 kg.

La fréquence de répétition des impulsions peut être variée en six gammes à degrés de décades de 15 Hz à 15 MHz et l'amplitude peut être variée de 0 à 10 V dans 50 Ω. Les temps de montée et de chute peuvent être variés de 10 nsec à 10 msec en trois gammes à degrés. Le retard peut être varié de 30 nsec à 30 msec en six gammes à degrés de décades. La largeur peut être variée de 30 nsec à 30 msec en six gammes à degrés de décades. La sortie de synchronisation est positive à 3 V, les temps de montée et de chute sont de 15 nsec et la durée est de 50 à 80% du cycle, de régime, selon le réglage de la commande de fréquence. L'entré d'entraînement extérieur exige une tension positive de 2 V, avec durée minima de montée de 20 nsec et de 30 nsec.

Une impulsion unique peut être fournie par une commande à bouton-poussoir sur le panneau frontal, et un commutateur sur le panneau frontal permet le déclenchement périodique des impulsions de sortie, "arrêt" à 0 V et "marche" à 7 V positifs.

EE 75 770 *pour plus amples renseignements*

RELAIS DE COMPTAGE À REBOURS

Distributeurs: D. Robinson & Co. Ltd, 5-7 Church Road, Richmond, Surrey

(Illustration à la page 791)

Le relais de comptage à rebours Rodene est un dispositif permettant de compter n'importe quel nombre d'impulsions de tension continue ou alternative. Chaque impulsion reçue excite une solenoïde qui déplace une roue de rochet d'une position. Lorsque le nombre préréglé est atteint, un microrupteur est mis en action et dirige à nouveau toute nouvelle impulsion vers la bobine de réenclenchement séparée ou vers le circuit extérieur de l'utilisateur.

Tous les relais de comptage à rebours peuvent être réglés de manière à compter n'importe quel nombre jusqu'à 20. Deux ou plusieurs relais peuvent être reliés en cascade de manière à ce que l'impulsion de réenclenchement du relais A actionne la bobine à degrées du relais B afin de déclencher le comptage par décades.

Les relais de comptage à rebours sont fournis soit pour réenclenchement automatique soit pour réenclenchement manuel. Les types à réenclenchement comportent un index qui reste à la position préréglée durant le minutage. L'index des types à réenclenchement manuel se déplace à la même vitesse que la roue à rochet, donnant ainsi à tout moment une indication visuelle du nombre d'impulsions reçues.

Le réenclenchement automatique s'effectue par une seule impulsion à une bobine séparée qui déclenche un rochet et permet à un ressort à retour de réenclencher le relais en 50 msec.

Un ou deux bancs de commutateurs peuvent être montés au lieu ou en plus du microrupteur et ils peuvent être utilisés pour déclencher une série de circuits extérieurs à travers l'ensemble du cycle. N'importe quelle longueur d'impulsion de plus de 25 msec peut actionner le relais, mais comme dans tous les relais, la dissipation est élevée afin d'activer le fonctionnement et par conséquent la bobine ne doit pas être excitée de manière continue. La vitesse maxima est de 5 plots par seconde et la durée du relais est de l'ordre de 3 millions d'opérations. Tous les modèles sont fournis avec les connecteurs "AMP-Faston."

EE 75 771 *pour plus amples renseignements*

ÉTALON DE FRÉQUENCE

General Radio Co. (U.K.) Ltd, Marlow Road, Bourne End, Buckinghamshire

(Illustration à la page 791)

L'oscillateur de fréquence standard type 1115-B est le tout-dernier de la série d'étalons de fréquence a cristal de quartz de la société General Radio Co. Le nouvel instrument utilise un cristal getter à cinquième harmonique de 5 MHz dans un oscillateur à transistors à 3 étages. Le cristal, l'oscillateur et les circuits CAG sont tous enfermés dans un four à commande proportionnelle. La fréquence d'oscillation est divisée pour fournir des sorties à 1 MHz et 100 kHz ainsi qu'à 5 MHz.

La stabilité à court terme (déviation à 95 % des limites de confiance) est d'une partie dans 10^{11} (efficaces) pendant une durée moyenne de 10 sec, 100 parties dans 10^{11} (efficaces) pendant une durée de 300 μsec. Le vieillissement caractéristique est inférieur à une partie dans 10^{10} par jour après un an. La largeur de la ligne spectrale est inférieure à 0,25 Hz sur la bande X. Le facteur de bruit est de 145 dB et au-dessous pour une largeur de bande de 1 Hz sur la sortie de 5 MHz.

Le nouvel étalon utilise les composants actifs constitués de corps solides dans un coffret spécialement conçu et construit pour les applications mobiles et ardues. L'alimentation par batterie au nickel cadmié incorporée est mise en oeuvre automatiquement et actionne l'étalon pendant 35 heures sur courant secteur alternatif.

EE 75 772 *pour plus amples renseignements*

COFFRET CLIMATIQUE

Fisons Scientific Apparatus Ltd, Loughborough, Leicestershire

La division Weyco de la société Fisons Scientific Apparatus Ltd, vient d'ajouter un nouveau coffret climatique à sa gamme d'appareils de contrôle du milieu.

Le nouvel appareil, portant la référence CM/49, a les caractéristiques suivantes:

Gamme de température: 25° C à 150° C contrôlée jusqu'à ± 1° C;

Gamme d'humidité: 12% au maximum contrôlée jusqu'à ±1 %;

Capacité de la chambre de travail: 9,4 pieds cubes.

Un système breveté de débit d'air assure la distribution de l'air conditionné dans l'ensemble de la chambre de travail. L'humidité est introduite par l'injection d'eau atomisée. Un élément de réfrigération hermétiquement scellé est incorporé à l'appareil et permet non seulement le fonctionnement à basse température mais également d'obtenir des humidités inférieures à l'humidité ambiante. Toutes les commandes sont montées à l'avant du coffret afin de faciliter son utilisation. On a accès au composant électrique en soulevant le panneau de commande à charnières. Un enregistreur à deux plumes, enregistrant de manière continue la température et l'humidité, constitue un accessoire standard.

Le nouveau coffret répond à la plupart des besoins des ingénieurs et du personnel de contrôle de la qualité des industries chimiques et de peintures pharmaceutiques.

EE 75 773 *pour plus amples renseignements*

RÉCEPTEUR HF À STABILITÉ ÉLEVÉE

The Marconi Co. Ltd, Chelmsford, Essex

(Illustration à la page 791)

Un nouveau récepteur universel HF à haute stabilité, type H2301, couvrant la bande de fréquences de 500 Hz à 30,5 MHz, vient d'être réalisé par la société Marconi. Ce récepteur compact et économique permet la réception de signaux modulés en amplitude, d'ondes constantes et de bandes latérales uniques. En lui ajoutant un adaptateur spécial, type H5011, il peut être utilisé avec des circuits de téléimprimeurs à une seule voie.

La gamme de fréquence est couverte en 30° de 1 MHz avec un chevauchement de 100 kHz, et la précision de l'étalonnage est de 1 kHz. L'étalonnage du cadran est linéaire et présentée de manière à ce que la fréquence soit indiquée en combinant les lectures sur des échelles séparées de "MHz" et "kHz". On peut vérifier l'étalonnage à n'importe quel moment à l'aide d'un oscillateur à cristal de 100 kHz incorporé.

On obtient une grande sélectivité au moyen d'une technique de double conversion, employée avec le contrôle piezoélectrique du premier oscillateur local afin de donner une stabilité exceptionnellement élevée pour ce type de récepteur.

La deuxième MF fonctionne à 500 kHz et un montage à cathode asservie fournit une sortie à cette fréquence, pouvant être reliée à l'équipement auxiliaire. Cinq positions de selectivité sont prévues, dont deux emploient des filtres passe-bande. Un filtre BF permet la réception sélective d'ondes entretenues.

Toutes les commandes de panneau sont placées de manière pratique afin de faciliter le fonctionnement et le récepteur comprend un haut-parleur incorporé. Des sorties audio séparées peuvent être fixées avec des commandes de gain indépendantes.

Des composants de haute qualité sont utilisés dand le récepteur qui est de construction fort robuste, étant composé de six sous-châssis et de deux doubles coffrets blindés, boulonnés l'un à l'autre. Il peut être monté sur châssis ou sur banc d'essai et utilisé à partir de n'importe quelle tension alternative secteur.

L'adaptateur est entièrement transistorisé et contrôle des décalages de 200 à 850 Hz à des vitesses télégraphiques pouvant atteindre 100 bauds.

Le récepteur Marconi H2301 était à l'origine un récepteur Eddystone, mais il a été modifié pour répondre à une spécification Marconi de fonctionnement à bande latérale unique et à modulation par déplacement de fréquence.

EE 75 774 *pour plus amples renseignements*

INSTRUMENT DE MESURE MULTIGAMMES

Oxley Developments Co. Ltd, Priory Park, Ulverston, Lancashire

(Illustration à la page 791)

La société Oxley Developement Co. Ltd, vient d'introduire sur le marché britannique le "MONOC," qui constitue un instrument de mesure multigammes de poche et d'une conception inédite.

