ELECTRONIC ENGINEERING

VOL. 36 No. 438

AUGUST 1964

Commentary

'RADIO RESEARCH 1963' has recently been published and is the report of the work carried out last year by the Radio Research Station of the Department of Scientific and Industrial Research.

As is well known, the Radio Research Station's efforts are largely concentrated on essentially geophysical problems and for many years the work of the Station has been devoted mainly to studies of the propagation of radio waves over a large range of distances and over practically the whole radio frequency spectrum. In its turn this work has led to practical and theoretical investigations of the ionized and un-ionized parts of the atmosphere through which the waves travel and by which they are affected in transit. The knowledge which these investigations brings is very valuable and is contributing to a better appreciation of the fundamental processes at work in the everyday uses of radio waves and in the uses to which they will be put in the future: to understand the physical principles involved in transmission through a medium is important, since the knowledge so obtained can be applied over the whole range of uses of radio waves.

While a great deal of the Radio Research Station's work is of a purely theoretical and fundamental nature, the Radio Research Board, whose major duty is to advise the Director of the Radio Research Station on the content of the research programme, is concerned to ensure that investigations of a more practical nature are also included in the programme. To this end investigations are in hand towards understanding more clearly what determines the magnitude of the field strength laid down by a transmitter when the radio-wave energy reaches the receiver by way of the ionosphere.

During recent years the Station has become increasingly involved in international scientific ventures and notable here is the part it is playing in the U.K. contribution to the International Quiet Sun Years (I.Q.S.Y.) enterprise and in space research.

The I.Q.S.Y. project is a follow-on to the International Geophysical Year (I.G.Y.). During the I.G.Y. held in 1957-58 geophysicists spread widely over the earth co-operated in an international programme of observations. This period of observation, which had been planned to coincide with a maximum in the sunspot cycle, proved to be exceptionally fortunate as it happened that, on this occasion, the maximum was greater than any since the start of observations some two hundred years ago. Never before had observations been made with such advanced techniques or so widely over the earth and the addition to scientific knowledge was enormous. However, in many cases comparable observations were not available near sunspot minimum. To remedy this omission another period of international observation known as the International Quiet Sun Years has been organized to occupy the period near sunspot minimum and will run from January 1964 to January 1966. During the period of the I.Q.S.Y. the Radio Research Station will take part in the international programme by repeating some of the observations it made during the I.G.Y. Routine vertical soundings of the ionosphere will be made at Slough and at the outstations in Singapore and Port Stanley (Falkland Islands) and special observations of ionospheric drift and absorption will be made at Singapore.

So far as space research is concerned the Radio Research Station's contribution is twofold: direct space research and the maintenance of 'space science services'. In all of the research work concerned with the ionosphere and solar-terrestial relations, the Station is making increasing use of rockets and artificial satellites. In total, about one-quarter of the Station's effort is devoted to the design and execution of experiments made with the help of rockets and satellites, while another one-quarter is devoted to providing 'space science services' to assist other workers in space science in the United Kingdom and elsewhere.

The Station is making contributions to the National space research programme through experiments in rockets fired from the range at Woomera in Australia; to the combined U.K.-U.S.A. programme through an experiment to be performed in the third joint U.K.-U.S.A. satellite; to the programme of the European Space Research Organization (E.S.R.O.) through an experiment to be performed in one of their first two satellites; and through a collaborative arrangement with the Canadian Defence Research Board, to the analysis of results obtained by their 'Alouette' topside sounding satellite.

The 'space science services' provided by the Station include the maintenance and operation of the N.A.S.A. Minitrack Station at Winkfield, the reception of telemetry at Singapore and in the Falkland Islands and the processing of telemetry data after transmission.

There is no doubt that the advent of new research techniques such as rockets, satellites and lasers have increased enormously the scope of the work carried out by the Radio Research Station while they have, in return, made a great deal of the routine investigations appear more meaningful than sometimes they have appeared in the past.

Demodulation Circuits for PAL Colour Television Receivers

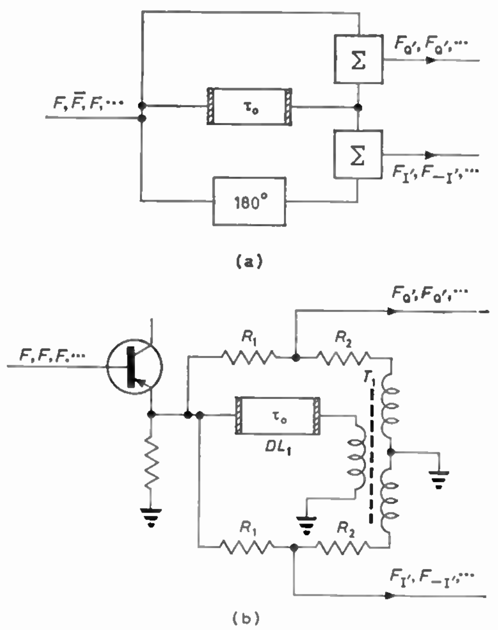
(Part 1)

By W. Bruch*

In this article some characteristic circuits are described for the demodulation of PAL colour signals. A switchable decoder permits processing the signal according to the PAL_s principle in addition to evaluation of the signal by the PAL_{DL} method; disconnexion of the PAL circuits allows operation with NTSC signals. A number of basic circuits and their practical design are described together with suggestions for reducing the cost of the delay line modulator.

(Voir page 576 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 583)

WITH the PAL colour television system[2,3] the two quadrature components making up the modulation signal are sequentially transmitted on alternate lines. Two demodulation methods are available for reception. With the delay-line demodulator, PAL_{DL} †, undistorted F_Q', and



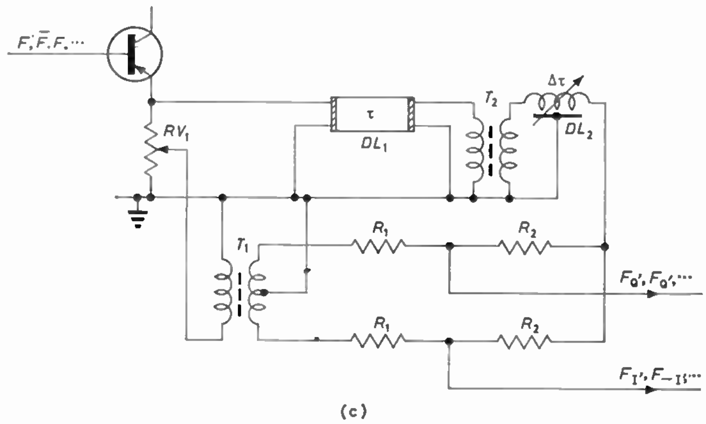


Fig. 1. Signal separating circuit of the PAL system

(a) block diagram of delay-line demodulator (synchronous demodulators are omitted)
(b) basic circuit
(c) basic circuit with additional delay-line for fine adjustment

corresponding F_I' signals are obtained by respective addition and subtraction of the direct and delayed carrier signals in their correct phase relation. It is also possible to use the 'simple PAL system', PAL_s, where suitable commutation of the NTSC synchronous demodulator allows the eye to perceive mean values of successive colour signals on the screen. For both these systems a number of circuits has been developed which has allowed the PAL system to be brought to its present state of development. Some of the more important of these circuits are described here in detail.

Demodulation with Respect to I' and Q'

Demodulation whereby the I' and Q' chrominance signals impressed on the transmitted carrier are retrieved at the receiver seems the most obvious method and indeed gives the best possible signal-to-noise ratio. The basic principle of this type of demodulation using a delay-line is illustrated in Fig. 1(a) and can be realized in practice by a variety of circuits. Circuits relying entirely on passive elements, i.e. those working without valves or transistors, for the formation of the sum and difference terms indicated in Fig. 1(a) have so far proved the most successful as they ensure long-term stability and temperature independence. Such a circuit is outlined in Fig. 1(b).

As is seen from Fig. 1(b) the sum ($2F_Q'$) and difference ($2F_I' - 2F_I', 2F' \ldots$) sub-carriers are each formed across a resistance network. Both these resistance matrices and the delay line are supplied from a cathode-follower or, better still, from an emitter-follower where transistor circuits are

* *Telefunken A.G., Hanover.*

† *Demodulators using delay lines are designated 'PAL_{DL}' or 'Standard PAL' while those without delay-lines are termed 'Simple PAL,' here abbreviated to 'PAL_s'[1]. The NTSC chrominance sub-carrier is written as F and its complex conjugate $\bar{F}$. F_Q', F_I', $F_{(B'-Y')}$ etc. represent the carriers modulated with the Q', I', $(B'-Y')$ etc. chrominance signals[2,3] c.f.*

used. Input and output impedances of the currently used ultrasonic delay line are in the region of 50Ω. It is possible, therefore, to use a step-up transformer (T_1) with band-pass characteristics at the output of the delay line to make good the voltage drop of approximately 16dB due to termination and insertion losses of the line. The transformer (T_1) secondary is wound bifilarly and great care is taken to ensure symmetry. One of these bifilar secondary windings gives the voltage for the summation signal (Q') while the other produces, with a 180° phase shift, the (I') signal for the difference channel. Because of the symmetrical arrangement of the transformer secondaries its primary circuit is completely independent of the output. Any interference reaching one of the secondaries, whether inductively or by conduction through the resistance chains, is cancelled by an equal and opposite signal which is bound to exist at the other secondary winding. The two sum and difference subcarriers appearing across the two resistance networks are then rectified in high input resistance synchronous demodulators thus yielding the I' and Q' video signals. The I' demodulator output is moreover switched through 180° on alternate lines. Switching is performed by a bistable multivibrator circuit which in turn is triggered by the line fly-back pulse. A transmitted reference signal ensures the correct phase relation between transmitter and receiver [1-3]. Although a number of suitable gated-beam demodulators, using for instance the 6GY6 or 6BN6 type valves are available, discussion is here confined to diode-ring-demodulators, also referred to as clamping-diode demodulators.

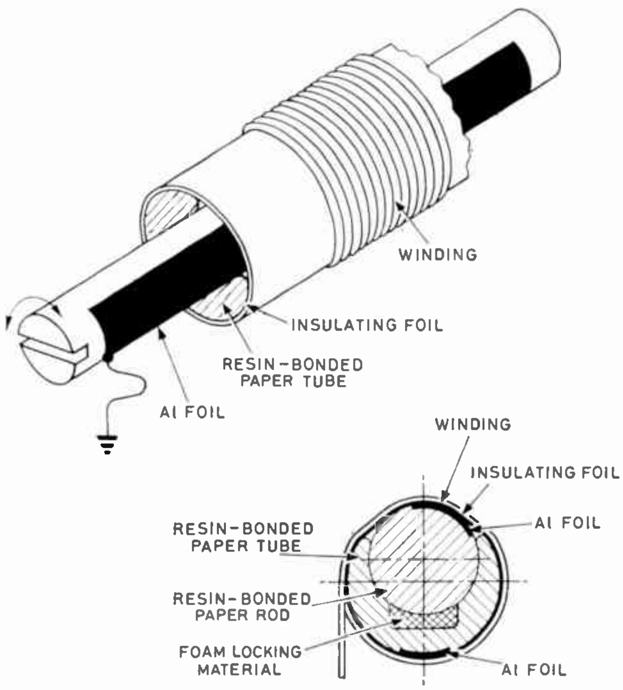


Fig. 2. Adjustable delay-line (DL_2)

To enable the delay time to be set accurately a small adjustable delay line DL_2 has to be added to the circuit of Fig. 1(b). It is not economically possible to produce such a small delay line with a low characteristic impedance of the same order as that of the ultrasonic line DL_1. It is, however, feasible to produce cheap and stable delay lines with characteristic impedances of 2kΩ or more. A transformer to match the two delay lines must therefore be provided so that the complete circuit takes the form of Fig. 1(c).

The construction of the adjustable delay line DL_2 is shown in Fig. 2. A rotatable rod of insulating material, alongside of which a strip of earthed metal foil is fixed, is eccentrically arranged within a single layer coil. Rotation of the rod allows the capacitance and thereby the delay time and characteristic impedance of the line to be varied over a sufficiently wide range. The line is terminated by a 2·2kΩ resistor equal to the mean value of the characteristic impedance of the delay line. The graph of Fig. 3 shows the possible adjustment of delay time expressed in terms of phase displacement ($\Delta\phi$) of the chrominance subcarrier. Since the delay adjustment is inevitably associated with a variation in characteristic impedance, the adjustable range must be limited to avoid serious mismatching with fixed termination resistances.

Where wider tolerances in the transmission times of the ultrasonic delay lines have to be expected, it is convenient to stock the fine adjustment lines in two or three different

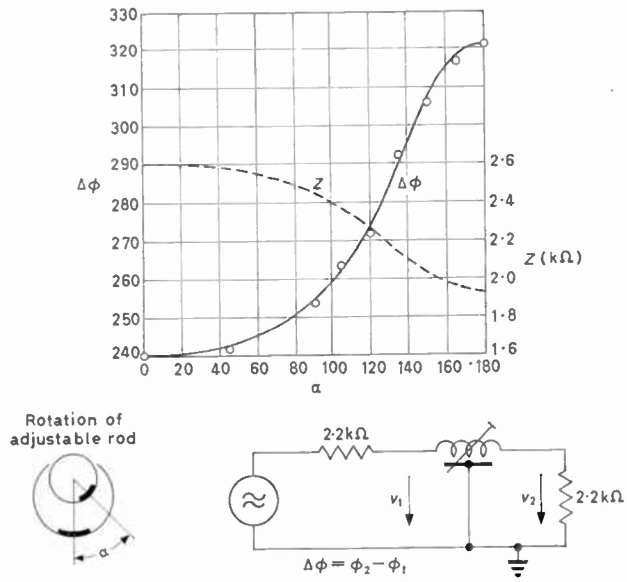


Fig. 3. Range of adjustable delay-line shown in Fig. 2 expressed as phase displacement $\Delta\phi$ at 4·43Mc/s against angular rotation α of adjustable rod

lengths so that the overall delay can always be correctly set. The temperature coefficient of the delay coil was measured to be $2 \times 10^{-4}/°C$ so that it can be entirely neglected[4]. The total delay time τ_0 must differ by exactly ±0·25 cycle from the actual line duration which, with quarter-line offset, corresponds to 283·75 cycles[5]. Adjustment is facilitated by the fact that the delay period may be either 284 or 283·5 cycles of the chrominance carrier. A half-cycle adjustment is obtainable simply by reversal of the terminal connexions to transformers T_1.