Grâce à une précision sur tension continue de 1,5 % et une précision sur tension alternative de 2,5 %, l'appareil couvre des gammes pouvant atteindre 1000 V c.c. et c.a. Des gammes de courant atteignant 1 A c.c. et 10 c.a. sont également prévues. L'instrument comporte un ohmmètre qui ne nécessite aucun réglage et qui peut effectuer des mesures jusqu'à 2 MΩ. Il y a, en outre, de nombreux accessoires permettant d'étendre les gammes.

En raison de l'originalité de la conception on a pu ajouter un dispositif à échelle à miroir comportant une échelle de 9 cm dans un coffret ne mesurant que 15,24 cm × 10,16 cm × 5,08 cm (approximativement).

Le mouvement, qui est bien protégé par les diodes et les fusibles, est pratiquement indestructible. En raison de son équilibrage minutieux, l'instrument peut être utilisé dans n'importe quelle position.

Un bracelet élastique permet de tirer parti pleinement de la légèreté de poids de l'instrument. Il peut, en effet, être porté sur le poignet, laissant ainsi l'usage libre des deux mains.

Le sélecteur de gammes unique permet le fonctionnement sans difficulté aucune avec une seule main, rendant ainsi l'instrument des plus praitques.

EE 75 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Un système d'enregistrement multivoies de la contrainte des arbres

par C. I. Sach

Résumé de l'article aux pages 736 à 739

Cet article décrit un système d'enregistrement de contrainte réalisé spécifiquement pour étudier la réponse des arbres, mais qui trouve une application générale dans les laboratoires et pour les travaux de compagne exigeant un certain nombre d'enregistrements extensométriques de la résistance. Cet appareil fonctionne sur batterie et il est étalonné de manière à demeurer insensible aux variations de la tension d'alimentation. Son absence de linéarité est inférieure à 1% de la sortie maxima et il convient à la mesure de la contrainte dynamique car ses composants de fréquence vont jusqu'à 10Hz.

Transformateurs de rapport toroïdaux; leur circuit équivalent et la mesure de leurs paramètres

par A. J. Binnie et T. R. Foord

Résumé de l'article aux pages 740 à 745

Cet article réunit une série d'analyses publiées sur les circuits équivalents qui représentent les types les plus utiles de transformateurs toroïdaux utilisés dans les réseaux en pont à basse fréquence de précision. Des méthodes sont indiquées qui permettent de mesurer divers paramètres utilisés dans ces circuits équivalents. Les effets de capacités bobinées entre elles ne sont pas considérés de sorte que les circuits et équations dont il est question dans cet article ne s'appliquent pas nécessairement aux fréquences acoustiques élevées.

Un réseau sélectif transistorisé à résistance par capacité par T. H. Appleby

Résumé de l'article aux pages 746 à 749

Il est maintenant généralement connu que la micro-miniaturisation électronique améliorera finalement la fiabilité des appareils électroniques en réduisant leurs dimensions. Les techniques basées sur les circuits constitués de corps solides ainsi aue sur les films minces ne permettent pas l'emploi d'inducteurs pour des fréquences inférieures à 30MHz de sorte qu'il faudra avoir recours à des circuits transistorisés de résistance par capacité. Le circuit décrit dans cet article résulte de recherches entreprises au sujet de ces circuits.

Un nouvel intégrateur et son utilisation pour le maintien de la sensibilité dans les systèmes de mesure du zéro par H. C. Bertoya

Résumé de l'article aux pages 750 à 755

Il s'agit ici d'un nouvel intégrateur constituant l'équivalent électronique d'un moteur asservi. Il peut être utilisé en particulier pour maintenir la sensibilité d'un système de mesure du zéro actionné par servomécanisme qui compare le comportement d'un élément inconnu (ou "d'essai") à un élément normal (ou de "référence").

La xérographie appliquée à un imprimeur de calculatrice à action rapide par R. H. Dagnall et P. F. T. C. Stillwell

Résumé de l'article aux pages 756 à 760

Les auteurs donnent un aperçu des conditions exigées pour l'impression des données de calculatrices commerciales.

Ils décrivent un appareil à imprimer fonctionnant "en ligne" ou "hors ligne" à partir d'une calculatrice numérique utilisant la xérographie pour produire un enregistrement à impression continue. Des éléments choisis de donnée fixe (comme par exemple des schémas de formes) sont projetés par un réservoir à film et imprimés en même temps que les données variables qui sont affichées sous forme de caractères sur une paire de tubes cathodiques. La sélection de la forme ainsi que sa disposition sont commandées par programme; les circuits de contrôle détectent certaines erreurs ou certains défauts de fonctionnement de la machine et font en sorte que toute forme erronée soit clairement marquée.

Le papier se déplace à une vitesse constante de 20cm par seconde, donnant une vitesse d'impression de 2.800 lignes par minute.

Dispositif électronique pour compter et enregistrer les insectes dans la recherche agricole par E. J. Brach et W. J. Mason

Résumé de l'article aux pages 761 à 764

Pour pouvoir lutter contre les insectes dans le blé et dans la farine, les entomologistes doivent étudier leurs habitudes. Par exemple, ils doivent connaître le nombre d'insectes se trouvant dans une unité de blé ou de farine, ainsi que le période de leur activité maxima. Les auteurs décrivent un instrument permettant de déterminer ces facteurs. Il se compose d'un piège à insectes, d'un transducteur optique, d'un circuit de comptage, d'une minuterie et d'un appareil imprimeur qui enregistre le nombre d'insectes sortant d'un récipient à farine ou à blé pour chercher de l'eau.

Stabilisation de la tension dans les appareils électroniques hybrides par P. Visontai

Résumé de l'article aux pages 764 à 765

La fourniture de la tension d'alimentation pour les transistors d'appareils électroniques utilisant à la fois des lampes et des transistors pose des problèmes particuliers.

Si l'une des lampes porte suffisamment de courant continu pour répondre aux besoins des parties transistorisées de l'appareil, on peut alors utiliser une seule alimentation.

L'auteur décrit un dispositif qui rend possible, pratiquement sans frais supplémentaires, la stabilisation de la tension d'alimentation réduite d'un groupe de transistors contre les variations de consommation de courant continu d'un autre groupe de transistors alimentés par la même source.

La mesure du spectre de bruit par diodes tunnel dans la gamme de fréquence de 5kHz à 10MHz par R. A. Giblin

Résumé de l'article aux pages 766 à 769

Cet article décrit la mesure du bruit dans les diodes tunnel au germanium à courant de pointe de 1mA de 5kHz à 10MHz, dans la gamme de tensions de polarisation de 0 à 450mV, en ce qui concerne en particulier la zone de conductance négative de la caractéristique. On a utilisé à cet égard une méthode de comparaison s'appliquant à l'ensemble de la gamme de fréquence. Des difficultés particulières ont été éprouvées dans la mesure du bruit. Le problème de la stabilisation de la diode a été étudié de manière approfondie. L'article indique un circuit ayant permis de mesurer simultanément la caractéristique statique et la conductance d'inclinaison négative. Les résultats obtenus montrent que le spectre obéit à une loi de basse fréquence jusqu'à 10MHz pour des tensions de polarisation dépassant 300mV et se réduit graduellement à un spectre plat au-dessous de 50mV.

Circuit à amplificateur à contre-réaction à un seul étage pour le calcul analogique à répétition par R. E. King

Résumé de l'article aux pages 770 à 771

L'auteur décrit un amplificateur simple de courant continu à circuit à contre-réaction dont le gain du circuit ouvert est d'environ 2000. Il présente également une analyse de sa réponse de fréquence. L'amplificateur convient particulièrement pour le calcul analogique à répétition dans la gamme de fréquences acoustiques.

Générateur transistorisé de dents de scie à variation linéaire et à plots négligeables par J. K. Moss

Résumé de l'article aux pages 772 à 775

L'auteur décrit un générateur linéaire de dents de scie à pente bien définie et à plots initiaux négligeables. Les dents de scie sont synchronisées avec un déclencheur extérieur. Grâce à la conception particulière du circuit de commutation, le retard entre l'application de l'impulsion de déclenchement et le commencement de la dent de scie est maintenu à une valeur réduite. On peut aussi obtenir le balayage positif et négatif.

Un convertisseur fréquence linéaire/tension par W. P. O'Grady

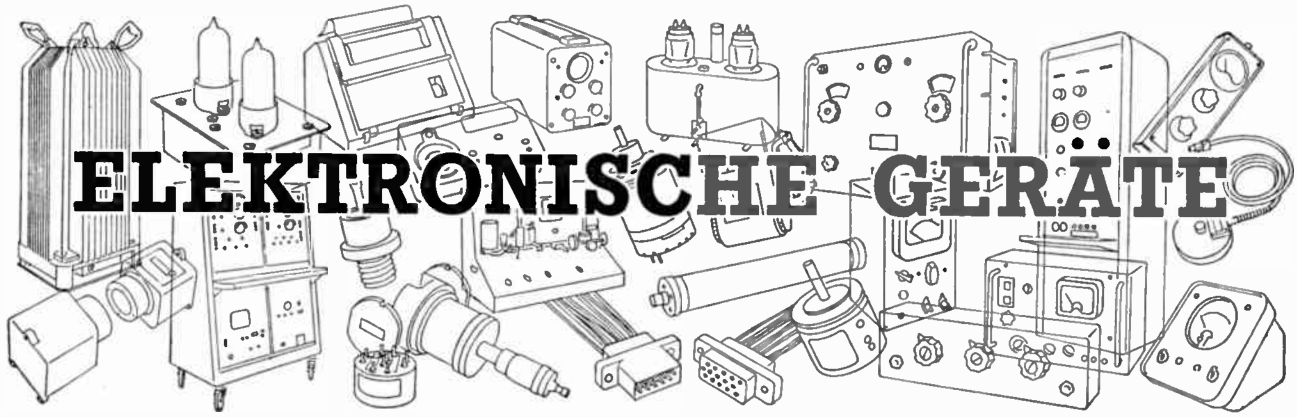
Résumé de l'article aux pages 776 à 778

Un circuit semi-conducteur convertissant linéairement la fréquence en tension est étudié dans cet article. L'auteur examine certaines méthodes permettant d'améliorer la caractéristique de phase entre la fréquence d'entrée et la tension de sortie, ainsi que des moyens de réduire la tension d'ondulation superposée à la tension de sortie.