Fig. 4 shows the circuit of a transistor demodulator used for demonstrating both the PAL and NTSC systems*. With

---

** The PAL to NTSC switch-over facility has proved very valuable in comparative tests on the two systems particularly as no other parameters of the decoder need be changed. It is necessary, however, for such comparisons that the PAL and NTSC systems are worked on the same subcarrier frequency, i.e. an NTSC system with the quarter-line offset necessary for PAL[5] or, as is common now, PAL system with a line scan displaced by $0{\cdot}88 \times 10^{-3}$ cycle.*

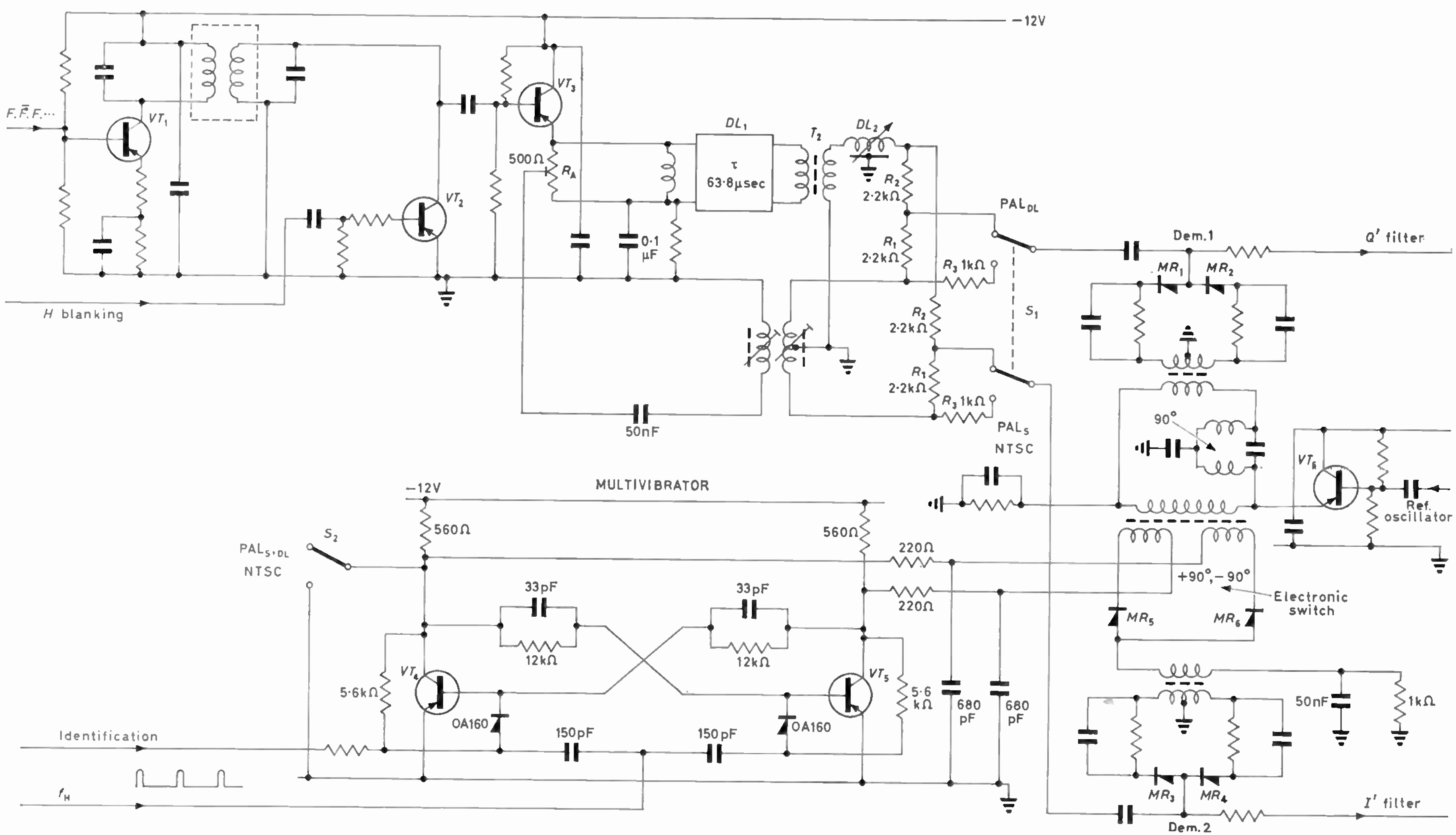


Fig. 4. PAL_{DL} demonstration circuit capable of being switched to PAL_S and NTSC operation

suitable I' and Q' filters good signal-to-noise ratio and low cross-modulation in the colour channels are achieved. If the transformer T_2 and the delay lines are disconnected by operation of switch S_1 (Fig. 4) the PAL_{DL} demodulator becomes a PAL_s circuit. If now, in addition, the bistable switching circuit is bypassed with switch S_2 (Fig. 4) in the down position, an NTSC demodulator is obtained. Thus it is easily possible not only to convert a PAL_{DL} demodulator to NTSC working in the factory but also to design sets for dual operation on, for instance, PAL and NTSC. The carrier commutating circuit using diodes MR_5 and MR_6 and the controlling multivibrator (VT_4 and VT_5) shown in Fig. 4 are identical with the circuits described[2] for PAL_s demodulators.

With switch S_1 (Fig. 4) in the PAL_s/NTSC position, matching resistors R_3 are brought into circuit so that the following diode demodulators should work from the same source impedance as they do when the delay lines are connected. This is important for a fair comparison between the properties of PAL_{DL} and PAL_s as it avoids changes in colour saturation on the receiver screen when switch S_1 is operated: this also applies to a comparison between the PAL_{DL} and NTSC systems, provided always, similar chrominance carrier offsets and modulation methods are used.

Fig. 5. Voltages at the outputs of PAL demodulator

(a) transmitted NTSC signal: I'—output voltage with incorrect alignment of PAL_{DL} receiver

(b) as (a) but with correct receiver alignment

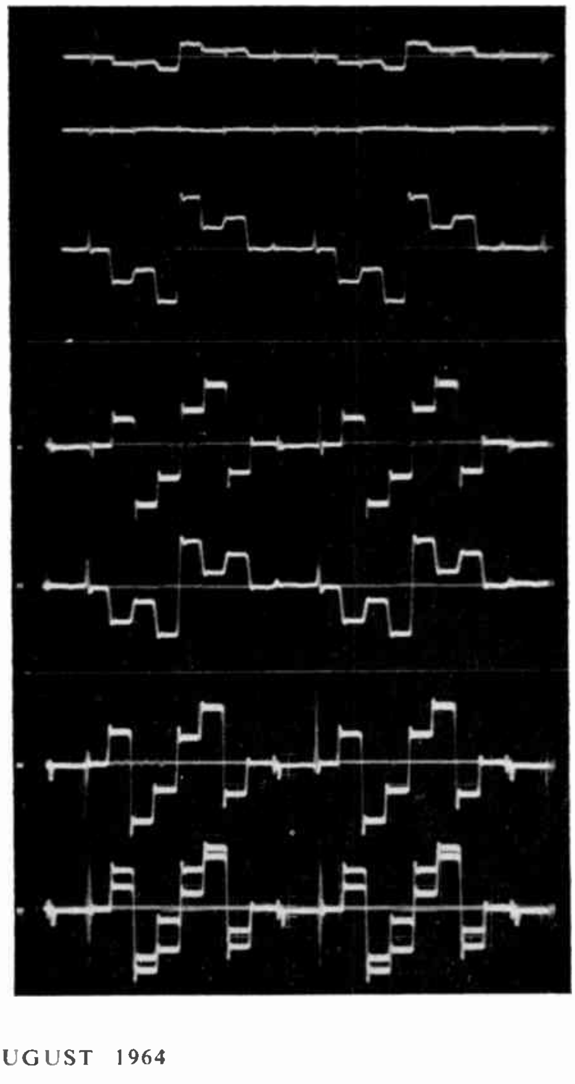
(c) transmitted NTSC signal: Q'—output voltage with correct receiver alignment (NTSC-Q'—signal)

(d) transmitted PAL signal: I'—output voltage of PAL_{DL} receiver

(e) as (d) but Q'—output voltage

(f) output of PAL_{DL} I'—demodulator with 30° error in reference carrier phase and correspondingly increased chrominance amplitude

(g) as (f) but measured on a PAL_s receiver



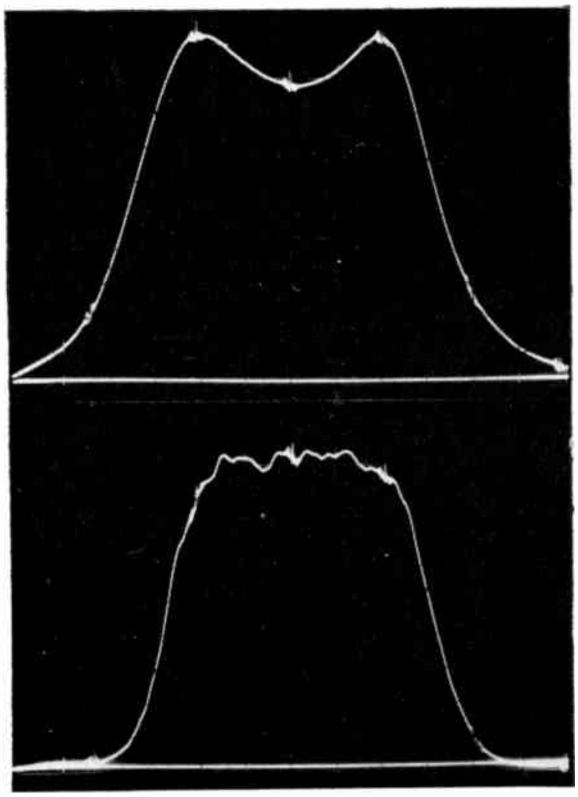


Fig. 6. Pass-band (logarithmic) of colour channel as measured with sliding frequency generator (wobbulator) applied to input

(a) direct path

(b) delayed path

In order to obtain colour-true picture backgrounds, the output valves of the three colour-difference amplifiers or, with monitor decoders, the R'-, G'-, B'- output stages should be clamped during line flyback*. Moreover to maintain the black-level accurately in all three channels the colour signal is blanked out during line fly-back by blocking of transistor VT_3 (Fig. 4) which removes both burst and noise during d.c. restoration.

Alignment of the delay-line bridge circuit is easily achieved with switch S_1 (Fig. 4) in the PAL_{DL} position by means of an NTSC modulated (colour test chart) signal. The output voltage of the I' demodulator as observed on a c.r.o. screen is gradually reduced to zero by alternate adjustment of resistor R_A and of the fine-adjustment delay line DL_2. Fig. 5(a) shows an I' demodulator output voltage such as might be expected before alignment, while Fig. 5(b) shows the c.r.o. trace for a properly balanced circuit with the corresponding Q' output voltage reproduced in Fig. 5(c). Thus aligned, both demodulators will now give the correct I' (Fig. 5(d) and Q' (Fig. 5(c) outputs when operated with PAL signals. Adjustment of the reference carrier

* *Diode clamps, now frequently replaced by switched transistors, are used for d.c. restoration to ensure maintenance of the true black-level as transmitted[6].*

phase is not critical, it can be checked with switch S_1 in the PAL_s position when a 'venetian blind' effect will be visible if the setting is incorrect. Fig. 5(f) shows the same signal as Fig. 5(d) but now with a phase error displacement of 30°. If the amplitude of the chrominance signal is suitably increased the same undistorted voltage as in Fig. 5(d) is obtained. Fig. 5(g) shows the I' signal in a PAL_s system with the phase error as obtained for Fig. 5(f). This shows several pairs of signals superposed in order to show the deviation of successive lines from the required shape of Fig. 5(d), the effect of which however averages out.

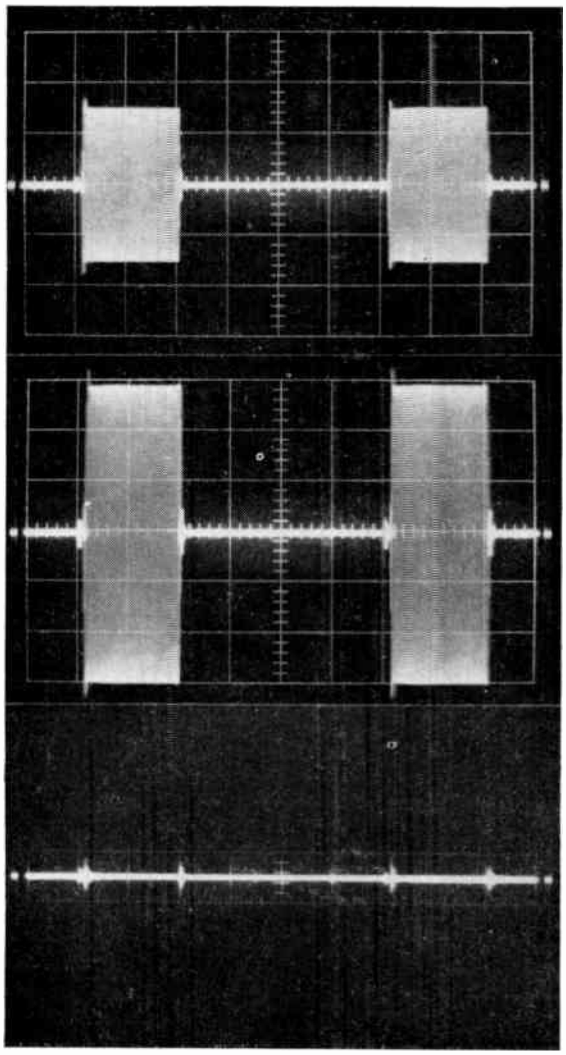
Fig. 6(a) and 6(b) show the pass-band of the colour channel through the direct and the delayed paths respectively as it was recorded with a sliding frequency generator (wobbulator) connected to the input. A section of a pulse modulated chrominance carrier at the input to the Q'-diode demodulator is shown in Fig. 7(a), a direct and a delayed signal being reproduced side by side for comparison. The sum ($2Q'$) and difference (I') of the pulse signals is illustrated respectively in Fig. 7(b) and 7(c).

Fig. 7. Section of pulsed carrier at input to Q'-demodulator direct pulse on left; delayed pulse on right

(a) before addition

(b) after addition at carrier frequency (2Q')

(c) the same pulses at input to I'—demodulator (subtraction)



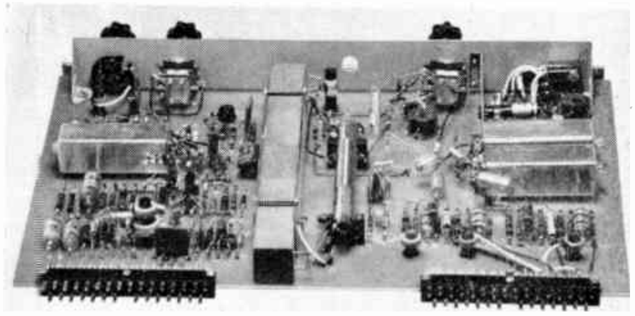


Fig. 8. Demodulator of Fig. 4 built as plug-in sub-assembly

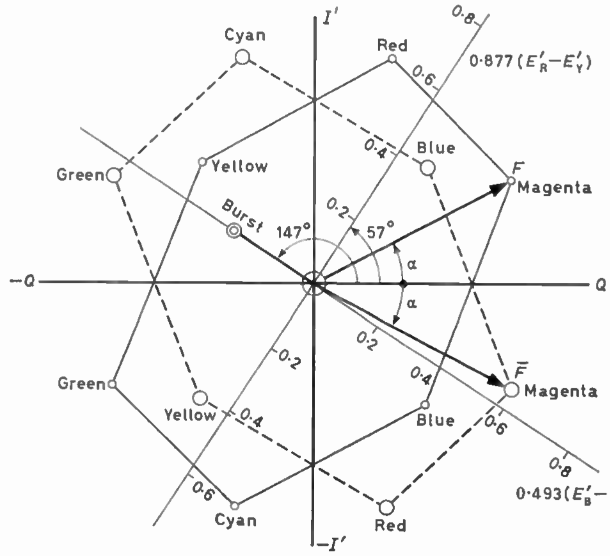


Fig. 9. Vector representation of PAL colour test chart

A demodulator according to Fig. 4 built as a plug-in sub-chassis is shown in Fig. 8.

Circuits Without Delay Line for Direct Demodulation of the $R' - Y'$ and $B' - Y'$ or the X' and Z' Signals in Simple PAL Receivers

I'-, Q' decoding and matrixing requires a high number of stages (for signal reversing); it is not favoured by manufacturers due to the high cost, which is increased further by the circuits for delay time equalization.

For NTSC receivers so-called equi-band demodulators are commonly used and particularly those which supply two of the picture signals, $(R' - Y')$ and $(B' - Y')$, directly while the $(G' - Y')$ signal is obtained from a simple matrix and reversing circuit. As in the PAL_{DL} receiver, these signals can also be produced directly by suitable demodulating circuits in PAL_s receivers without delay line. If Q' is chosen as the horizontal reference a convenient vector representation for the PAL system can be produced. With NTSC transmissions and a reference carrier lagging the datum Q' by 33° (referred to as $\angle(B' - Y')$ a synchronous demodulator produces a signal of 0·493 $(B' - Y')$ and a corresponding signal* of 0·877 $(R' - Y)$ with a

* *After demodulation, the signals $(R' - Y')$ and $(B' - Y')$ as well as X' and Z' appear multiplied by differing amplification factors. These are constants in the NTSC equation[2] which must be taken into account in the gain of the output stages following demodulation. For the sake of simplicity these constant multipliers will for the time being be neglected in the following discussion.*

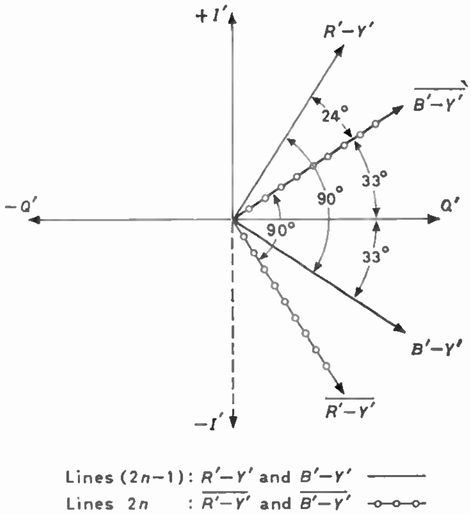


Fig. 10. Demodulation axes for odd and even lines with (R'−Y'), (B'−Y') demodulation in PAL$_S$ receiver

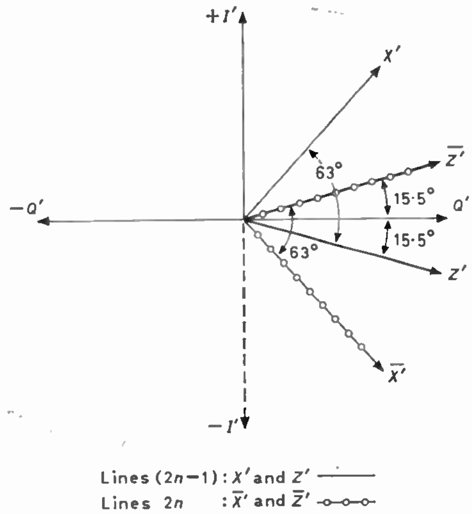


Fig. 11. Demodulation axes with X' and Z' demodulation in PAL$_S$ receiver

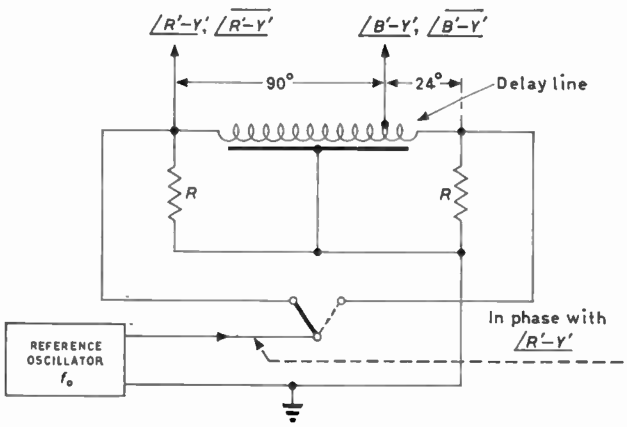
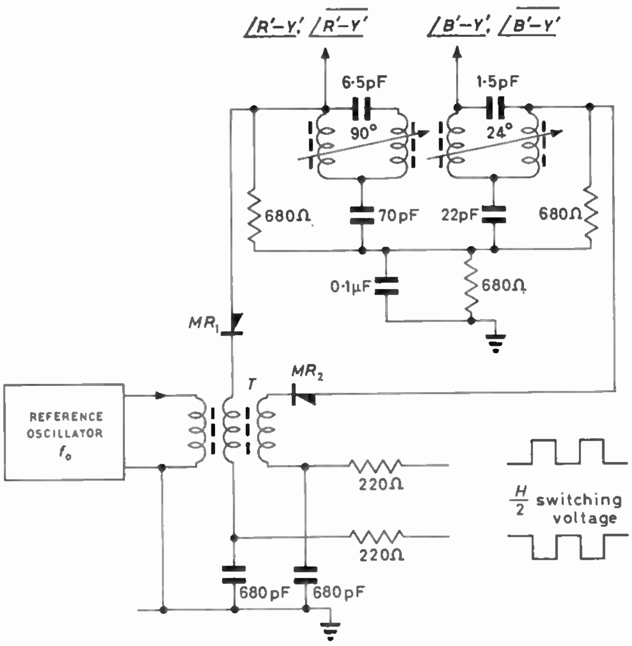


Fig. 12 (above). Mirroring circuit for reference carrier of (R'−Y')−, (B'−Y')− demodulation in PAL$_S$ receivers with delay line

Fig. 13 (below). Mirroring circuit for reference carrier of (R'−Y')−, (B'−Y')− demodulator with equalizer circuits



reference carrier phase angle $\angle(R'-Y')$ of 57° leading. If with PAL transmission the modulation axis I' is switched through 180° to $-I'$ on alternate lines, demodulation at the receiver must also be modified as otherwise incorrect colour signals would result. Fig. 9 shows a vectorscope diagram of a bar colour test transmitted by PAL. Since the endpoints of the colour vectors are mirrored in the Q' axis when the colour signal is transmitted in the complex conjugate form $\overline{F}$, the demodulation axes must be similarly mirrored as shown in Fig. 10. The 33° lagging phase $\angle(B'-Y')$ must now lead by 33° and because $\angle\overline{B'-Y'}$. Correspondingly the $\angle\overline{R'-Y'}$ phase which led by 57° must be made to lag by that angle $\angle(R'-Y')$. The new axes have the same relative positions in the new $-I'Q'$ quadrant as they originally occupied in the $I'Q'$ quadrant. Thus it becomes possible to demodulate with reference to $(R'-Y')$ and $(B'-Y')$, say, while the I' signal is being reversed at the transmitter provided the reference carrier phases are being appropriately reversed in the two demodulators. In Fig. 11 this concept is extended to the X' and Z' axes. X', Z'-demodulation, such as for instance RCA use in all their NTSC colour receivers, makes it possible to recover signals to operate the three output valves in particularly simple, stable colour difference circuits. According to the design of the circuits the X' and Z' axes are however differently defined. The phase angles shown in Fig. 11 correspond to the definitions given by Carnt and Townsend[8].

It is seen that it is necessary to demodulate with respect to four different demodulation axes in order to obtain two colour difference signals. It might be expected that rather complex circuits would be required to obtain these four phase references. This would indeed be the case if these four voltages were produced by four phase shifting circuits requiring a complicated electronic switch to connect the two demodulators in turn to pairs of the phase-shifters. Two simple examples will demonstrate, however, that this complication is unnecessary and that the phase switching can quite easily be attained with a degree of complexity not exceeding that required for I'-, Q'-demodulation. One of these examples, using demodulation axes according to

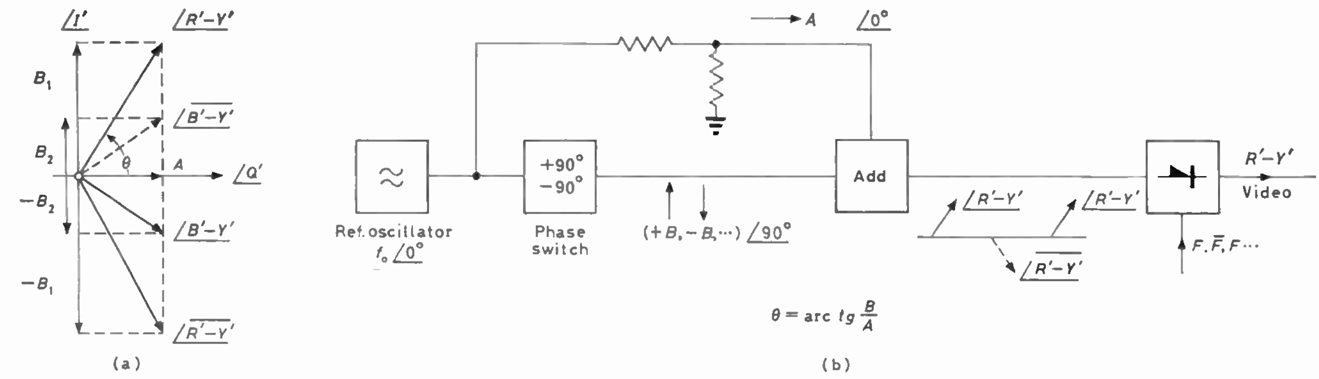


Fig. 14. Vector representation and block diagram for an alternative reference carrier mirroring circuit for (R′ − Y′) −, (B′ − Y′) − demodulation with PAL_S

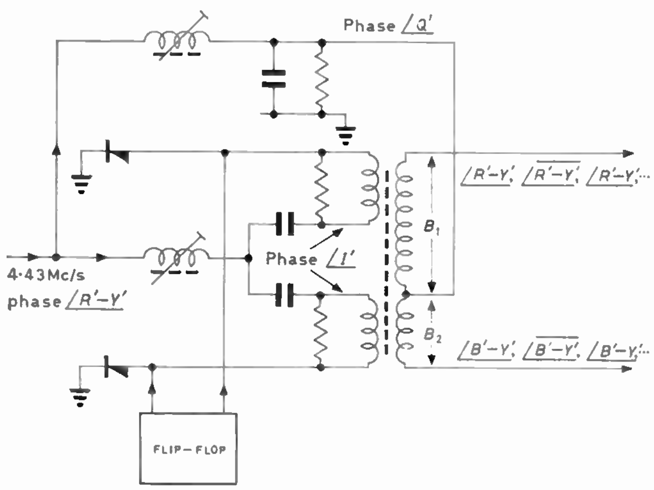


Fig. 15. Suggested circuit for Fig. 14

Fig. 10, is shown in Fig. 12. The reference carrier of phase $\angle(R'-Y')$ is switched alternately on to one of the terminated ends of a delay line. The voltages for the synchronous demodulators are picked off a tapping along the delay line. With the switch (Fig. 12) in the left-hand position, during those line intervals in which signal F is being transmitted, voltage of phase $\angle(R'-Y')$ is applied directly to the red demodulator, which works with respect to $R'-Y'$ as its reference, and then, after a shift of 90°, to the blue demodulator having $B'-Y'$ as its reference. The short right-hand section of the delay line is not used in this switch position. During those line intervals in which the chrominance carrier is modulated with the conjugate $\bar{F}$ signal the switch is on its right-hand contact. Then the reference carrier of phase $\angle(R'-Y')$ is applied to the other end of the delay line and it is taken off at the tap after a phase shift of 24° and thence fed to the blue demodulator which now demodulates with respect to the conjugate axis $\overline{B'-Y'}$ as shown in Fig. 10. After a further phase rotation of 90° the voltage is taken from the far end of the line to the red demodulator which thus works with respect to $\overline{R'-Y'}$. Thus the required mirroring of the demodulation axes is achieved. That part of the delay line which produces the 90° shift is thus operative during both switch positions.

This circuit has a number of advantages. The carrier amplitude does not affect the phase rotation; only a single pole switch is required; the same section of the delay line is utilized for the two 90° phase shifts; the phase delays can be trimmed very simply allowing for adjustment with the set in use.

A practical embodiment of the circuit using equalizer networks as phase-shifters and diode switches controlled by square waves is shown in Fig. 13. The transformer merely needs to have two similar secondary windings as against the push-pull transformer required for the circuit of Fig. 1. With the secondaries wound bifilarly equal amplitudes and phase relations are ensured. Equalizer networks as phase-shifters are preferable to delay lines as they are easier to design and manufacture for low characteristic resistances. Unlike other phase shifting circuits they can be loaded at intermediate points of the chain and can be used in either direction. Such a circuit for X' and Z' reference carriers was constructed for use with an existing RCA receiver and tested.

Another switching method also requiring only a single pole switch uses two reference carriers at right-angles one of which is switched through 180° on alternate lines. These quadrature signals are thus combined linearly to produce the required vector mirroring. One of the voltages is of amplitude A, say, and lies along the $\angle Q'$ axis, the other voltage, B, is switched from $\angle I'$ to $\angle -I'$ on alternate lines. Fig. 14 illustrates the principle for $(B'-Y')$ and $(R'-Y')$ demodulation. A completed receiver circuit as used by the Italian Television Services[8]—RAI—is shown in Fig. 15.

REFERENCES

1. EBU-Document Com.T (E) 75.
2. BRUCH, W. Farbfernsehsysteme-Uberblick über das NTSC-, SECAM- und PAL-System. *Telefunken Ztg. 36*, No. 1/2, 70 (1963).
3. BRUCH, W. Das PAL-Farbfernsehsystem. Prinzipielle Grundlagen der Modulation und Demodulation. *Nachrichtentech. Z, 17*, No. 3, 109 (1964).
4. SCHOLZ, W. Verzögerungsleitunger für den Laufzeitausgleich in Farbfernseh-Codier- und Decodiergeräten. *Radio Mentor. 30*, 111 (1964).
5. BRUCH, W. Wahl eines Präzisionsoffsets für den Farbhilfsträger im PAL-Farbfernsehsystem. *Telefunken Ztg. 36*, No. 1/2, 89 (1963).
6. WOLF, J. Zur Dimensionierung der Klemmschaltung. *Elektronische Rundschau. 9*, No. 2, 53 (1955).
7. CARNT, P. S., TOWNSEND, G. B. Colour Television. (Iliffe Books, London, 1961).
8. NTSC-Simple PAL Colour Television Receiver. RAI-Radio-televisione Italiana, Laboratorio Ricerche. (Oct. 1963).
9. U.S.A. Patent No. 2 868 872.
10. BRUCH, W., DIRKS, E. Die Synchronisierung der sequentiellen Umschaltungen beim PAL-Farbfernsehsystem. (To be published).