Un compteur à décades à cathode froide de 50Hz par R. D. Ryan

Résumé de l'article à la page 779

Un multivibrateur simple npn fournit un circuit d'entraînement économique pour un tube de comptage à décades au gaz et à cathode froide de 50Hz. Les diodes de rétablissement à tension continue sur les barres de guidage assurent des niveaux constants d'impulsions en escalier.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 786 bis 791

Fotoelektrisches Suchgerät

Simmonds Relays Ltd, Edinburgh Place, Temple Fields, Harlow, Essex

(*Abbildung Seite 786*)

Das neuste fotoelektrische Visolux-Gerät der Simmonds Relays ist das fotoelektrische Tor für reflektiertes Licht Typ RL.1 zum Abtasten von Papier, Kunststoff und anderen sich bewegenden aufgespulten Materialien auf gedruckte Kantenmarken ohne direkte Berührung. Ein mit einem hochempfindlichen Verstärker ausgerüstetes RL.1 kann durch seine Abtastung andere Vorgänge, wie z.B. nachfolgendes Abschneiden, auslösen.

Die RL.1-Ausrüstung besteht aus einem kompakten Suchkopf mit Projektorlampe und Empfänger und ist insgesamt 65 mm lang, 18 mm breit und bis zu 30 mm tief. Anschluss des Suchkopfes an den Hauptverstärker erfolgt über ein 2 m langes Kabel.

Der Impulsverstärker-Schalter ist volltransistorisiert mit eingebautem Netzanschluss für 240 V, 50 Hz, Netzschalter und Netzsignallampe und hat drei Bedienelemente zur Einregelung von Empfindlichkeit, Kontrast und Zeitverzögerung. Die Abfallverzögerung des Ausgangsrelais ist zwischen 40 und 150 ms einstellbar. Die Abmessungen des mit drei Befestigungslappen ausgestatteten Verstärkers sind 190 × 140 mm. Der Relaisschalter ist mit Umschaltkontakten für 2 A induktionsfrei bei 240 V~ ausgerüstet.

EE 75 751 *für weitere Einzelheiten*

Miniaturrelais

B & R Relays Ltd, Temple Fields, Harlow, Essex

(*Abbildung Seite 786*)

Das Miniatur-Einsteckrelais D05/P-D55/P hat drei Umschaltkontakte und ist für die internationale 11-Stift-Fassung ausgelegt. Ein ungewöhnliches Konstruktionsmerkmal ist die peripherische Anordnung der Kontakte, die in Einzelkammern untergebracht sind, wodurch die Isolierung verbessert und Überschlagen zwischen Kontakten verhindert wird. Durch einen Rollvorgang beim Schliessen jedes Kontaktes wird gute Kontaktgabe erzielt. Der geringe Verschleiss bewirkt eine längere Kontaktlebensdauer.

Die drei Umschaltkontakte bestehen aus Silber und sind jeweils für 6 A, 250 V~ oder 30 V— bemessen. Die erforderliche Spulennennleistung ist 2 W oder 4 VA und der maximale Gleichstromwiderstand der Spule 10 200 Ω. Typ D05 ist für Betrieb mit Spannungen bis zu 170 V— und Typ D55 mit Spannungen bis zu 350 V~ geeignet. Das nur 106 g wiegende Relais kann für die Tropen oder hochfeuchte Verhältnisse mit einer vakuumimprägnierten Spule ausgerüstet werden. Es wird mit angeschnappter Haube aus Makrolon geliefert, auf deren Stirnseite das Anschlussdiagramm eingraviert ist.

Das Modell D05/P-D55/P wird nur in einer Ausführung geliefert, und Verbindungen zu den Anschlussstiften können nicht geändert werden. Die Abmessungen sind 59 mm hoch auf einem Sockel von 51 × 40 mm.

EE 75 752 *für weitere Einzelheiten*

Drahtwiderstände

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

(*Abbildung Seite 786*)

Miniature Electronic Components Ltd hat ihr Fertigungsprogramm für eingekapselte Präzisionsdrahtwiderstände durch den neuen Typ P8S ergänzt.

Dieser Typ ist 8 mm hoch, hat einen Durchmesser von 7,1 mm mit Druckschaltungsanschlüssen für das Rastermass von 5,08 mm. Es ist der kleinste Widerstand des Sortimentes und für 0,2 W bei 85°C bemessen. Der Betriebstemperaturbereich ist −65°... +150°C.

Das Programm umfasst 16 Typen mit axialen oder Druckschaltungsanschlüssen sowie radialen Lötfahnen. Widerstandswerte bis zu 4,5 MΩ sind mit Standardtoleranzen von 1 Prozent, 0,5 Prozent und 0,1 Prozent lieferbar.

Volle Baumustergenehmigung nach der britischen Vorschrift DEF.5113 wurde für die Typen P12, P34 und P56 erteilt.

EE 75 753 *für weitere Einzelheiten*

Fotoelektrischer Schalter

Lancashire Dynamo Electronic Products, Rugeley, Staffordshire

(*Abbildung Seite 786*)

"Vigilite" ist ein neues fotoelektrisches Schaltgerät Serie LPC.2, dessen Fertigung bei der Lancashire Dynamo Electronic Products (M.I.-Gruppe) angelaufen ist.

Das Gerät ist in zwei Ausführungen —mit Ferngeber und Fernzelle oder mit eingebauter Zelle und Ferngeber— lieferbar und für die Steuerung von Zähl-, Sortier-, Partiezähl- und Suchvorgängen mit einer Höchstrate von 5 Vorgängen je Sekunde in Gebäuden oder im Freien geeignet.

Die Schaltung ist volltransistorisiert und in Drucktechnik mit direkt auf die Druckkarte montierten Bauelementen ausgeführt. Die 2,4 mm dicke Karte passt in das Steuergehäuse, in dem sie mit Schrauben befestigt ist.

Das Gusseisengehäuse mit vier vorstehenden Lappen für Wandmontage hat einen abnehmbaren Deckel, dessen Kanten mit einem Dichtungsstreifen wetterfest gemacht sind. Die Aussenabmessungen sind 156 mm breit, 216 mm hoch und 78 mm tief.

Das interne Relais ist mit einem Arbeits- und einem Ruhekontakt ausgerüstet, die für Schaltzwecke dienen können und für 5 A, 230 V oder 1 A, 440 V~ (induktionsfrei) bemessen sind.

Durch entsprechendes Positionieren eines internen Steckverbindungspaares kann man das Relais so schalten, dass es entweder beim Auftreffen des Lichtstrahles oder dessen Unterbrechung abfällt. Auf diese Weise lässt sich je nach Wahl des Kunden Betrieb mit "Aussetzen bei Licht" oder "Aussetzen im Dunkeln" einstellen.

Empfindlichkeits- und Differenzregler sind vorhanden.

Das Gerät kann an Einphasennetze für 110 V oder 250 V 50 Hz angeschlossen werden.

EE 75 754 *für weitere Einzelheiten*

Miniaturrelais

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Abbildung Seite 787)

Diese preiswerten Miniaturrelais gibt es für zwei- oder vierpoliges Umschalten mit vergoldeten oder Silber-Kontakten. Sie werden mit staubdichten durchsichtigen Hauben und auf Einstecksockel montiert geliefert.

Die Kontakte sind für 1 A bei 100 V~ oder 24 V— bemessen, die Spulen für Spannungen bis zu 85 V— ausgelegt. Der Kontaktdruck ist besser als 10 g.

EE 75 755 *für weitere Einzelheiten*

NOR-Simulator

Brensal Electronics Ltd, Charles Street, Bristol 1

(Abbildung Seite 787)

Dieses in sich geschlossene Gerät ist für Einsatz in der Industrie wie auch in technischen Lehranstalten bestimmt, wo Vorführen oder Prüfen einfacher logischer Funktionen verlangt wird. Alle Signalverbindungen werden mit Hilfe von mitgelieferten Verbindungsschnüren auf dem Schaltfeld der Frontplatte hergestellt. Auf diese Weise können Systeme direkt von Logik-Diagrammen beschaltet werden.

Folgende Moduln sind vorhanden:

24 NOR-Elemente, davon 16 mit 6 Eingängen und 8 mit 2 Eingängen; alle Elemente haben drei Ausgangsbuchsen, die man auf Wunsch zum Treiben von 6 Eingängen verwenden kann.

2 Zeitschalter mit über den Bereich 0...60 s regelbarer Verzögerung; je nach verwendeten Eingangsbuchsen stehen zwei verschiedene Ausgangsarten zur Verfügung.

4 Binärzähler mit vom Schaltfeld zugänglichen Verbindungen für Eingang, Einstellen, Rückstellen, sowie Q und Q-Ausgänge.

12 Kippschalter für Einsatz als Eingangsorgane, die entweder O V oder 24 V als Eingangssignale für die NOR-Elemente wählen.

8 Ausgangseinheiten, die Ausgangsbedingungen simulieren, je Stufe jedoch nur einen Eingang haben.

Alle Moduln haben Signallampen, die aufleuchten, wenn der Ausgang 0 V ist.

Zum Steuern von externen Geräten steht eine begrenzte Leistung zur Verfügung.

Die Ausrüstung ist für Anschluss an 200...250 V~ ausgelegt und in ein Holzgehäuse mit abnehmbarem Deckel eingebaut. Aussenabmessungen sind 673 × 432 × 190 mm.

EE 75 756 *für weitere Einzelheiten*

Grossschirm-Oszillograf

Vertrieb: Claude Lyons Ltd, Hoddesdon, Hertfordshire

(Abbildung Seite 787)

Die zwischen den Messoszillografen mit kleinem Schirm und Oszillografen mit Fernsehröhrenanzeige bestehende Lücke wird durch eine Auswahl vortrefflicher, von Constructions Radio-électriques & Electroniques du Centre (C.R.C.) gefertigter Geräte geschlossen, deren Vertrieb Claude Lyons Ltd übernommen hat. Sie haben alle Einrichtungen, die man von einem Hochleistungs-Messoszillografen erwarten kann und sind mit einer speziell entwickelten 180-mm-Elektronenstrahlröhre bestückt, deren nutzbare Anzeigefläche doppelt so gross ist wie die eines 125-mm-Modelles.

Ein neues Gerät in diesem Programm ist der Grossschirm-X-Y-Oszillograf Modell OC746 mit identischen X- und Y-Kanälen, von denen jeder eine ganze Reihe von Einschüben aufnehmen kann. Bei Verwendung als X-Y-Oszillograf werden zwei identische Verstärker-Einschübe (Einspur-, Zweispur-, Differenz- oder Kleinsignalverstärker) benutzt. Mit einem Verstärkereinschub und dem Zeitablenkgerät BT7461 im X-Kanal ensteht ein vielseitiger herkömmlicher Oszillograf.

Mit den Einschüben HF5661 (einspurig) und BT7461 hat das Modell OC746 vertikale Ablenkkoeffizienten von 50 mV/cm...20 V/cm bei 0...1 MHz. Der Zeitablenkbereich ist 0,5 μs/cm... 2,5 s/cm mit Dehnung und vollen Synchronisierungs- und Triggereinrichtungen. Ein Spannungseicher im Hauptchassis gibt eine genaue 1-kHz- Rechteckwelle mit Pegeln von 0,5 mV bis zu 100 V ab. Der Einschub BF5662 hat einen Differenzeingang und Ablenkkoeffizienten von 1mV/cm...50 V/cm und der Kleinsignaleinschub BF5672 Ablenkkoeffizienten von 100 μV/cm... 50 V/cm mit Differenzeingang, während der Doppelspureinschub CE5673 zerhackte oder abwechselnde Ablenkungsdarstellung mit Koeffizienten von 50 mV/cm...20 V/cm gibt.

EE 75 757 *für weitere Einzelheiten*

Elektronisches Tachometer

S. Smith & Sons (England) Ltd, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Abbildung Seite 787)

Ein neues industrielles elektronisches 152-mm-Tachometer wurde von der Smiths Industrial Division angekündigt. Zwei Standardmodelle mit Messumfang 0...10 000 UPM und 0...15 000 UPM sind lieferbar, und auf Bestellung sind Ausführungen bis zu 1 000 000 UPM erhältlich.

Das Tachometer erfordert keinerlei Kupplung mit sich bewegenden Teilen und ist für Verwendungszwecke, in denen mechanisches Abgreifen unmöglich ist, ideal geeignet.

Der Drehspulanzeiger mit seiner 270°-Kreisskala ist im selben Gehäuse wie die transistorisierte Treiberschaltung untergebracht. Eisenhaltige Ansätze eines rotierenden Maschinenteiles, die in unmittelbarer Nähe eines getrennt montierten magnetischen Fühlkopfes vorbeilaufen, erzeugen Impulse, die eine transistorisierte Schaltung in einen ihrer Frequenz proportionalen Ausgang umwandeln. Auf diese Weise ist die Zeigerstellung von der Geschwindigkeit der Welle, auf der die Ansätze befestigt sind, abhängig. Das Tachometer benötigt eine Gleichstromversorgung von 24 V; da der höchste Stromverbrauch jedoch nur 100 mA ist, kann man Primärbatterien verwenden.

Die eisenhaltigen Ansätze können durch Angiessen, Bearbeitung oder Hartlöten an einer Welle, einer Riemenscheibe, einem Stirnzahnrad oder einer Schwungscheibe entstehen. Fühlköpfe für Einsatz mit solchen Ansätzen können um die Peripherie herum angeordnet werden, oder ein anderer Fühlkopf kann die Drehung eines geschlitzten Stückes, das in die Mitte eines überhängenden Wellenendes eingeschraubt ist, abtasten.

Für Mehrfachanzeige kann ein Fühlkopf jede beliebige Anzahl von Anzeigegeräten treiben. Bei Verwendung eines Schalters mit niedrigem Kontaktwiderstand ist es auch möglich, Anlagen mit mehreren Fühlköpfen und nur einem Anzeigegerät zu betreiben.

EE 75 758 *für weitere Einzelheiten*

Zeit- und Vorgang-Registriergeräte

Bowmar Instrument Ltd, Sutherland Road, London, E.17

(Abbildung Seite 788)

Bowmar Instrument Ltd gibt bekannt, dass die Fertigung ihrer Zeitfolgeanzeiger Typ 1440 und Vorganganzeiger Typ 1989 nunmehr in Grossbritannien angelaufen ist.

Diese Geräte eignen sich durch ihre ausgezeichnete Lesbarkeit, Digitalanzeige auf vier Trommeln und Miniaturausführung vor allem für Einsatz an Bord von Flugzeugen. Die auf schwarzen Trommeln eingravierten weissen Zahlen sind 3,2 mm hoch und zeigen bis zu 9999 Stunden an. Der Aussendurchmesser ist 17 mm, die Länge 46 mm und das Gewicht ungefähr 50 g. Unter den verschiedenen lieferbaren

Einbauausführungen sind u.a. solche für dichten Einbau in Frontplatten. Die Leistungsaufnahme ist 1,1 W bei 115 V, 400 Hz, einphasig. Für Anschluss an 28V— sind Wechselrichter mit Abmessungen von ungefähr 25 × 25 × 38 mm lieferbar.

Der Vorgangsanzeiger ist in ein ähnliches Gehäuse eingebaut und kann bis zu 9999 zählen. Für den Eingang sind entweder 24...28 V— bei 2 W oder 115 V, 400 Hz bei 2 W erforderlich. Höchstzählrate ist 10 Vorgänge je Sekunde bei 50 ms Impulsdauer.

Die beiden obigen Geräte sind nichtrückstellbar und den Anforderungen des Pflichtenblattes MIL-M-7793C entsprechend konstruiert.

EE 75 759 *für weitere Einzelheiten*

Überlastungsschutz-Thermistor

Standard Telephones & Cables Ltd, Connaught House, 63 Aldwych, London, W.C.2

Der erste einer neuen STC-Serie von Thermistoren mit positivem Temperaturkoeffizienten ist jetzt lieferbar. Es ist der PTC120, der speziell als Schutzelement für Motoren, Generatoren, Transformatoren und andere Ausrüstungen gegen übermässige Temperaturen, die die Isolation der Wicklung beschädigen könnten, entwickelt wurde. Die Schwellenwert-Nenntemperatur dieses Types ist 120°C.

Das Thermistorelement PC120 hat einen Durchmesser von nur 4,75 mm. Das Gehäuse dieses kleinen Bauelementes wird trommelentgratet, so dass es ohne Gefährdung der Lackisolation in Wicklungen eingelegt werden kann.

Die Anschlüsse des PTC120 kommen in 50 cm Standardlänge als verdrillte Doppelleitung. Andere Längen können geliefert werden. Für Drehstrombetrieb gibt es Mehrfachanordnungen des PTC120.

EE 75 760 *für weitere Einzelheiten*

Miniaturlötkolben

Antex Ltd, Grosvenor House, Croydon, Surrey

(*Abbildung Seite 788*)

Das Antex-Sortiment wurde durch das neue Modell C24ON ergänzt. Es handelt sich um einen Miniaturkolben für Netzspannungen, der zwischen 60 und 85 g wiegt. Er ist mit einer "Ferraclad"-Spitze ausgerüstet, deren Lebensdauer fünfmal so lang sein soll wie die üblicher vernickelter Spitzen und die ausserdem eine grössere Wärmekapazität und Wärmespeicherfähigkeit besitzt. Die Spitze lässt sich einfach und ohne Beschädigungsgefahr für den Kolben auswechseln. Der Kolben ist robust konstruiert und für Dauerbetrieb geeignet.