(To be continued)

An Economical Silicon Switching Transistor and its use in Ring Counters

By J. Palmer*

A silicon transistor ring counter was described by Ringrose and Bradley[1,3]. The design of the counter is discussed in detail and suggestions are made which enable an economical version to be made, the cost ratio being about a decade of the new for a single trigger pair of the older one. There are limitations in speed etc., and these are discussed. The counter uses a low cost silicon transistor and diode. A design example is given. The counter may be used as a straightforward counter or as a type of commutator, shifting a unique condition or set of conditions along a ring, one stage at a time.

(Voir page 576 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 583)

THIS article was stimulated by the development of a transistorized ring counter by Ringrose[1,3] following an earlier valve design[2].

With certain reservations it is believed that this circuit and many others of similar simplicity may be manufactured at a cost ratio of a decade for the price of a single trigger pair. This saving is achieved by the use of the silicon alloy transistor, the OC703. It is clearly acknowledged that there will almost certainly be limitations on the maximum speed of operation compared with the design by Ringrose, but it is felt that this is usually acceptable as long as these limitations are understood.

The Basic Circuit

The circuit functions as follows (with reference to Fig. 1(a)). The clear state finds all even number transistors 'on' and odd ones 'off' (except VT_1). The circuit is taken to a desired state for zero count by reversing the condition of one pair, in this case $VT_2\,VT_3$. This state is shown in Fig. 1(b). It will be observed that the collector current of VT_1 supplies the emitter current of VT_3 only (neglecting leakage currents).

A square wave into C_1 is differentiated by C_1R_{B1}. The negative excursion merely switches VT_1 harder 'on' but the positive excursion switches VT_1 'off' for a short time.

While VT_1 is 'off' the emitter of VT_3 is open-circuited causing its collector to rise towards $-V_{cc}$. On the way negative it takes VT_2 base into conduction, VT_2 collector falls towards earth and tends to cut off VT_3 base current. The action is accumulative until the pair have changed condition to VT_2 'on' and VT_3 'off'.

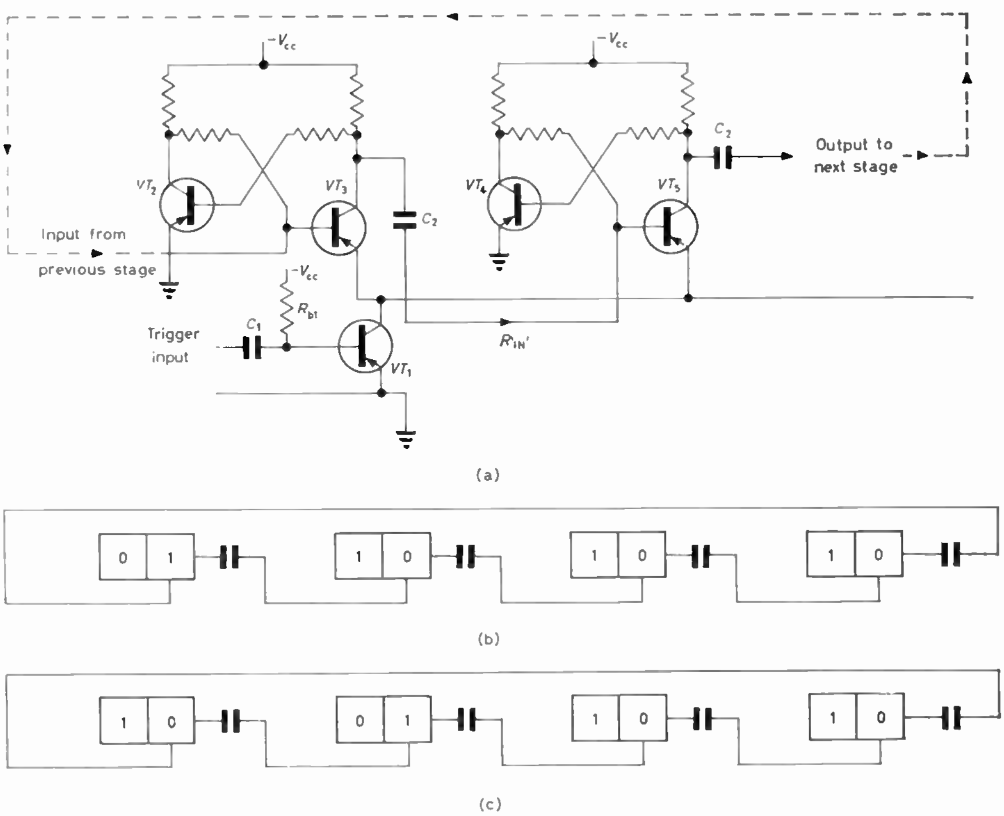
The collector of VT_3 is also loaded to a.c. through C_2 by the base circuit of VT_5 represented by R_{in}. It is arranged that the time-constant of C_2R_{in}, is longer than the time-constant of the differentiating circuit at the base of VT_1. Consequently when VT_1 is again allowed back into conduction its collector sees only one path to $-V_{cc}$ again, and this is now through the emitter of VT_5 due to the negative pulse still retained on VT_5 base. The resultant state is shown in Fig. 1(c).

The unique state has now shifted along to the next pair. The action is limited in repetition rate by the size of the capacitor C_2 required to couple the negative pulse to the following stage. It can cope with quite random pulses to a lower rate however, and of course will retain its unique state indefinitely in the absence of a pulse to VT_1.

Stable States

Fig. 2(a) shows the state of the non-unique pairs while (b) shows that of the unique condition.

Fig. 1(a). The ring (b) and (c) Shifting the unique state



* *Brush Clevite Co. Ltd.*

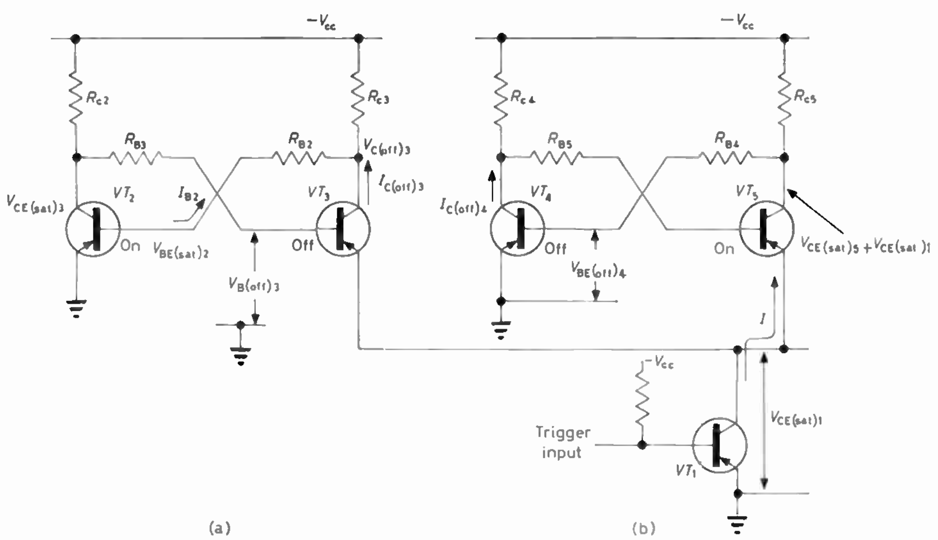


Fig. 2. The stable states

In Fig. 2(a) VT_2 is 'on', its collector saturating to a level sufficiently near to earth to keep VT_3 'off'. This is not too difficult because the emitter of VT_3 is at a small negative potential, i.e. $V_{ce(sat)}$ of VT_1. This is an important point which relaxes what might have been a stringent requirement, $V_{ce(sat)}$ of VT_2. Thus the first requirement is:

$$V_{CE(sat)2} < V_{B(off)3}$$

where $V_{B(off)3}$ is (maximum base-emitter potential allowed while keeping $I_{C(off)3}$ to a low value) minus ($V_{ce(sat)}$ of VT_1). Hence:

$$V_{ce(sat)2} < V_{be(off)3} - V_{ce(sat)1} \quad \ldots\ldots\ldots\ldots \quad (1)$$

The collector of VT_2 rises to a value determined by V_{cc}, R_{C3}, R_{B2}, $V_{BE(sat)2}$ and only slightly by the current out of VT_3, designated $I_{C(off)3}$. This now sets the qualifying condition for $I_{C(off)3}$, which should be considerably less than I_{B2} in the worst condition of VT_3. Assuming this to be so:

$$V_{o(off)3} \simeq \frac{(V_{cc} - V_{BE(sat)2})\,R_{B2}}{R_{B2} + R_{C3}} + V_{BE(sat)2}$$

It considerably aids in the definition of the base current of VT_2 if $V_{C(off)3}$ is large compared with the expected spreads of $V_{BE(sat)2}$.

Because:

$$I_{B2} = \frac{V_{C(off)3} - V_{BE(sat)2}}{R_{B2}}$$

In Fig 2(b) VT_5 is 'on', its collector saturating to a level which, even when added to $V_{CE(sat)1}$, is sufficiently near earth to keep $I_{C(off)}$ (the current out of VT_4 collector), small compared with the base current of VT_5, even in the worst condition of VT_4, VT_5 and VT_1.

The worst conditions are with: high $V_{ce(sat)5}$
high $V_{ce(sat)1}$
low $V_{be(off)4}$

because:

$$V_{be(off)4} > V_{ce(sat)5} + V_{ce(sat)1} \quad \ldots\ldots\ldots \quad (2)$$

It is now clear that a high $V_{ce(sat)1}$ is a help in Fig. 2(a) but a penalty in Fig. 2(b). As it might have some finite value the worst case overall is when it is high.

The Transistor

Although the Brush transistor type OC704 was considered, it was decided that the OC703 was more suitable. The former specification limits V_{ce} to 3V, by no means impossible as will later be seen, but nevertheless rather tight in view of the preference to keep $V_{o(off)3}$ large compared with the spread of $V_{BE(sat)2}$.

The guaranteed specification of the OC703 shows the following relevant points:

$V_{CE(max)}$ ($V_{BE} = +0{\cdot}5$V) $= -80$V

$I_{CBO(max)}$ ($V_{CB} = -10$V) $= -0{\cdot}1\mu$A. $I_{C(max)} = -50$mA.

$I_{B(max)} = -15$mA. $V_{CE(sat)}$ ($I_C = -7$mA; $I_B = -1$mA) $= -320$mV. h_{21e}. ($V_{CE} = -6$V; $I_C = -1$mA) > 10.

Although $V_{ce(max)}$ is defined at a positive base bias of 0·5V, its high value incates that it will be ample even when having a slight negative bias. This will be dealt with later although more from the viewpoint of $I_{C(off)}$ at a given small forward bias and a reasonable voltage at the collector.

$I_{C(max)}$ and $I_{B(max)}$ are ample.

$V_{CE(sat)}$ is quoted but not the equivalent spread of V_{BE}.

It was decided to investigate a random selection of OC703 for $V_{CE(sat)}$ and $V_{BE(sat)}$ at various values of I_C and I_B.

It happened that a batch of 63 was already being measured for another task. Its only preselection had been for a 90V I_{CBO} and 47V I_{CEO} and I_{EBO}, all at currents to be less than 1μA. This group was measured at 10mA I_C and 1mA I_B and the results are displayed in Fig. 3.

It will be observed from Fig. 3 that a reasonable yield

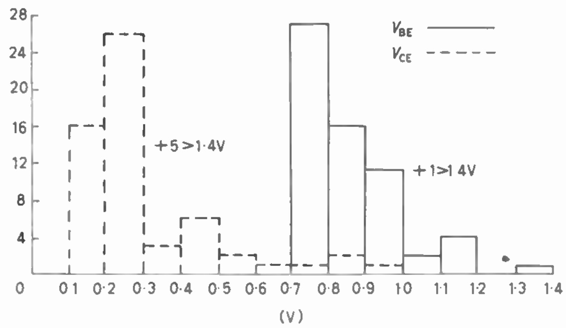


Fig. 3. Distribution of $V_{CE(sat)}$ and $V_{BE(sat)}$ at $I_C = -10$mA, $I_B = -1$mA for the OC703

would be obtained by placing limits of V_{BE} of 0·7 to 1·2V and a maximum V_{CE} of 0·5V. To be precise 47 transistors out of the 63 passed this test. The main fall-out is seen to be in bottoming voltage $V_{CE(sat)}$ which is not surprising considering the rigour of the test on such low gain devices. This is unlikely to be important because in practice such a high saturated gain will not be demanded. Indeed Ringrose[1] states "in fact it may be considered usual to use only one value of resistor throughout the whole counter", i.e. an approximate demanded gain of:

$$2 \times R_c/R_c \quad \text{(because } R_b = R_c\text{)}$$

or 2.

If, however, a higher gain is demanded, say 5, this will be satisfactory. Alternatively a gain of 10 could be obtained with greater yield by reducing the I_C towards 2mA, at about which current β is a maximum.

The next test was to check in the circuit of Fig. 4 what value of $I_{C(off)}$ could be expected with the worst $V_{CE(sat)}$ i.e. $-0{\cdot}5$V on the base. Preliminary checks indicated that this was too negative to switch off the OC703 satisfactorily so it was necessary to reduce the $V_{CE(sat)}$ to 0·35V max. This reduced the 47 passes to 41. The remarks of the previous paragraph still apply however.

The collector voltage was set at -27V as being as high as anyone is likely to require in this arrangement.

The results of this test are shown in Fig. 4. From this it

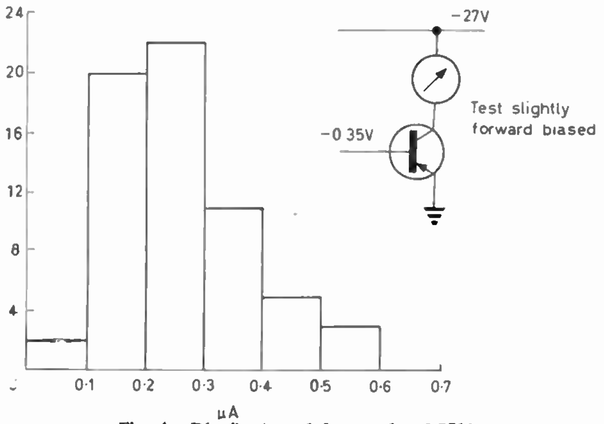


Fig. 4. Distribution of $I_{C(off)}$ for OC703 at (V_{CE} −27V V_{BE} −0·35V)

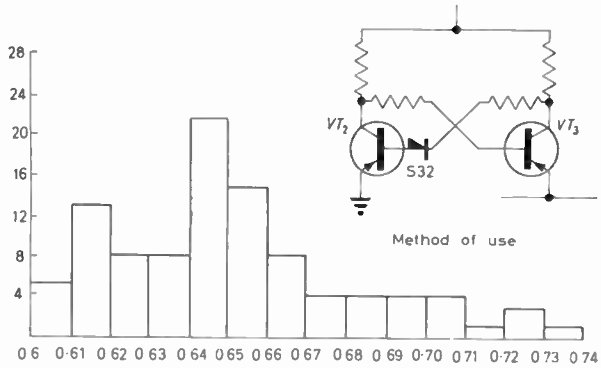


Fig. 5. Distribution of S32 diode forward voltage at 1mA

will be seen that all were below 0·8μA except one which was 2·5μA. This was therefore considered a safe value of $V_{BE(off)}$.