EE 75 761 *für weitere Einzelheiten*

Bausätze für abgeschirmte Räume

Belling & Lee Ltd, Great Cambridge Road, Enfield, Middlesex

(*Abbildung Seite 788*)

Zusätzlich zu ihrer Auswahl abgeschirmter Verkleidungen kann Belling & Lee Ltd nunmehr modulare Verkleidungen der Serie "100" liefern, deren Konstruktionsmerkmale man normalerweise nur bei auf Bestellung hergestellten Anlagen findet.

Die neue Serie stützt sich auf austauschbare Baugruppen mit Metallgerippe, die sich schnell in einen abgeschirmten Raum mit bis zu 100 dB Abschwächung bei Frequenzen von 1...500 MHz zusammenbauen lassen.

Zu den Merkmalen gehören leitende Dichtungen für leistungsfähiges HF-Zusammenkitten von Feldern und eine Tür neuer Konstruktion mit Schiebesitz, die automatisch festen Kontakt macht und in Position bleibt.

Modulare Fensterfelder können in Streckmaterial oder in Honigwabenform mit einer Lage gegen HF abgeschirmter Kanäle geliefert werden. Letztere arbeiten nach dem Prinzip der "Hohlleiter unter Grenzfrequenz", die die HF unter der Grenzfrequenz stark dämpfen, jedoch in Vorwärtsrichtung gute Sicht erlauben und der Belüftung keinen Widerstand bieten.

Für elektrische Stromversorgung, Fernsprechen und andere Dienste gibt es ein hochbeanspruchbares Durchführungsfeld mit geeigneten Filtern.

Ein Hauptvorteil der neuen Serie ist die Anpassungsfähigkeit in Bezug auf Raumgrösse und -anordnung, denn die Verkleidungen können ohne Schwierigkeit modifiziert und jederzeit ausgebaut werden. Man kann sie auch jederzeit auseinandernehmen und kompakt einlagern oder zu einem anderen Standort transportieren. Das Material wird in Bausatzform geliefert, und zwar mit Baugruppen, Holzfussbodenfeldern mit Linolfliessen, sowie allen erforderlichen Muttern, Bolzen und HF-Dichtungen. Die Standardfelder sind 2,24 m hoch und 1,12 m breit; die Türen und Durchführungsfelder für die Dienste kommen alle in dieser Modulgrösse. Für Sonderverwendung sind halbe Moduln, Arbeitstische usw. erhältlich, und es gibt eine umfangreiche Auswahl elektrischer Filter.

EE 75 762 *für weitere Einzelheiten*

Selbstabgleichender Schreiber

Vertrieb: Wessex Electronics Ltd. Midsomer Norton, Somerset

(*Abbildung Seite 788*)

Der Latronics Recordette-4 ist ein selbstabgleichender Hochleistungs-Potentiometerschreiber für vielseitigen Einsatz. Das Gerät bietet kleine Abmessungen, Tragbarkeit, Streifenvorschübe in breiter Auswahl. Schreiber mit und ohne Tinte, sowie viele andere Eigenschaften einschliesslich einer Genauigkeit von ½% des Bereiches bei Empfindlichkeit von ½% des Bereiches, Vollautomatik und Endwertabgleich in ½ s.

Die Eingangsstufen Typ 1 und 2 haben einen für Vollausschlag zwischen 10 und 100 mV regelbaren elektrischen Bereich. Die Eingangsstufe Typ 3 hat einen zwischen 10 mV und 100 V einregelbaren Bereich, und die Eingangsstufe Typ 4 ist für Anschluss von Thermoelementen ausgelegt und ermöglicht Direktregistrieren von Temperaturen.

Recordette-4 wurde für zuverlässigen Langzeitbetrieb konstruiert. Direkter Zugang zu Streifentrieb, Schreibstift, Eingangsstufe und allen Bedienelementen vereinfacht die Bedienung erheblich.

Streifenvorschübe sind zwischen 6,35 mm/h und 1220 mm/h erhältlich; der 100 mm breite Streifen kann entweder aufgewickelt oder für sofortige Besichtigung und Abreissen durch einen Schlitz im unteren Teil der Tür gespeist werden. Die Ausführung für Frontplatteneinbau erfordert einen Ausschnitt von 171,5 × 222 mm.

EE 75 763 *für weitere Einzelheiten*

Labor-Oszillograf

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(*Abbildung Seite 789*)

Telequipment Ltd hat den Oszillografen D43R für Einbau in 19"-Gestelle herausgebracht, der bei nur 178 mm Frontplattenhöhe 406 mm tief ist.

Der D43R wurde für breite Verwendung in Industrie, Laboratorien und Medizin entwickelt und wird mit einer Auswahl von fünf Verstärkereinschüben geliefert, darunter zwei neue.

Die 10-cm-Planschirm-Elektronenstrahlröhre mit Nachbeschleunigung des D43R wird mit 4 kV betrieben; die Zeitablenkung hat 18 vorwählbare, geeichte Schriebgeschwindigkeiten von 0,5 s/cm...1 µs/cm. Die Anstiegzeit ist 23 ns, die Eingangsimpedanz 1 MΩ, und das Gerät ist für Netzanschluss an 100...240 V (50...100 Hz) ausgelegt.

Es gibt Verstärkereinschübe A, B und C (Universal, Differential und für extrahohe Verstärkung) sowie die neuen Typen D und G. Type D, ein Hüllkurvenmonitor, ist ein abgestimmter Verstärker mit vier umschaltbaren Bereichen von 2,5...32 MHz. Auf Resonanz abgestimmt hat er einen Ablenkfaktor von etwa 1 V/cm, eine Modulationsfrequenz ist als ein Synchronisierausgang vorhanden, und der Eingangswiderstand ist 50 Ω.

Typ G ist ein kombinierter Breitband- und Differentialverstärker, dessen Differentialeingänge eine Unterdrückung von besser als 200:1 von 0...1 MHz (Sinuswelleneingang) haben, die bei 12 MHz nicht unter 30:1 fällt. Die höchste Eingangsspannung im 10 mV/cm Bereich ist 10 V_{SS}.

EE 75 764 *für weitere Einzelheiten*

Digitalvoltmeter-Ratiometer

Digital Instruments Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(*Abbildung Seite 789*)

Digital Instruments Ltd hat vor kurzem ihr umfangreiches Programm für Digitalinstrumentierung durch das Digitalvoltmeter DM2022 ergänzt. Dieses neue Gerät hat die Messgenauigkeit und Zuverlässigkeit des Voltmeters DM2020, ist jedoch mit einer längeren Skala (39999) und einer eingebauten Ratiometerschaltung ausgerüstet.

Das neue Instrument hat eine Messunsicherheit von 0,0025 Prozent des Skalenendwertes, ±0,01 Prozent der Anzeige und das hohe Auflösungsvermögen von $2{,}5 \times 10^{-5}$. Der Messumfang 0...2 kV wird in fünf Bereichen überstrichen, deren niedrigster eine Empfindlichkeit von 10 μV hat. Die Anzeige kann extern so umgesetzt werden, dass sie von Analogspannungseingängen kg/cm², °C usw. direkt angibt. Die Eingangsimpedanz ist höher als 25 GΩ für die beiden niedrigsten Bereiche und 10 MΩ für die höheren. Der Eingang ist erdfrei, so dass in der Signalquelle gegenwärtige Gleichtaktsignale unterdrückt werden.

Die Ratiometerschaltung ist eine bedeutende Einrichtung des Gerätes, die z.B. für Potentiometereichungen und Analogrechnermessungen Vergleichsmessungen zwischen Spannungsverhältnissen und externen Bezugsspannungen ermöglicht.

Das DM2022 enthält zahlreiche Einrichtungen, die von den älteren DM-Instrumenten her bekannt sind, u.a. Maximum- und Minimum-Arbeitsweise mit voller Genauigkeit und Dekadenausgänge zum Treiben von Druckwerken, Lochern usw. in jedem beliebigen von sechs Codes.

EE 75 765 *für weitere Einzelheiten*

Nahwirkungsschalter

W. H. Sanders (Electronics) Ltd, Gunnels Wood Road, Stevenage, Hertfordshire

(*Abbildung Seite 789*)

Die britische Atombehörde (UKAEA) wird seit einiger Zeit mit diesen Bauelementen beliefert, die jetzt auch an die Industrie abgegeben werden. Die Betätigung des Schalters hängt von der Vervollständigung eines magnetischen Kreises durch externe eisenhaltige Stoffe ab; das Gerät kann eingesetzt werden, um die Gegenwart von Arbeitsstücken oder Maschinenteilen nachzuweisen, ohne dass diese berührt werden. Diese Anforderungen bestehen, wo es praktisch unmöglich ist, herkömmliche Schalter zu verwenden, z.B. in Gegenwart korrodierender oder entflammbarer Dämpfe, Staub oder Flüssigkeit, oder wo radioaktive Stoffe gehandhabt werden.

Der Nahwirkungsschalter wird betätigt, wenn sich eisenhaltige Stoffe innerhalb 12,5 mm vom Suchkopf befinden. Der wirksame Abstand kann mittels eines Empfindlichkeitspotentiometers eingeregelt werden. Der Suchkopf ist vollkommen imprägniert und in Epoxydharz gekapselt und wird durch die in den meisten Industrieverfahren anzutreffenden Bedingungen nicht beeinflusst.