Summarizing the use of the OC703 in the two stable states it is seen that one may use a worse $V_{CE(sat)}$ of 0·35V and a spread of $V_{BE(sat)}$ between 0·7V and 1·2V.

Further to this a bottomed transistor at its worst will reduce the collector current of the worst following transistor to less than say 5μA.

All conditions now seem satisfactory except the worst case of inequality (2) in which:

$$V_{BE(off)4} > V_{CE(sat)5} + V_{CE(sat)1}$$

There are two approaches to this problem.

(a) Place a low cost silicon diode in series with the base of VT_2 and all even number transistors. This is illustrated in Fig. 5. The S32 diode is suitable and the spread of forward voltage drop for a hundred samples chosen from five different batches is also shown.

They were measured at a forward current of 1mA. This is an entirely satisfactory solution except that it increases the cost of the pair by about 20 per cent. The cost is still within the limits proposed in the introduction, i.e. a decade for the price of a trigger pair.

(b) Alternatively the original simplicity may be maintained by taking the risk of using random devices and expecting to get a possible reject pair for a foreseeable percentage of the production. Three devices have to be at their worst limit at the same time. The risk is not too severe. It is greatly reduced by the simple expedient of selecting a lower limit $V_{CE(sat)}$ device for VT_1 position. For a decade a ratio is required of 1 (for VT_1) to 20 (for 10 pairs). This is easily satisfied by a figure of 0·15V max (see Fig. 3).

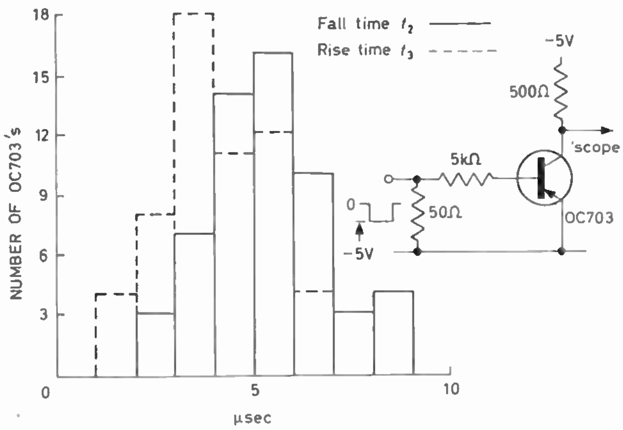


Fig. 6. Rise and fall times and test circuit

Transient Conditions

An admitted reduction in maximum repetition rate of the counter using OC703, to that of the original, is one of the limitations which must be understood. This is largely a function of the transistor and the data sheet is of little help in this respect. Consequently a functional test was applied to the same group of devices previously tested.

Fig. 5 shows the circuit in which a current of about 10mA is switched by a base current of < 1mA. (Spread of actual values may be calculated with reference to Fig. 3).

t_1 = hole storage time

t_2 = fall time (switching off)

t_3 = rise time (switching on)

The usual 10 and 90 per cent levels were taken.

The results were displayed on an oscilloscope.

The t_1's were all below 2μsec and typically 0·7μsec.

Under these conditions it is safe to say that $t_1 < 2{\cdot}5\mu$sec, $t_2 < 12\mu$sec and $t_3 < 10\mu$sec.

The effect of placing a 3 000pF 'speed up' capacitor across R in Fig. 6 was then measured and the times were reduced to $t_1 + t_2 < 0{\cdot}55\mu$sec and $t_3 < 0{\cdot}9\mu$sec.

It should not be assumed that switching times of this value are claimed in this counter. It must be recalled that this device has an f_α quoted as typically 0·5Mc/s and it is being compared with the SA496 which has a minimum f_1 of 7·2Mc/s. These readings are merely an attempt to get the two arrangements in perspective.

Clearing the Counter and Resetting Zero

The counter is cleared by taking VT_1 to the 'off' conditions for a period long enough for whichever of the coupling capacitors (C_2 etc) is to discharge and charge again. As VT_1 is kept off the next stage cannot come to its unique condition so the counter relaxes to its cleared state even when VT_1 is allowed to come 'on' again. The desired 'zero' is achieved by (for example) earthing the collector of the odd numbered transistor momentarily. A simple circuit illustrating these actions is shown in Fig. 7.

The two processes can be combined.

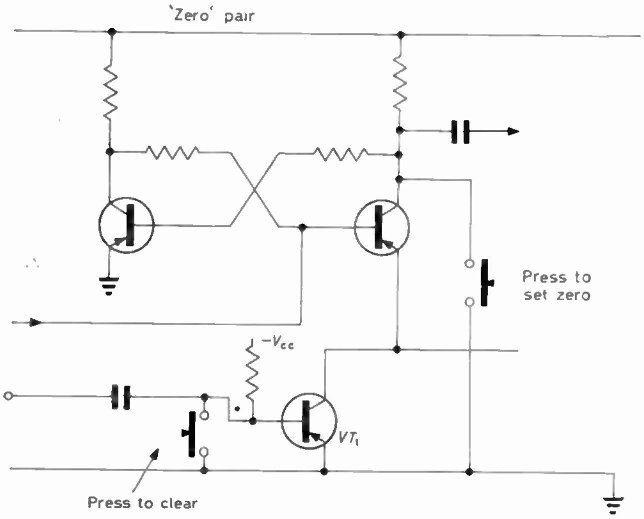


Fig. 7. Clearing and resetting

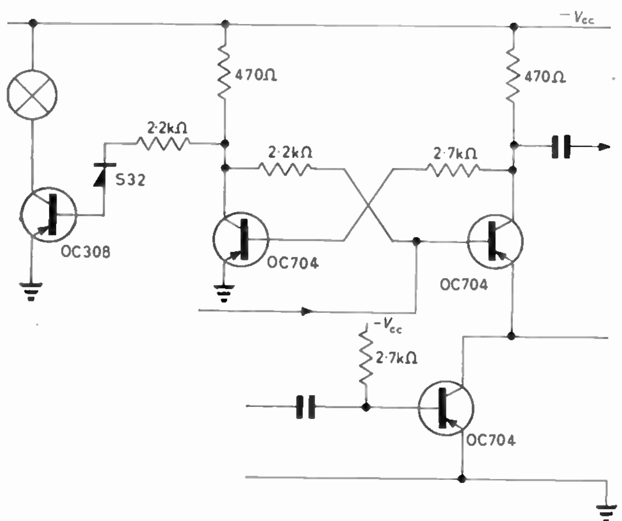


Fig. 8. Lamp driving circuit

Read-Out

This will largely depend on the type of application. If it is desired merely to indicate its own count this can be done by either a lamp in the collector of the odd numbered transistor in each pair or by a separate lamp driver. The former pre-supposes a low current lamp (liable to cost more than the rest of the circuit). The second method is shown in Fig. 8. As a germanium transistor was used as lamp driver a silicon diode was put in series with its base to allow the substantially higher $V_{ce(sat)}$ of a silicon device to switch it off.

The values of resistor used in Fig. 8 were in fact used with OC704. It is doubtful whether they would be satisfactory when used with all OC703 transistors because they demand a gain of about 7 at about 12mA even without the read-out transistor connected. Furthermore they exceed the voltage rating of the OC704. The OC701 is clearly a satisfactory solution.

In fact the circuit of Fig. 8 worked well with V_{cc} varied between −9V down to −2·4V. The lamps were rather dim in the latter state.

Application

When connected in the form of a ring the arrangement can be used as a kind of commutator in the same way as it is used to light lamps in sequence. It helps if the supply to be commutated has an earthy or at least commoned line. The scheme may also be used to shift a combination of 'on' and 'off' states along one stage at a time. The VT_1 transistor must be capable of bottoming well with additional collector current. This is usually achieved by reducing its base resistor value.

Design Procedure (see Fig. 9)

Two factors tend to dictate choice of component values. The load or read-out fashion and the requirement to define the base current fairly accurately in the face of

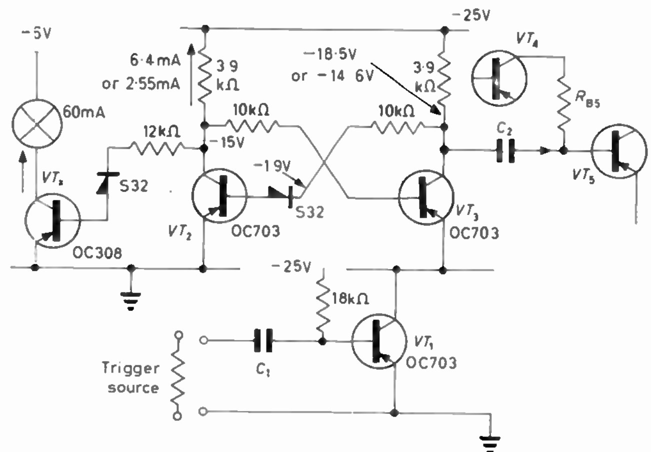


Fig. 9. Design example

rather a wide spread of base emitter voltage. For this example a 6V 60mA lamp is a suitable read-out. The OC308 is a suitable driver and has a gain of greater than 50 at 60mA, i.e. requiring an I_b of 1·2mA. The collector of VT_1 may conceivably reach −0·35V and the $V_{BE(sat)}$ of VT_3 may reach −1·2V. These infer that the voltage source for reasonable current definition should be 15V.

For reasons explained a silicon diode is required in series with the base of the lamp driver. The S32 diode is suitable and from Fig. 8 (the maximum V_f of about 0·7) + (the V_{be} of OC308 of 0·3V) = 1V. The collector of VT_2 when 'off' is 15V hence R_b drops 14V. This fixes R_b at 14V/1·2mA = 11·7kΩ, i.e. 12kΩ.

The collector load resistor is a compromise between two considerations.

(1) It is desirable to keep it large so that the current of the 'on' transistor is kept down to the optimum 2mA region.

(2) It is desirable to keep it small for the reason now shown: the capacitor C_2 couples the negative going step of VT_3 collector to the base of VT_5. At first VT_1 is open-circuit and so VT_5 cannot conduct. Hence C_2 only looks through R_{B5} to the collector of VT_4 which being 'on' is at about 0V. Thus to avoid overloading VT_3 collector and endangering the

switching of the first pair it is desirable to keep the load resistance of VT_3 small.

A compromise is reached by using a V_{cc} of 25V. Thus with R_{B2} in parallel with R_{B5} (the transitory state until VT_5 conducts) the ratio of voltage across R_{L3} to R_{B2} is 10:14. A degree of symmetry is acquired if the base current for VT_3 is about the same as that for the lamp driver VT_x, i.e. 1·2mA. Hence R_{L2} is set at 10V/2·4mA or 4kΩ. If it is set at 3·9kΩ then according to the ratio just calculated two base resistors in parallel = (3·9kΩ/10) × 14 = 5·5kΩ.

Hence R_{B2} etc. are set at 10kΩ.

Now with VT_2 'on' its collector current can be 25V/3·9kΩ = 6·4mV. The current in R_{L3} has two values, the transitory one of:

$$\frac{25\text{V} - (V_{be} + V_t)}{\left(\dfrac{10\text{k}\Omega \,.\, 10\text{k}\Omega}{10\text{k}\Omega + 10\text{k}\Omega}\right) + 3{\cdot}9\text{k}\Omega} = \frac{25 - (1{\cdot}2 + 0{\cdot}7)}{5\text{k}\Omega + 3{\cdot}9\text{k}\Omega}$$

$$= 23{\cdot}1/8{\cdot}9\text{k}\Omega = 2{\cdot}3\text{mA}$$

Which divides approximately equally between the base of VT_2 and via C_2 through R_{B5}.

Thus the transitory state demands a β in VT_2 of about (6·4/1·1)mA, i.e. 5·8. This is reasonable, as has been seen earlier.

Then with the 'turn on' of VT_1, followed by VT_5, the capacitor C_2 charges to its steady state value set by the collector of VT_3:

$$\text{Current in } R_{L3} = \frac{25\text{V} - 1{\cdot}9\text{V}}{10\text{k}\Omega + 3{\cdot}9\text{k}\Omega} = 23{\cdot}1/13{\cdot}9\text{k}\Omega = 1{\cdot}66\text{mA}$$

all of which flows out of VT_2 base.

The demanded steady state β of VT_2 = 6·4mA/1·66mA = 3·85 which is easily satisfied.

The collector of VT_3 rises to:

$$-(1{\cdot}9\text{V} + 1{\cdot}66\text{mA} \times 10\text{k}\Omega)$$
$$= -18{\cdot}5\text{V}.$$

The trigger transistor has to switch 6·4mA maximum. It has been seen that it is desirable to have a low $V_{ce(sat)}$ for VT_1. Hence as there is no pretence to speed it is reasonable to bottom it hard. If a β of 5 is assumed an I_B of 1·28mA is required. Hence an R_B of say:

$$\frac{25\text{V} - 1{\cdot}2\text{V}}{1{\cdot}28\text{mA}} = 18{\cdot}6\text{k}\Omega.$$

The next step is to choose a value of C_1 which is just sufficient to guarantee to turn off VT_1. One approach is to figure that the current out of VT_1 base should be diverted through C_1 for at least long enough to account for hole storage time plus fall time of its collector current. 20μsec is sufficient to account for this sum of delays.

$$\text{Hence } C_1 = \frac{20 \times 10^{-6}}{18 \times 10^3} \simeq 1 \times 10^{-9}\text{F}.$$

i.e. 1 000pF

The capacitor C_2 is the final value to decide.

The principle upon which the circuit operates requires that:

$$C_2 R_{B5} > C_1 R_{B1}$$

$$\text{Hence } C_2 > \frac{C_1 R_{B1}}{R_{B5}}$$

$$> \frac{10^{-9}\text{F} \times 18 \times 10^3}{10^4}$$

$$C_2 > 1{\cdot}8 \times 10^{-9}\text{F}$$

Say C_2 may be 10^{-8}F
$= 0{\cdot}01\mu$F.

The input to the trigger differentiating circuit should be a positive current pulse of duration in excess of 20μsec and amplitude greater than 1·3mA.

Clearing the ring and resetting zero has been dealt with earlier.

The lamps used for the read-out may be supplied either from a separate 6V line or from the 25V line through appropriate resistors.

Note (1). It is assumed that with these silicon devices and operating current levels, the effects of leakage currents may be neglected.

Note (2). This counter was constructed using random devices and worked up to 150kc/s with a 6V input pulse.

Conclusions

It has been shown that a silicon ring counter of the type described[1] can be constructed at a cost which makes it an economical proposition. The changes introduced by using a cheap silicon transistor in place of the very sophisticated device have been discussed in detail.

Other suitable pnp silicon alloy transistors are the OC430, OC440 and OC445.

The rapid development in planar transistors has made a significant impact on their price structure. The following npn devices appear to be suitable candidates for this type of circuit, 2N706, BSY51, BSY73, 2N2256, 2N2257 and ST-01.

Acknowledgment

The author wishes to thank the management of Brush Clevite Co. Ltd for permission to publish this article and to draw attention to the work of the authors of Ref. 1 and 3 and to acknowledge its value.

REFERENCES

1. RINGROSE, J. W. A Transistorized Ring Counter (Letter). *Electronic Engng.* *33*, 815 (1961).
2. WESTOBY, P. J. A New Type of Ring Counter. *Electronic Engng.* *31*, 375 (1959).
3. RINGROSE, J. W., BRADLEY, D. J. The SA496 Silicon Transistor in a Directly Coupled Ring Counter. Labstract 9 (Semiconductors Ltd, Swindon).

Solar Cells for Use in Heart Machine

A new safety device for use in Hurt's cardiac by-pass apparatus has been evolved by Standard Telephones & Cables Ltd. This apparatus is used in the Drew method for open cardiac surgery which involves cooling the patient's body to 12°C at which temperature the circulation may be arrested for periods up to one hour without damage to the brain. Circulation must, however, be maintained during the cooling and warm-up period. This is done by means of two pumps by-passing the right and left sides of the heart. It is vital that the blood volume of the patient be constant and that no air be injected.

The problem was to find a reliable and instant method of stopping the by-pass pumps in the event of their blood reservoirs falling below a safe level. Continued pumping, even for a few seconds in these circumstances would introduce fatal air into the patient.

A special solar-cell was devised to stop the pumps automatically. The design had to meet three essential requirements:
(a) The photo-cells which detect the fall in reservoir level had to be highly sensitive because the transparency of the plastic containers is not high; the material becomes yellowish after sterilization and also carries a blood film on the inside surface during operation.
(b) The cells had to give a high output, sufficient to operate relays directly so that amplifiers and other potential causes of failure could be avoided.
(c) The cells themselves had to be of the highest quality and reliability available.

High output solar cells (STC type PV20) were chosen because no additional lighting or optical focussing was found necessary beyond the 30W lamp used to illuminate the machine itself.

The pump is restarted automatically when the rising blood level in the reservoir again obscures the light to the photo-cell. A separate switch also permits over-riding manual control of the pump when required.

A Self-Tuning 5kW Linear Amplifier

By J. Wood*, M.I.R.E., Assoc.I.E.E.

This article describes an automatic tuning r.f. linear amplifier with an output of 5kW peak envelope power (p.e.p.). The amplifier is designed to be driven from a low power driver transmitter providing an output power of 500W p.e.p. over the frequency range from 2·5 to 25Mc/s. The amplifier incorporates a servo system controlled from discriminators which enables the amplifier to be automatically adjusted to the required frequency in a few seconds—the longest wavechange is 10sec with average channel changes of 5sec.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

LESS than two decades ago all high power transmitters were designed to be manually readjusted to a new frequency. The length of time taken to perform this operation varied from 10min to 1h depending on the complexity of the work involved in shutting down the transmitter and performing all the necessary circuit changes, including removing all the tank inductors and substituting replacements.

Later designs of high power transmitters employed variable tank inductors, usually driven from chains and gears manually controlled by the operator. This, and other improvements made a substantial contribution to the reduction of the off traffic period by reducing the wavechange times down to something of the order of 10min maximum. The introduction of motor driven tank inductors followed shortly afterwards, mainly with the object of decreasing the physical effort demanded from the manual operator.

Following on from this came the introduction of multichannel transmitters employing positioning mechanisms, although the use of such mechanisms was usually limited to power stages delivering less than 1kW for reasons of the torques demanded by large tuning capacitors and inductors. The number of channels provided rarely exceeded 12, smaller numbers between 4 and 8 being more common. With such systems wavechanges could be made in times of the order of 30sec. It must be borne in mind that these changes were manually operated in the first place to the desired settings corresponding to one particular frequency and one particular aerial in order to preset the mechanisms.

The logical development from the multi-turn mechanisms selector was the introduction of servo systems employing servo motors and positioning potentiometers. Such transmitters represented a very great improvement in auto-tuned transmitters, since the number of channels available was limited only by the number of positioning potentiometers or fixed resistors connected to a selector switch. The tremendous advantage of this type of multi-channel autotuning transmitter over that utilizing multi-turn mechanisms is that the frequencies of any channel not in use can be changed without interrupting traffic on the channel in use.

New Concept

The 5kW h.f. linear amplifier described here is a departure from the types of transmitters described in that it is entirely self-tuning in the strictest sense, being automatically tuned and controlled on the correct aerial loading by three discriminators which operate respectively on the frequency of the incoming drive signal, the phase difference between the driving r.f. voltage and the developed r.f. voltage in the 5kW π-network and finally the amplitude relationships of these voltages.

* *Racal Electronics Ltd.*

The chief advantages of this automatic tuning amplifier over the pre-positioned servo tuned transmitters is a faster and more accurate tuning up and which is independent of the aerial impedance within a s.w.r. or 2 to 1. The tank impedance of the 5kW output is automatically maintained at the current value for any subsequent changes of aerial loading.

The tuning and loading time is fast and under certain circumstances where the frequency change is say 100kc/s at 7Mc/s is effectively zero time. An average frequency change takes 3 to 4sec.

Basic R.F. Design

To achieve the high speed tuning facility it is of primary importance that the number of variable tuning elements should be as small as possible. Secondly, no switching of high power tank inductors or capacitors is permissible because this would introduce difficulties at the power involved as well as significantly increasing the channel changing time under certain conditions. It was therefore decided to employ wideband techniques wherever possible. Vacuum capacitors with a wide range from minimum to maximum capacitance, used in conjunction with a specially designed variable tank inductor suitable for high speed operation provide a convenient solution to the problem of the tank network design.

To meet the alternative output impedance of 50Ω unbalanced feed and 600Ω balanced feed it was decided to provide the 50Ω output impedance from the tank network and then to produce the 600Ω balanced output by means of a separate wideband transformer of 50 to 600Ω impedance conversion. The advantage of this arrangement is that the number of variable elements in the 5kW tank network can be reduced to two, though for other reasons it is preferable to use a π-network arrangement of three elements.

A further significant problem associated with self-tuning transmitters is the design of the neutralizing circuits which must give complete coverage without any necessity for switching into separate bands. This can be achieved by the use of low capacitance ceramic valves or by the use of conventional valves in a grounded grid configuration.

Choice of Transmitting Valve Type

The advantages of employing a grounded grid arrangement are well known[1-5]. The configuration is particularly adapted to automatic transmitters operating over the h.f. range since no neutralizing is necessary and the low input impedance is well suited to matching into a coaxial cable via an impedance conversion device. The input impedance remains sensibly constant throughout the operating cycle even when grid current is drawn. In fact, the only design requirement that was in any doubt was the power gain since, in general, grounded grid stages do not yield large power gain.

AUTOMATIC TUNING CONSIDERATIONS

Analysis of a three-element π-network shows that the magnitude of the dependent variable functions can be minimized by employing a high ratio of Z_1 to Z_2 terminating impedances (Fig. 1).

This in fact is the case if the output impedance is very low as in coaxial terminations. Therefore, so far as the initial automatic tuning or the coarse tuning phase is concerned it is theoretically possible to electrically gang together the three tuning elements via their servo systems and to drive the complete automatic tuning from a pilot servo loop driven in turn from a coarse tuning discriminator. It will be necessary to arrange the ganging and tracking of the 5kW tank network to the geometrical mean of the range of aerial impedances likely to be presented. In this way it is possible to arrange that at any operating frequency and for any complex aerial impedance within the limits of the specification the coarse tuning will always be sufficiently accurate to realize a reasonable amount of r.f. power output, though it will not be correctly tuned. The final and perhaps the most difficult task of ensuring that the 5kW stage is automatically corrected for the mistuning by measuring the phase-angle of the tank network is now also theoretically possible, as is also the problem of ensuring that C_2, the loading capacitor is automatically tuned to suit the aerial impedance. In general it is C_1 which determines the phase-angle of the load impedance presented to the valve and C_2 which determines the magnitude of the load impedance. It is possible therefore to lock the tuning inductor in the coarse tuned position and to carry out the final tuning on the capacitors C_1 and C_2. An adjustment to C_1 will call for a minor readjustment to C_2 and vice versa by their respective servo systems but eventually and in a short space of time (about 1sec) both servo systems will have reached a null response and the amplifier will be correctly tuned.

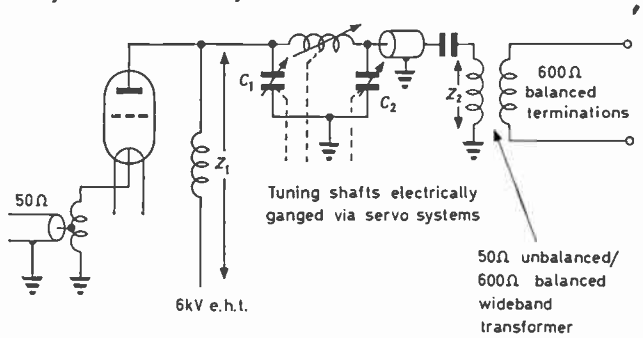


Fig. 1. π**-network with high ratio terminating impedances**

CAPACITORS AND INDUCTORS

Practical considerations as to the suitability of components indicated that the π-network proposed for the 5kW tank circuit should use vacuum variable capacitors for the terminating impedances and a continuously variable tank inductor for the series arm. All three elements should have a straight line law for turns against capacitance/inductance.

Vacuum variable capacitors are the ideal solution to the requirement chiefly because of their small bulk coupled with their high electrical ratings and wide range of minimum/maximum capacitance.

The tank inductor should be capable of high speed tuning to give rapid selection of any portion of any turn. To maintain a reasonable tank circuit efficiency the coupling between the used and unused turns should not be such as to bring the natural unloaded Q value to less than 150. It may also be necessary to employ some method of law correction between the tank inductor and the gear box to yield the correct law since the natural tendency for any inductance is to follow a square law (of inductance versus turns). However, the large transmitting inductances can usually be constructed to follow an approximately linear law.

Detailed Design

AUTOMATIC TUNING (see Fig. 2)

The automatic tuning of the amplifier is achieved through three separate high speed servo loops employing high torque d.c. motors and servo amplifiers with positioning precision potentiometers. These servo motors drive the π-network elements via suitable reduction gears equipped with automatic slipping clutches. The three servo loops can, for purposes of manual tuning be electronically controlled from a single control on the front panel of the equipment. In this condition the three tuning elements are ganged together electrically through the servo positioning systems. Two further controls engraved 'resonance' and 'loading' provide the necessary facility for manually steering the capacitors to yield the correct loading and resonance. However, the principle mode of operation is fully automatic where the positioning servos are self-tuned by means of three built-in discriminators which operate respectively on frequency, phase and amplitude.

The sequence of operation is as follows. The first phase of automatic tuning is the coarse tuning phase which commences on the receipt of a low level pilot carrier. This signal is fed into a discriminator connected to a servo loop which produces a voltage proportional to wavelength, this operation takes less than 1sec. This voltage is then automatically applied to the three positioning servos which carry out the coarse tuning. When this is completed, which takes an average time of 4sec, a sensing relay automatically initiates phase three, in this a phase discriminator operates from the peak r.f. drive voltage applied to the cathode and the peak r.f. voltage developed at the anode of the transmitter valve. The output from this discriminator is used to operate the servo loop which drives the resonance capacitor of the tank π-network to yield exact resonance.

Simultaneously an amplitude discriminator operating from these same two voltages is used to operate the main positioning servo loop controlling the loading capacitor of the π-network to produce the exact degree of aerial loading. Fine tuning and aerial loading occurs simultaneously and sequentially since a change in the setting of the element will usually call for a corresponding adjustment to the other. The time taken for these operations is of the order of 1sec. It should be noted that the automatic tuning circuits are always energized to enable the amplifier to be readjusted slightly from time to time to compensate for the effect of temperature changes upon the impedance characteristics of the aerial.

COARSE TUNE UNIT

The coarse tune unit consists of a bridge type RC discriminator, a d.c. servo amplifier and a permanent magnet d.c. motor. A small r.f. input to the discriminator produces a d.c. error voltage which, after amplification by the servo amplifier is used to drive the motor.