Zwei Ausführungen des Nahwirkungsschalters sind lieferbar, und zwar Typ FSPS/042 und der kleinere Typ PDM/3171, der für Einsatz mit einem bistabilen Transistorverstärker geeignet ist. Das kleinere Modell lässt sich in bestehende Transistorlogiksysteme einbauen oder auch getrennt als normales Schaltbauelement verwenden.

EE 75 766 *für weitere Einzelheiten*

Wellenformgenerator

Servomex Controls Ltd, Crowborough, Sussex

(*Abbildung Seite 789*)

Als Weiterentwicklung des NF-Wellenformgenerators LF 51 hat Servomex Controls Ltd ein neues Gerät herausgebracht, das 90 deutliche Wellenformen im Ton- und Servofrequenzbereich erzeugen kann und trotzdem so klein und leicht ist, dass es sich in einer Hand tragen lässt. Dieser bemerkenswerte Fortschritt in der Konstruktion von Instrumenten dieser Art wurde durch weitgehende Verwendung von Transistoren und gedruckten Schaltungen erzielt.

Der neue Generator LF 141 hat einen Frequenzbereich von 0,002 Hz...2 kHz und erzeugt alle 90 Wellenformen ohne irgendwelche Zusatzgeräte. Die Ausgangsspannung von 0...10 V_s wird in drei Bereichen mit Höchstspitzenspannungen von 10 V, 1 V und 0,1 V abgegeben.

Das Gerät hat eine direkte Frequenzskala, und die Ausgangsspannung kann entweder symmetrisch um den Nullleiter oder polarisiert entnommen werden. Das ergibt zusätzliche Wellenformen und einige verzögerte Wellenformen. Ein besonderes Merkmal des Gerätes ist, dass man es tasten kann; Anlegen einer externen Verriegelungsspannung stoppt den Hauptgenerator und erlaubt Abgabe von ein oder mehreren Schwingungen.

Ein weiteres Merkmal ist der völlig getrennte Hilfsoszillator zum Triggern des Hauptgerätes, was eine grosse Mannigfaltigkeit neuer Wellenformen erzeugen kann. Da der Hilfsoszillator ausser dem scharfen Triggerimpuls auch eine Dachschräge erzeugt, besteht die Möglichkeit, einen Oszillografen mit ausreichender Gleichstrom-Triggerpegelregelung zu synchronisieren, um den Einsatz der Hauptwelle zu sehen. Das entspricht einer in den Oszillografen eingebauten niederfrequenten Verzögerungsleitung mit einer Höchstverzögerung von einigen Sekunden.

Die Grundschaltung des Integrators ist neu und beseitigt die Verzögerung, die normalerweise nach Anlegen eines Triggerimpulses auftritt. Verwendung einer Baugruppe mit veränderlicher Phase erzeugt durch Abwesenheit der flachen Spitzen der Dreieckwellenform des Grundgenerators eine verbesserte Sinuswelle. Mit diesem Gerät kann die Welle bei jedem gewünschtem Phasenwinkel beginnen.

EE 75 767 *für weitere Einzelheiten*

Wärmedruckbindung

G. V. Planer Ltd, Windmill Road, Sunbury-on-Thames, Middlesex

(*Abbildung Seite 790*)

Die kürzlich von G.V. Planer Ltd angekündigte Ausrüstung ist ein vielseitiges Werkzeug zur Herstellung von Wärmedruckbindungen in Forschung und Fertigung. Die Anlage ist besonders für das Binden von Anschlüssen auf Dünnfilm-Mikroschaltungen sowie Halbleiter-Werkstoffen und -Elementen geeignet.

Konstruktionsmerkmale der Apparatur sind neue Präzisionsmikromanipulatoren mit X-, Y- und Z-Bewegung und spezielle "Schachfiguren"-Bedienelemente mit Doppelachse, unabhängige thermostatische Regelung des Bindemeissels und Trägers, sowie eine anpassungsfähige Anordnung des Mikromanipulators und der Trägerposition auf der Arbeitsfläche.

Die Anlage ist mit zwei Mikromanipulatoren ausgerüstet, von denen einer den thermostatisch geregelten Bindemeissel mit der veränderlichen Belastung handhabt, der andere die Drahtvorschubeinrichtung mit Druckluftbetätigung. Letztere ist für Standardspulen, die Drähte zwischen 0,013 mm und 0,25 mm Durchmesser aufnehmen, konstruiert. Auf der Düse der Vorschubeinrichtung sitzt ein Drahtschneider. Bindemeissel und Träger können auf Temperaturen bis zu 600°C gehalten werden, und man kann dabei über der Arbeitsfläche Schutzgas einführen.

Ein Doppelmikroskop mit Stereogummilinse und von 7 bis 30× oder 14 bis 60× kontinuierlich regelbarer Vergrösserung und veränderlicher Beleuchtung ist eingebaut.

Betriebsbauteile und Steuerpult sind auf einem Stahltisch mit 75 × 120 cm Oberfläche aufgebaut.

EE 75 768 *für weitere Einzelheiten*

Überschall-Prüfgerät

Ultrasonoscope Co. (London) Ltd, Sudbourne Road, Brixton Hill, London, S.W.2

(*Abbildung Seite 790*)

Das als Baumuster V angekündigte neue Überschall-Prüfgerät ist volltransistorisiert und mit einer 127-mm-Elektronenstrahlröhre bestückt, die genaue Dickenprüfungen und Beobachtung im Aussendienst erlaubt.

Einstellen lassen sich Frequenzen von 5, 2,5 und 1,25 MHz, und die Verstärkung wird mit einem geeichten Abschwächer geregelt. Die höchste Ablenkgeschwindigkeit ist 2 μs, so dass 6,35 mm dicker Stahl über den ganzen

Bildschirm dargestellt werden kann. Die Zeitablenkung hat kontinuierlich regelbare Verzögerung. Die Impulsfrequenz ist zwischen 50...1000 Hz veränderlich, um Geisterechos zu beseitigen. Durch die Metallstruktur hervorgerufene Streuungen kann man mittels eines Stufenschalterreglers unterdrücken. Es besteht die Wahl zwischen gleichgerichteter und nicht gleichgerichteter Darstellung, da sich für eine Anzahl schwieriger Aufgaben die nicht gleichgerichtete Anzeige als notwendig erwiesen hat.

Die Hauptbedienelemente sind auf der Frontplatte, die übrigen nach Verwendungszwecken gruppiert an der Seite angeordnet. Die aufladbare Batterie hat eine Lebensdauer von 7 Stunden und lässt sich im Handumdrehen auswechseln. Auf Wunsch kann das Baumuster V auch vom normalen Netz betrieben werden, für welchen Fall das Batterieabteil einen Netzanschluss aufnimmt. Die Aussenabmessungen sind 241 × 178 × 546 mm und das Gewicht einschliesslich Batterien ungefähr 12,7 kg.

EE 75 769 *für weitere Einzelheiten*

Impulsgeber

Vertrieb: B. & K. Laboratories Ltd. 4 Tilney Street, Park Lane, London, W.1

(*Abbildung Seite 790*)

Der Mehrzweck-Impulsgeber 6613 der Texas Instruments gibt zusammenfallende positive und negative Ausgangsimpulse, deren Frequenz durch einen internen Zeitgeber, ein externes Signal oder einen manuellen Einmalschalter bestimmt wird. Die Impulsdauer ist bis zu einem Tastverhältnis von 90% regelbar. Die Verzögerung kann für Ausgang vor oder nach dem Synchronisierungsausgang eingestellt werden. Die Ausgangsamplitude ist von 0...10 V einstellbar, und für den Ausgang ist ein Überlastungsschutz vorgesehen. Es wird durchgehend Druckkartenkonstruktion verwendet; das komplette Gerät hat Aussenabmessungen von nur 216 × 216 × 305 mm und wiegt ungefähr 4.5 kg.

Die Taktimpulsfrequenz ist in sechs dekadenabgestuften Bereichen von 15 Hz bis zu 15 MHz wählbar, die Amplitude zwischen 0 und 10 V an 50 Ω. Anstiegs- und Abfallzeiten sind in drei Bereichen von 10 ns bis zu 10 ms einstellbar, die Verzögerung in sechs Dekadenbereichen von 30 ns...30 ms. Die Impulsdauer kann man in sechs Dekadenbereichen zwischen 30 ns und 30 ms regeln. Der Synchronisierungsausgang ist 3 V positiv mit Anstiegs- und Abfallzeiten von 15 ns und einer Impulsdauer zwischen 50...80% Tastverhältnis in Abhängigkeit von der Frequenzeinstellung. Ein Minimum von 2 V positiv bei 20 ns Anstiegszeit und 30 ns Mindestdauer ist für den externen Treibereingang erforderlich.

Eine Drucktaste auf der Frontplatte kann einen Einzelimpuls auslösen, und ein Frontplattenschalter erlaubt Ausblenden von Ausgangsimpulsen, “Aus” bei 0 Volt und “Ein” bei 7 V positiv.

EE 75 770 *für weitere Einzelheiten*

Vor-Abzählrelais

Vertrieb: D. Robinson & Co. Ltd, 5-7 Church Road, Richmond, Surrey

(*Abbildung Seite 791*)

Das Rodene Vor-Abzählrelais ist eine Einrichtung zum Zählen jeder beliebigen Anzahl von Gleich- oder Wechselstromimpulsen. Jeder ankommende Impuls erregt eine Magnetspule, die ein Schaltrad einmal weiterrastet. Sowie die vorgewählte Zahl erreicht ist, wird ein Mikroschalter betätigt, der irgendwelche weitere Impulse zu einer getrennten Rückstellspule (wenn vorhanden) oder in die externen Schaltungen des Benutzers weiterleitet.

Jedes “Vor-Abzählrelais” ist für das Zählen jeder Zahl bis zu 20 einstellbar. Zwei oder mehrere Relais können in Serie geschaltet werden, so dass der Rückstellimpuls des Relais A die Schrittschaltspule des Relais B erregt, was Dekadenzählen ermöglicht.

“Vor-Abzählrelais” sind entweder für automatische oder manuelle Rückstellung erhältlich. Bei automatischer Rückstellung bleibt ein Zeiger während des Zählens in der vorgewählten Position, bei manueller Rückstellung dreht sich der Zeiger mit dem Schaltrad, so dass die Anzahl der empfangenen Impulse jederzeit sichtbar ist.

Die automatische Rückstellung wird durch einen an eine getrennte Spule gelegten Einzelimpuls eingeleitet, durch den eine Sperrklinke freigegeben wird, worauf eine Feder die Rückstellung des Relais innerhalb 50 ms bewirkt.

Ein oder zwei Schaltersätze können anstelle oder zusammen mit dem Mikroschalter benutzt werden, um während des zyklischen Ablaufes eine Reihe externer Schaltungen anzusprechen. Jede Impulsdauer über 25 ms erregt das Relais; da in allen Relais die Verlustleistungen hoch sind, um schnelle Betätigung zu erzielen, soll man die Spulen nicht kontinuierlich erregen. Die Höchstschaltrate ist 5 Schritte je Sekunde und die Lebensdauer des Relais in der Grössenordnung von 3 Millionen Betätigungen. Alle Modelle sind mit “AMP-Faston”-Anschlüssen lieferbar.

EE 75 771 *für weitere Einzelheiten*

Frequenznormal

General Radio Co. (U.K.) Ltd, Marlow Road, Bourne End, Buckinghamshire

(*Abbildung Seite 791*)

Der Frequenznormal-Oszillator 1115-B ist der neuste im Programm der General Radio für Quarzkristall-Frequenznormale. Das neue Gerät arbeitet mit einem gegetterten 5-MHz-Kristall (5ter Oberton) in einem dreistufigen Oszillator. Kristall, Oszillator und die automatische Verstärkungsregelung sind zusammen in einem Thermostaten mit Proportionalregler untergebracht. Die Oszillatorfrequenz wird für Ausgangssignale bei 1 MHz und 100 kHz unterteilt und kann auch bei 5 MHz entnommen werden.

Die Kurzzeitkonstanz (Abweichung bei Vertrauensgrenzen von 95%) ist 1×10^{-11} (eff) über eine Mittelwertzeit von 10 s, 100×10^{-11} (eff) über 300 µs. Eine typische Alterung ist unter 1×10^{-10} je Tag nach einem Jahr. Die Spektrumlinienbreite liegt unter 0,25 Hz im X-Band. Die Rauschaustastung ist —145 dB für 1 Hz Bandbreite bei 5 MHz Ausgang.

In diesem neuen Normal finden durchweg aktive Silizium-Halbleiterelemente Verwendung, und seine Bauart wurde unter besonderer Berücksichtigung ortsveränderlichen Einsatzes und strengster Anforderungen entwickelt. Eine eingebaute Nickel-Kadmium-Stromversorgung schaltet sich bei Ausfall des Wechselstromnetzes automatisch ein und speist das Normal für 35 Stunden.

EE 75 772 *für weitere Einzelheiten*

Klimaschrank

Fisons Scientific Apparatus Ltd, Loughborough, Leicestershire

Im Rahmen ihres Programmes für Umgebungstest-Ausrüstungen hat die Weyco-Abteilung der Fisons Scientific Apparatus Ltd einen neuen Klimaschrank angekündigt.

Der neue Schrank Typ CM/49 hat folgende Eigenschaften:

Temperaturbereich: —25° C. ... +150°C regelbar auf ±1°C.

Feuchtigkeitsbereich: von 12% bis zum Maximum auf ±1% relative Luftfeuchtigkeit regelbar.

Volumen der Arbeitskammer: 0,26 m³.

Ein patentiertes Luftströmungssystem gewährleistet gleichmässige Verteilung der feuchten Luft in der Arbeitskammer. Feuchtigkeit wird durch Einspritzen zerstäubten Wassers eingeführt. Eine hermetisch dichte Kühleinrichtung ist eingebaut, die nicht nur Betrieb bei niedrigen Temperaturen, sondern auch die Erzielung eines Feuchtigkeitsgehaltes unter dem der Umgebung ermöglicht. Alle Bedienelemente sind vorn am Schrank angeordnet und alle elektrischen Elemente durch Schwenken des Bedienfeldes nach oben zugänglich. Zur Standardausrüstung gehört ein Zweikanalschreiber für kontinuierliches Protokollieren von Temperatur und Feuchtigkeit.

Der neue Schrank wird die Forderungen von Klima-Ingenieuren und Qualitätskontrollpersonal in den Sektoren Verpackung, Medikamente, Farben und Chemie zum grössten Teil befriedigen.

EE 75 773 *für weitere Einzelheiten*

Hochkonstanter HF-Empfänger
The Marconi Co. Ltd. Chelmsford, Essex
(Abbildung Seite 791)

Ein hochkonstanter Universal-HF-Empfänger H 2301 mit Frequenzbandüberstreichung von 500 kHz...30,5 MHz ist nunmehr von der Marconi Co. Ltd erhältlich. Der wirtschaftliche und kompakte Empfänger ist für Empfang von AM-, CW- und ESB-Signalen geeignet und kann mit dem Frequenzumtast-Zusatz H 5011 für Einkanal-Fernschreiberkreise eingesetzt werden.

Der Frequenzbereich wird in 30 Teilbereichen von 1 MHz mit 100 kHz Überlappen und einer Eichunsicherheit von innerhalb 1 kHz überdeckt. Die Skaleneichung ist linear und so angeordnet, dass die Frequenz durch Ablesen getrennter MHz- und kHz-Skalen bestimmt wird. Zur jederzeitigen Überprüfung der Eichung ist ein 100-kHz-Kristalloszillator eingebaut.

Anwendung der Zweifachumsetzungstechnik erzielt hohe Selektivität, wobei Quarzsteuerung im ersten Hilfsoszillator eine für diese Empfängerklasse ungewöhnlich hohe Stabilität ergibt.

Die zweite Zwischenfrequenz arbeitet mit 500 kHz, und diese Frequenz kann man über einen Kathodenfolger für Zusatzausrüstungen entnehmen. Von den vorhandenen fünf Selektivitätseinstellungen sind zwei mit Bandfiltern ausgerüstet. Für selektiven CW-Empfang ist ein Tonfrequenzfilter vorhanden.

Zur Erleichterung der Bedienung sind alle Einstellelemente auf der Frontplatte angeordnet und ein Lautsprecher eingebaut. Getrennte Tonausgänge mit unabhängiger Verstärkungsregelung sind vorhanden.

In dem dauerhaft konstruierten Empfänger, der aus sechs Chassis und zwei zusammengeschraubten doppelt abgeschirmten Kästen besteht, werden durchweg Bauelemente höchster Qualität verwendet. Er wird für Gestelleinbau oder im Tischgehäuse und für Anschluss an jedes Standard-Wechselstromnetz geliefert.

Der Frequenzumtast-Adaptor ist volltransistorisiert und erlaubt Umtastung zwischen 200 und 850 Hz bei Telegrafiegeschwindigkeiten bis zu 100 Bauds.

Der Marconi-Empfänger H 2301 war ursprünglich ein Eddystone-Empfänger, der entsprechend einem Marconi-Pflichtenblatt für ESB- und FI-Betrieb modifiziert wurde.

EE 75 774 *für weitere Einzelheiten*

Mehrbereich-Messgerät
Oxley Developments Co. Ltd. Priory Park, Ulverston, Lancashire
(Abbildung Seite 791)

Das vor kurzem von Oxley Developments Co. Ltd auf dem britischen Markt eingeführte Mehrbereich - Messgerät "Monoc" ist ein Tascheninstrument ungewöhnlicher Konstruktion.

Bei einem Messumfang von bis zu 1000 V≃ ist die Messunsicherheit 1,5 Prozent für Gleichspannungen und 2,5 Prozent für Wechselspannungen. Strombereiche sind bis zu 1 A— und 10 A∼ vorhanden. Ein Ohmmeter, das nicht justiert zu werden braucht, kann bis zu 2 MΩ messen. Für Bereicherweiterung gibt es zahlreichen Zubehör.

Die neuartige Konstruktion ermöglicht Einbau eines Spiegelmessgerätes mit ungefähr 81 mm Skalenlänge in ein Gehäuse von nur etwa 150 × 100 × 50 mm Abmessungen.

Das Messwerk ist sowohl durch Dioden wie Sicherungen geschützt; es ist unzerstörbar und kann durch sorgfältiges Ausgleichen in jeder Lage benutzt werden.