When C_1 (Fig. 3) is rotated so that its capacitive reactance is $-j680\Omega$ the bridge is balanced in magnitude. The r.f. voltages across R_2 and C_1 are applied to detector circuits so that the voltages are converted to d.c. before being

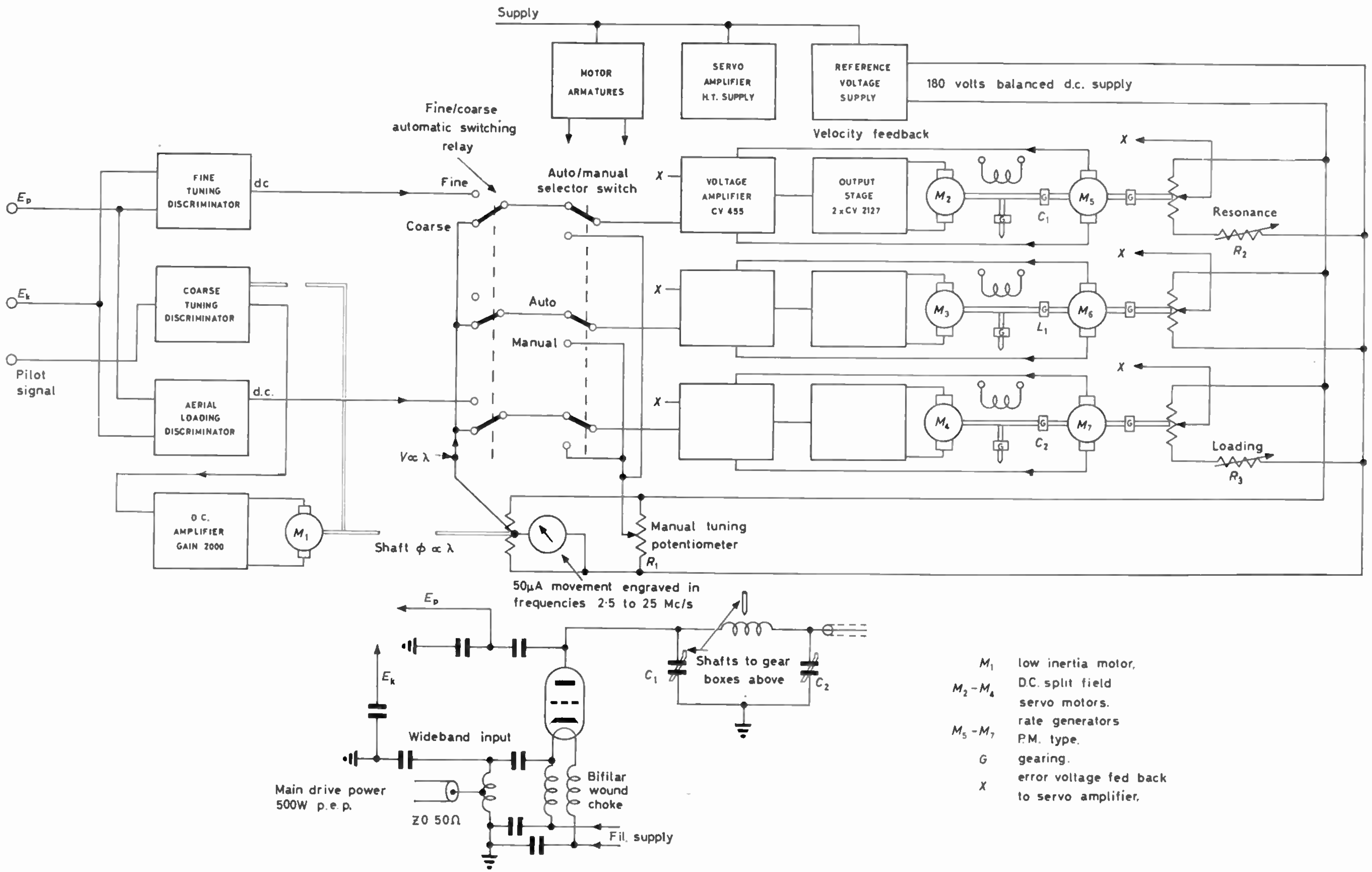


Fig. 2. Block schematic showing essential parts of auto tuning system

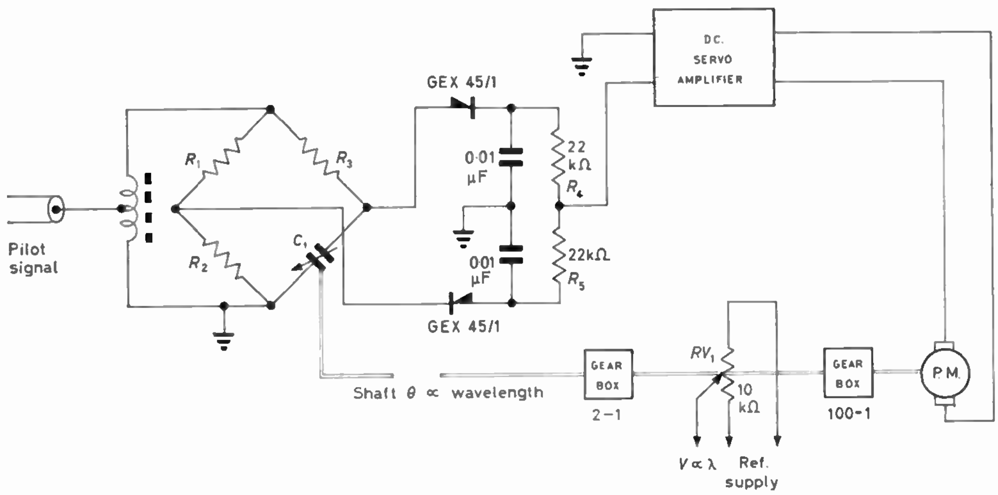


Fig. 3. Coarse tune unit

vectors derived from E_p. These three vectors are then rectified so as to produce a d.c. voltage which gives the discriminator response as shown. By careful design the effects of magnitude changes in the E_k components resulting from the use of an RC phase-shift network are reduced to an acceptable amount.

The fall off in signal amplitude of the discriminator output (Fig. 4(b)) as the phase-angle error increases from about 45° is fundamental due to the fact that the E_p is of a diminishing nature either side of resonance. The point from which the servo systems will pull in is determined by the amount of error signal available to overcome the static friction. Due to the high gain of the servo system and the fact that the servo motors develop an ample margin of torque, the range over which the fine tuning discriminator will operate the tuning capacitor C_1 is very wide at the low frequencies. The range over which it will operate is somewhat less for the frequencies above 15Mc/s due to the higher tank circuit Q. Even so, at 24Mc/s the discriminator will pull in from 3Mc/s either side of the resonance.

compared. The resistors R_4 and R_5 add two voltages but as the polarity of these voltages is opposite, the difference voltage will appear between their junction and earth.

This difference voltage is amplified by a d.c. amplifier and applied to a cathode-follower stage driving a d.c. permanent magnet motor which is coupled to C_1 and rotates so as to reduce the applied error signal.

The angular rotation of C_1 is thus proportional to wavelength which is the correct law required for positioning the 5kW π-network. Coupled to C_1 via a 2:1 step up gear is a precision potentiometer RV_1. The angular rotation of C_1 is thus converted to electrical information to operate the servo system.

Fine Tuning

This is shown in Figs 4(a) and (b) which give the electrical circuits for the fine tuning and aerial loading discriminators together with their associated responses.

The minimum lower limit to the accuracy demanded of the coarse tuning discriminator is dependent on the ability of the fine tuning discriminator to drive the servo tuning to the correct point from a position far down on the resonance of the tank circuit. The design of the fine position tuning discriminator is therefore of paramount importance, since a good design of fine tuning discriminator will ease the design problems of the coarse tuning systems.

An investigation was carried out on the problem of developing a fine tuning discriminator which would operate satisfactorily over the entire range of frequencies from 2·5 to 25Mc/s. As a result it was concluded that the only practical solution was to utilize the phase relationship existing between the r.f. driving voltage and the r.f. output voltage at the anode of the valve and to apply these voltages to a phase discriminator. Obviously no circuit can be devised to operate over such a frequency range without introducing some unwanted phase shifts, and the major problem was therefore to eliminate such effects while keeping the circuit simple.

Referring to Fig. 4(a) voltage E_p is fed into the grid of a phase-splitter via the screened capacitive divider which reduces the voltage from 6kV peak to about 12V. The valve acts as a phase-splitter producing two components with almost 180° phase difference. The voltage E_k from the cathode of the valve (being the driving voltage) is similarly reduced in amplitude to an almost equal voltage and its phase is simultaneously shifted so that it has a 90° phase difference to either of the two 180° opposed

Aerial Loading

In any transmitter tank network it is essential to maintain a close tolerance on the impedance offered to the valve. This is even more important in linear amplifier design. This means that to avoid the necessity for constant retuning, the transmitter must be equipped with an automatic self-loading servo which will enable the tank circuit impedance to be maintained at the optimum value for all frequencies and for all aerial impedances likely to be obtained in practise.

A loading discriminator can be obtained from the sampling of various combinations of r.f. circuit parameters existing in the grounded grid amplifier. Alternatively if a constant level tuning signal is available any one of these circuit parameters may be compared with a fixed d.c. level to yield a loading discriminator action. For preference, d.c. voltages rather than r.f. voltages are used so as to avoid the introduction of further errors associated with rectifying r.f. voltages over a wide frequency range.

Practical experiments have been made on all of these methods. Without doubt the simplest form of loading servo is obtained by comparing the valve cathode current with any one of the following parameters.

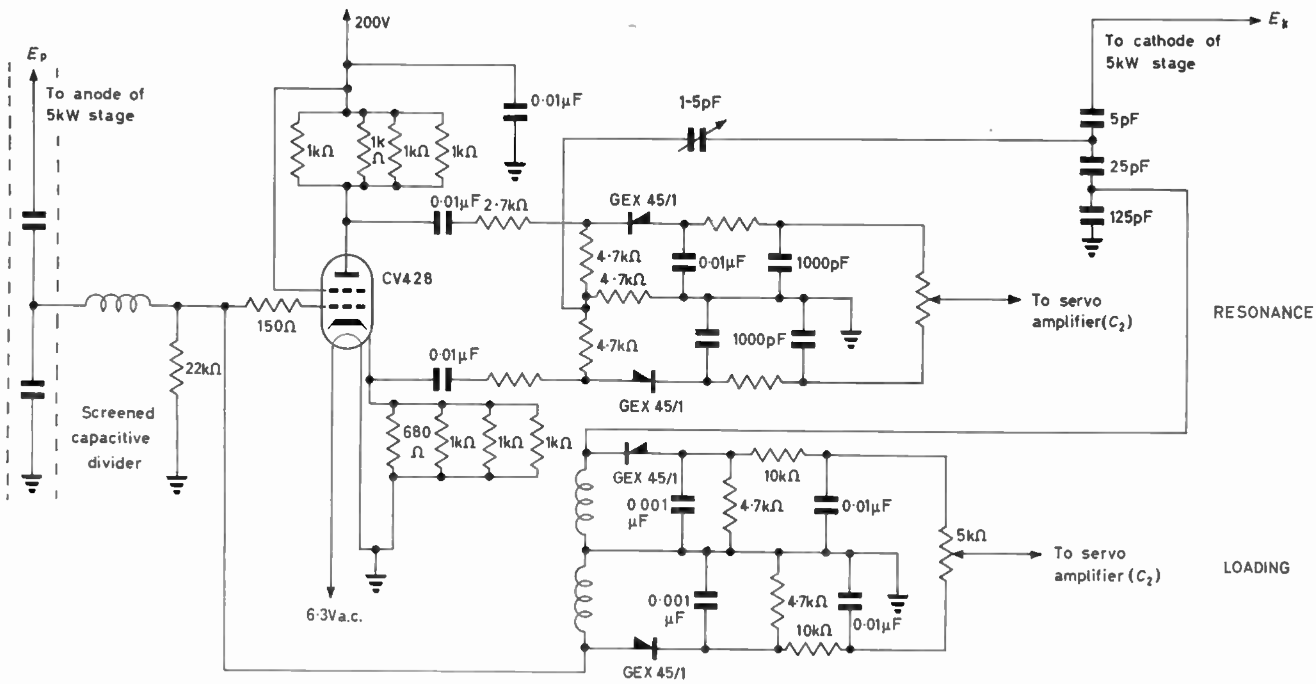
(1) A d.c. reference level.

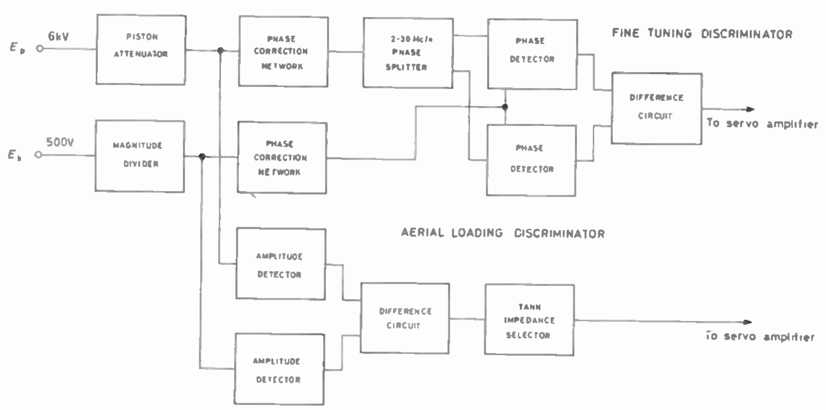
(2) E_k/E_p

(3) $I_{g(d.c.)}$

Of these the d.c. level is the simplest method. It should, however, be noted that in any of these methods with the exception of E_k/E_p it is essential that the driving voltage is held to a steady and correct level otherwise incorrect loading will result.

With the successful development of a screened capacitive divider for the purposes of reducing the r.f. anode voltage





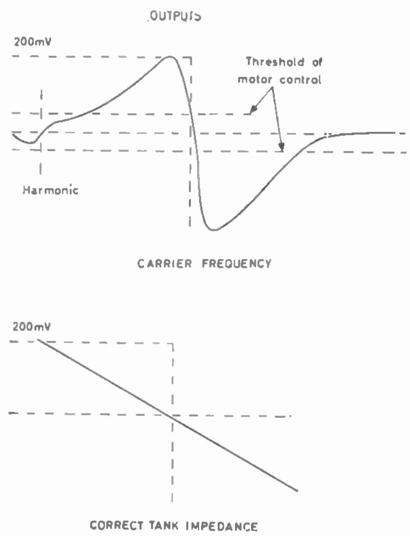


Fig. 4(a). Fine tuning discriminators
(b). Fine tune and aerial loading servo discriminators

for the fine tuning discriminator, the way was opened to carry out some experiments utilizing the magnitude relation ship between E_p and E_k for the purpose of producing the loading discriminator. These experiments showed a marked superiority of this type of loading discriminator over the alternative methods. Its principal advantage is its ability to maintain a constant tank impedance regardless of the type of signal (e.g. s.s.b., c.w., d.s.b.) or the level of the signal up to the limiting power level where E_p commences to flatten in waveshape due to overdrive.

Included in the design are two preset potentiometers for initial adjustment for correct conditions. Thereafter no further adjustment should be necessary unless it is desired to change the tank loading to the valve. This may be desired if the transmitter is to operate permanently on c.w. since a greater valve efficiency and power output can thereby be achieved.

Tuning Accuracy

The overall accuracy of an auto-tuning transmitter may be impaired by insufficient mechanical/electrical resolution. It is therefore essential that the overall resolution of any one of the servo loops is more than sufficient to realize the necessary tuning accuracy at the highest frequency when controlled from the manual positioning potentiometers.

In the final analysis when the transmitter is operated under self-tuning conditions the overall accuracy is dependent on the design of the fine tuning and aerial loading discriminators. The criterion of design should be that the tuning accuracy is never worse than that achieved by normal manual tuning methods, and for preference should yield greater accuracy.

Physical Aspect of R.F. Design

The r.f. design is intended to provide the maximum screening between the input and output circuits. The grid deck contains the grid components and the input circuit arrangement, the grid deck itself being earthed to r.f. through a built-in distributed capacitance plate on to the grid itself, thus minimizing grid inductance.

The tank inductor is positioned symmetrically in the anode compartment to give maximum efficiency. The tuning capacitor connects directly on to the valve jacket, thus eliminating any series inductance in this arm. For the same reason the loading capacitor connects directly on to the base of the tank inductor. The screened capacitive divider connects directly to the anode valve jacket. The anode deck situated some 3in below the grid deck is of silicon bonded laminated glass material and its purpose is to seal off this compartment so that all the air is forced down through the fins of the valve anode.

Protective spark gaps have been fitted across the tank inductor and the loading capacitor. These spark gaps conduct if abnormally high peak voltages are developed during the automatic tuning sequence, thereby protecting the components from possible damage. It may be noted that the spark gaps are located in the path of the air blast which assists in dispersing the ionized gases and eliminates the hazards due to this phenomenon.