Die leichte Konstruktion des Instrumentes wird durch ein elastisches Armband voll ausgenutzt, denn das Gerät lässt sich am Handgelenk tragen, wenn man nicht am Arbeitsplatz oder einer flachen Oberfläche arbeitet, und lässt beide Hände frei.

Da nur ein Bereichschalter vorhanden ist, kann er mit einer Hand bedient werden, was die bequeme Einsatzmöglichkeit noch erweitert.

EE 75 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein Mehrkanalregistriergerät für Baumbeanspruchungen von C. I. Sach

Zusammenfassung des Beitrages auf Seite 736-739

Dieser Beitrag beschreibt ein Aufzeichnungssystem für Beanspruchungen, das speziell für Untersuchungen der Reaktion von Bäumen entwickelt wurde, aber auch allgemeine Anwendungsmöglichkeiten im Labor und Aussendienst hat, wo Protokolle einer Anzahl von Dehnstreifen erwünscht sind. Es ist batteriebetrieben und so geeicht, dass Schwankungen der Stromversorgung die Eichung nicht beeinflussen. Das beschriebene Gerät hat eine Nichtlinearität von weniger als 1% des Höchstausganges und ist zum Messen dynamischer Beanspruchungen mit Frequenzkomponenten bis zu 10 Hz geeignet.

Ringförmige Verhältnistransformatoren—ihre Ersatzschaltungen und Messen ihrer Parameter von A. J. Binnie und T. R. Foord

Zusammenfassung des Beitrages auf Seite 740-745

Dieser Beitrag gibt einen Überblick, in dem veröffentlichte Analysen von Ersatzschaltungen, die die nützlichsten Typen ringförmiger Transformatoren für niederfrequente Präzisions-Brückennetzwerke darstellen, zusammengebracht werden. Verfahren zum Messen der in diesen Ersatzschaltungen verwendeten Parameter werden ebenfalls gegeben. Die Einflüsse der Wicklungskapazität werden nicht berücksichtigt, so dass die in diesem Beitrag gegebenen Schaltungen und Gleichungen nicht unbedingt bei höheren Tonfrequenzen gültig sind.

Ein transistorisiertes, selektives Kapazitäts-Widerstands-Netzwerk von T. H. Appleby

Zusammenfassung des Beitrages auf Seite 746-749

Man ist sich jetzt allgemein darüber klar, dass die Mikroelektronik die Zuverlässigkeit elektronischer Geräte schliesslich verbessern und ihre Abmessungen verringern wird. Festkörperschaltungen und Dünnfilmtechnik schliessen Verwendung von Induktivitäten unter 30 MHz aus, so dass transistorisierte Kapazitäts-Widerstands-Netzwerke gefunden werden müssen, die den LC-Frequenzgang simulieren. Die beschriebene Schaltung ist das Ergebnis einer Untersuchung solcher Schaltungen.

Ein neuartiger Integrator und seine Anwendung zur Aufrechterhaltung der Empfindlichkeit von Null-Messystemen
von H. C. Bertoya

Zusammenfassung des Beitrages auf Seite 750-755

Ein beschriebener neuartiger Integrator ist das elektronische Analog eines Stellmotors.
Eine spezielle Anwendungsmöglichkeit wird dann besprochen, bei der der Integrator zur Aufrechterhaltung der Empfindlichkeit eines Servo-Nullmessystems eingesetzt wird, das die Kenngrössen eines unbekannten Elementes oder "Prüflings" mit einem Normal- oder "Bezugs"-Element vergleicht.

Xerografie für Rechner-Schnelldrucker von R. H. Dagnall und P. F. T. C. Stillwell

Zusammenfassung des Beitrages auf Seite 756-760

Die an kommerzielle Drucker für Rechnerausgänge zu stellenden Forderungen werden kurz besprochen.
Ein Druckwerk für direkt oder indirekt mit einem Digitalrechner gekoppelten Betrieb, das xerografisch ein Endlosprotokoll erstellt, wird beschrieben. Gewählte Festdaten (z.B. Formularlinien) werden von einem Filmspeicher aus projiziert und gleichzeitig mit den variablen Daten, die als Zeichen auf einem Elektronenstrahlröhrenpaar dargestellt werden, gedruckt. Formwahl und -layout sind programmgesteuert; Monitorschaltungen entdecken eindeutige Fehler oder Versagen der Maschine und verursachen deutliches Markieren fehlerhafter Formen.
Das Papier hat eine konstante Vorschubgeschwindigkeit von ungefähr 20 cm/s, die Ausdrucken mit 2800 Zeilen je Minute erlaubt.

Eine elektronische Einrichtung zum Zählen und Registrieren von Insekten in der landwirtschaftlichen Forschung
von E. J. Brach und W. J. Mason

Zusammenfassung des Beitrages auf Seite 761-764

Zur Bekämpfung von Insekten in Weizen und Mehl müssen Entomologen ihre Gewohnheiten studieren, z.B. wie viele Insekten in einem Einheitsmass Weizen oder Mehl sind und zu welcher Zeit sie ihre grösste Tätigkeit ausüben. Eine Ausrüstung zur Bestimmung dieser Faktoren wird beschrieben. Sie besteht aus einer Insektenfalle, einem optischen Messwertwandler, Zählschaltungen, Zeitschaltungen und Druckwerk, die die Anzahl der einen mit Weizen oder Mehl gefüllten Behälter auf der Suche nach Wasser verlassenden Insekten registrieren.

Spannungskonstanthaltung in elektronischen Hybridapparaten von P. Visontai

Zusammenfassung des Beitrages auf Seite 764-765

In elektronischen Apparaten, die sowohl mit Röhren als auch Transistoren bestückt sind, entstehen Sonderprobleme in der Bereitstellung der Speisespannung für die Transistoren.
Wenn eine der Röhren so viel Gleichstrom führt, wie für die transistorisierten Teile des Apparates benötigt wird, so kann man mit einer Stromversorgung auskommen.
In einer beschriebenen Anordnung ist es—praktisch ohne Extrakosten—möglich, die niedrige Speisespannung einer Transistorengruppe gegen Schwankungen in der Gleichstromaufnahme einer anderen Transistorengruppe, die durch dieselbe Stromversorgung gespeist wird, zu stabilisieren.

Rauschspektrum-Messungen an Tunneldioden im Frequenzbereich 5 kHz . . . 10 MHz von R. A. Giblin

Zusammenfassung des Beitrages auf Seite 766-769

Rauschmessungen an Germanium-Tunneldioden für 1 mA Höchststrom zwischen 5 kHz und 10 MHz und im Vorspannungsbereich 0 . . . 450 mV werden unter besonderer Berücksichtigung des negativen Leitwertgebietes der Kennlinien beschrieben. Eine über den Gesamtbereich anwendbare Vergleichsmethode wurde angewendet und besondere, bei Rauschmessungen auftretende Schwierigkeiten besprochen. Das Problem der Stabilisierung der Diode wird eingehend behandelt und eine Schaltung gegeben, die gleichzeitiges Messen der statischen Kennlinien und des negativen Steilheitsleitwertes ermöglicht. Das Ergebnis lässt darauf schliessen, dass das Spektrum für Vorspannungen über 300 mV einem 1/f-Gesetz folgt und langsam in ein lineares Spektrum bei Spannungen unter 50 mV übergeht.

Ein Einstufenverstärker für wiederholte Analogberechnungen von R. E. King

Zusammenfassung des Beitrages auf Seite 770-771

Ein einfacher Gleichspannungsverstärker mit einer Verstärkung von ungefähr 2000 in der offenen Schleife wird beschrieben und eine Analyse seines Frequenzganges gegeben. Der Verstärker ist vor allem für wiederholte Analogberechnungen im Tonfrequenzbereich geeignet.

Ein linearer Transistor-Kippgenerator mit vernachlässigbarer Verweilzeit von J. K. Moss

Zusammenfassung des Beitrages auf Seite 772-775

Ein linearer Kippgenerator mit gut definierter Flanke und vernachlässigbarer Verweilzeit wird beschrieben. Der Sägezahn wird durch einen externen Trigger synchronisiert. Durch Entwurf geeigneter Schaltkreise wird die Verzögerung zwischen dem Anlegen des Triggerimpulses und dem Start des Sägezahns sehr klein gehalten. Positiv- und negativgehende Ablenkungen können erzeugt werden.

Ein linearer Frequenz-Spannungswandler von W. P. O'Grady

Zusammenfassung des Beitrages auf Seite 776-778

Eine Halbleiterschaltung, die Frequenz auf lineare Art in Spannung umsetzt, wird besprochen. Methoden zur Verbesserung der Phasencharakteristik zwischen Eingangsfrequenz und Ausgangsspannung, sowie Wege zur Herabsetzung der der Ausgangsspannung überlagerten Welligkeitsspannung werden in Betracht gezogen.

Ein 50-kHz-Kaltkathoden-Dekadenzähler von R. D. Ryan

Zusammenfassung des Beitrages auf Seite 779

Ein einfacher n-p-n-Multivibrator ist eine wirtschaftliche Treiberschaltung für ein gasgefülltes 50-kHz-Kaltkathoden-Dekadenzählrohr. Schwarzsteuerdioden an den Hilfselektroden gewährleisten konstante Schaltimpulspegel.