The tank inductor was developed in conjunction with the amplifier design to meet the usual requirements, i.e. high value of natural efficiency (or unloaded Q) freedom from unwanted electrical spurious resonances, low mechanical torque and low inertia, the two latter requirements being necessitated due to the high speed tuning requirement. All these requirements were finally achieved by a somewhat unusual arrangement. A stationary leadscrew carries an aluminium disk which is rotated on the leadscrew thread by means of a lower disk located with 3 vertical rods. The aluminium disk, via a ball contact, selects the desired turn and position on the turn and also serves to form a screen, thus reducing the magnetic coupling between the used and unused portions of the tank inductor, the unused portion being always shorted out. The total travel time from top to bottom of the tank inductor is 5sec and any part of the tank inductor can be selected with a positioning error of less than ⅛in over the entire length, which is approximately 30ft, thus yielding a resolution of about 0·1 per cent.

Future Trends

The TA.84 5kW h.f. linear amplifier was developed in 1956 and went into service in 1958 being at that time one of the first of the new generation of h.f. linear amplifiers designed on self-tuning principles and incorporating automatic aerial loading servo's.

It is considered that the automatic tuning techniques employed in the TA.84 are sufficiently advanced as not to warrant any significant design changes in the foreseeable future. Some significant improvements in valve design have taken place since the TA.84 was developed, notably the advent of the Eimac range of high power ceramic tetrodes intended specifically for s.s.b. operation. This type of valve will inevitably replace the grounded grid valve in future designs of h.f. transmitters.

Silicon high voltage rectifiers offer some possibility of a space reduction and also in heat dissipation, the existing xenon rectifiers in the e.h.t. supply have a total filament consumption of 180W. Similarly it is highly likely that the bulk of the servo system including the power supplies will be changed to using solid state components.

On the question of power output there is a tendency for planning authorities to consider 10kW as being an optimum compromise between the desired power for a long range circuit against the steeply rising costs of higher powered transmitters. It is considered therefore that the 5kW power output may have to be up rated to 10kW.

Finally with regard to the frequency range and the output power. There are indications that more and more users will follow the military trend in requiring transmitters which are capable of covering from the m.f. band of 1·5Mc/s up to 30Mc/s with continuous tuning. This is considered to be a practical requirement.

Acknowledgments

The author is indebted to the Board of Directors, Racal Electronics Ltd, for permission to publish this article.

REFERENCES

1. STRONG, C. E. The Inverted Amplifier. *Electronics.* p. 87 (July 1940).
2. BEVAN, P. A. T. High Power Television Transmitters. Part II. *Electronic Engng. 19,* 181 (1947).
3. JONES, M. C. Radio Frequency Voltage Amplifier. *Proc. Inst. Radio Engrs.* (July 1944).
4. MULLER, J. J. Cathode Excited Linear Amplifier. *Electrical Commun.*
5. STRONG, C. E. The Inverted Amplifier. *Electrical commun. 19n,* 32 (1941).

An Image Intensifier for Nuclear Research

Mullard Research Laboratories in conjunction with the Atomic Energy Research Establishment have developed a high-sensitivity image intensifier tube for use in experiments involving sub-atomic particles. The Authority has placed a contract with Mullard Ltd to supply six of these tubes, which are being made at the company's Mitcham factory.

In many experiments involving sub-atomic particles it is necessary to photograph one event of special interest among a million or so others which occur within a particular period. For example, it may be necessary to record the light track of a single particle involved in a collision process, or to determine the energy of the particle by means of the Cerenkov radiation it causes.

The amount of light produced by a single particle is extremely small. In the example quoted, the light produced would barely affect one grain of the emulsion on a photographic plate. Therefore, if useful photographs are to be obtained, the light must be intensified without its pattern being distorted. This can be done by means of image intensifier tubes incorporating photo-emitters which have a much higher quantum efficiency than photographic emulsions.

The Mullard two-stage image intensifier being supplied to the A.E.R.E. is such a device. It will be used as a low noise pre-amplifier in a system that produces useful photographs from extremely low-level light sources. The tube intensifies the light from a single particle into an image that provides a suitable input for a high-gain image amplifier.

Another difficulty which arises when investigating sub-atomic particles is the short duration of the light produced in a scintillation, or by Cerenkov radiation. The ancillary circuit needed first to recognize an event as one of interest, and make the decision to photograph it, takes longer to operate than the duration of the event.

The event must therefore be recorded and stored. The image intensifier does this by means of screens that will retain an image for a fraction of microsecond or several milliseconds.

The image intensifier contains two similar stages. In each stage the electrons emitted from the photo-cathode are focused electrostatically on to a fluorescent screen. Hence, the image on the screen is a reproduction of the pattern of the active region on the photo-cathode and its screen, making a total demagnification of 7·29:1 between the first photo-cathode (input) and the second screen (output). The first photo-cathode has a diameter of more than 150mm, but the maximum diameter of the final image is 20mm.

The screen of the first stage and the photo-cathode of the second stage are on the opposite sides of a mica sheet only 15 microns thick. This arrangement ensures highly efficient coupling between the two stages with little loss in image resolution.

Unlike the first stage of the image convertor, the second stage has an electrode fitted between the photo-cathode and the focus anode. When this electrode has a potential of about 200V negative, no electron from the photo-cathode reaches the output screen. It is used as a shutter to select the required event and to prevent noise generated by the system reaching the output screen. The screen in the image convertor may contain one of several phosphors, depending on how long an afterglow is needed. Screens with an afterglow ranging from a fraction of a microsecond to several milliseconds are available.

A Simple Phase Modulator

By C. T. Kohn*

An analysis is given of a simple phase modulator consisting of three reactances and a valve. Optimum working conditions are derived and the effects of a non-linear valve characteristic discussed. In a practical design, 10 per cent frequency distortion was obtained with 48° of deviation.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

IN equipment which is to be produced in quantity at a competitive price, preference is given to circuits which give reliable operation at low manufacturing cost, even if some sacrifice of quality of performance has to be made. It is an advantage if tuned circuits or critical operating conditions can be avoided.

In phase modulated transmitters, the need for a simple modulator has led away from the classical circuits[1] based on the addition of two voltages shifted by 90°, to the development of simpler circuits. One of these, using a valve as a variable impedance in series with a reactance which has for some time been in favour with designers, will now be discussed.

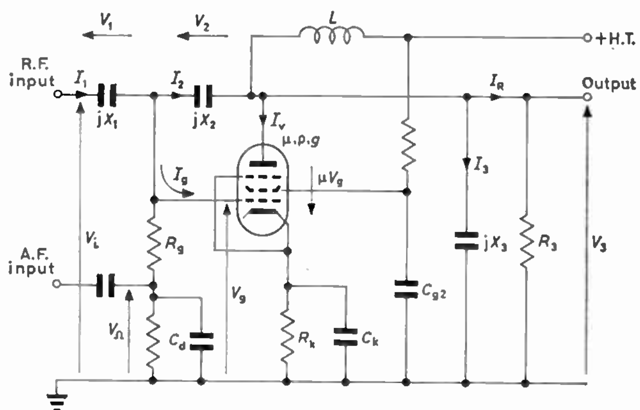


Fig. 1. Circuit diagram of the modulator

The Circuit

The circuit diagram of the modulator to be examined is shown in Fig. 1. The three reactances X_1, X_2 and X_3 act as a voltage divider, in which the phase of the output voltage across X_3 is shifted by the valve current. This phase shift should be as large as possible, and be proportional to the applied a.f. grid voltage.

In the circuit diagram of Fig. 1, it is assumed that the three reactances X_1, X_2 and X_3 are loss free, that the valve does not take any grid current, neither capacitive nor resistive, and that none of the auxiliary circuit elements affects the performance. The load is represented by the resistance R_3. If an r.f. voltage V_i is applied to the circuit, so that the voltage V_g appearing at the grid is small comared with the grid cut-off voltage, the following relations will apply:—

$$\begin{aligned} V_i &= V_1 + V_2 + V_3 \\ V_g &= V_i - V_1 \\ \mu V_g &= I_v\rho - V_3 \\ I_1 &= I_3 + I_v + I_R \end{aligned} \qquad \text{.............. (1)}$$

The solution of equation (1) gives the phase-angle ϕ between V_3 and V_i

$$\phi = \text{arc tan}\,\frac{gX_2(1+B)+A}{1-B(gX_2)^2-AgX_2} \qquad \text{.............. (2)}$$

the gain of the circuit

$$|V_3/V_i| = \frac{X_3}{X_1+X_2+X_3}\sqrt{\left[\frac{1+(gX_2)^2}{1+(BgX_2+A)^2}\right]} \qquad \text{.............. (3)}$$

and its input impedance

$$Z_1 = (V_1/I_1) = \mathrm{j}(X_1+X_2+X_3)\,\frac{1+\mathrm{j}BgX_2+\mathrm{j}A}{1+\mathrm{j}gX_3(1+(1/\mu))+\mathrm{j}(X_3/R_3)} \qquad \text{............. (4)}$$

where:

$$A = (X_2/R_3)\,.\,\frac{X_1+X_2}{X_1+X_2+X_3} \qquad \text{........................ (5)}$$

$$B = (X_3/X_2)\,.\,\frac{X_1}{X_1+X_2+X_3}\left(1+\frac{X_1+X_2}{\mu X_1}\right) \qquad \text{.............. (6)}$$

For the purpose of modulation, a suitable grid bias V_{go} is set up, resulting in a mutual conductance of the valve g_o, and an audio frequency voltage $V_\omega \sin \omega t$ where ω is the angular frequency of the modulating voltage, is applied to the grid. If a valve is used whose mutual conductance g is a linear function of the grid bias voltage the valve conductance becomes

$$g = g_o(1+(V_\omega/V_{go})\sin \omega t)$$

where V_{go} is the fixed grid bias measured from the cut-off voltage.

Operation Without Resistive Load ($R_3 = \infty$, $A = 0$)

Deviation and Distortion

The phase shift ϕ is a function of the valve transconductance g and a single circuit parameter B. At small values of gX_2, ϕ increases linearly, but flattens off beyond $gX_2 = 1$ and approaches ϕ_{max} asymptotically, for $gX_2 \rightarrow \infty$. If $B > 0$, ϕ reaches a maximum of 180°, if $B = 0$ $\phi_{max} = 90°$, and even less if $B < 0$. Thus only the case $B > 0$ need be considered.

The rate of increase of ϕ as a function of gX_2 increases with B, and for good sensitivity a high value of B would be adopted. However, modulation distortion is the decisive factor in the choice of B, and this was evaluated for various values of B. It was assumed that operation extends over the most linear part of the $\phi = f(gX_2)$ curves, from $g = 0$ to $2g_o$, the $g = f(V_g)$ characteristic being linear. As higher harmonics are small, only the second harmonic content ϕ_2/ϕ_1 is plotted as a function of the fundamental output in Fig. 2.

Fig. 2 shows that for the same deviation angle the distortion is smallest if $B = 1$, and reaches 20 per cent at $\phi_1 = 60°$. The distortions for $B = 2$ and 0·5 are equal and about 10 per cent higher than for $B = 1$. The inferiority of the case $B = 0$ is obvious.

* *Norbury Instruments Ltd.*

Fig. 3 gives the values of g_0X_2 necessary for obtaining a specified fundamental deviation ϕ_1 for various values of B, as a function of this deviation.

Incidental Amplitude Modulation

Gain variations within one cycle of the applied a.f. voltage, give rise to amplitude modulation incidental with the wanted phase modulation. From equation (3) (with $A = 0$), the gain remains independent of gX_2 if $B = 1$, so that there is no amplitude modulation at all. If $B < 1$, the gain increases on modulation; for $B > 1$, it decreases. If, for example, a phase modulation of $\pm 25°$ is obtained with $B = 0$, the modulation index is 0·41. This is very high compared with 0·085 obtained with $B = 0{\cdot}5$ or 2, for the same deviation. For a deviation of $\pm 50°$, $B = 0{\cdot}5$ gives 24 per cent and $B = 2$ 20·5 per cent of amplitude modulation. Thus, in the range important in practice $0{\cdot}5 < B < 2$ the amount of a.m. is not excessive and can be removed in the usual manner, by limiting in the following stages.

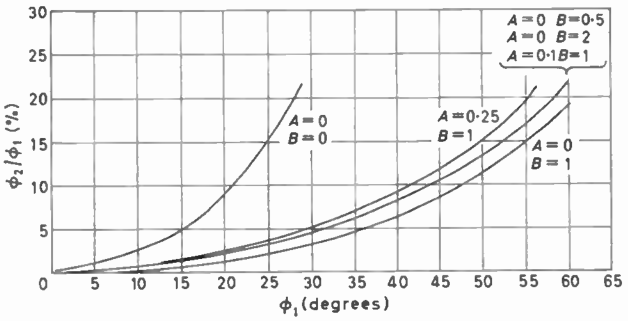


Fig. 2. Second harmonic content ϕ_2/ϕ_1, as a function of the fundamental component ϕ_1, for various values of B and A

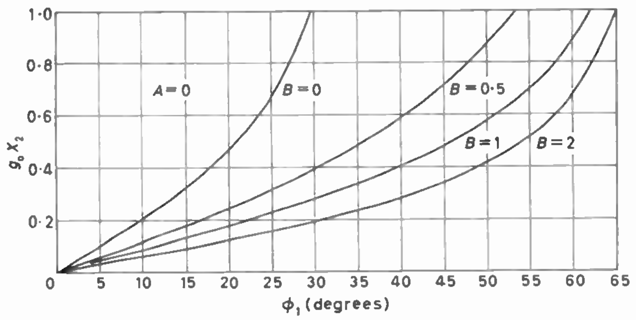


Fig. 3. Values of g_0x_2 necessary for obtaining a required deviation ϕ_1, as a function of the fundamental deviation component ϕ_1, for various values of B

Input Impedance

It is desirable that the input impedance remains constant over the modulation cycle, because this helps to avoid additional amplitude modulation and reaction on the preceding oscillator.

From equation (4), the input impedance without valve current is $j(X_1+X_2+X_3)$ and is modified on modulation by the terms in g of equation (4). This effect can be reduced considerably if $B = 1$ and $X_2 = X_3$ or $X_1 \gg X_2 + X_3$.

Effect of a Resistive Load (R_3)

In this case there is some deterioration of performance because there is some phase shift already at $gX_2 = 0$, which reduces the overall available phase swing. At $A = 0{\cdot}1$ which corresponds to Q-factors R_3/X_3 as low as 10, the distortion is approximately the same as for $B = 2$, $A = 0$ (Fig. 2). Thus, the circuit is capable of driving quite appreciable loads without unduly large deterioration.

The amplitude modulation and the input impedance are very little affected by a resistive load.

Operation with a Tuned Circuit

An examination of circuits used in practical applications shows that the modulator is used, as a rule, with a tuned circuit instead of the reactance X_3. If this circuit is tuned to resonance, the following equations apply:

$$\phi = \arctan \frac{gX_2(1+B)+A}{1 - B(gX_2)^2 - AgX_2} \quad \ldots\ldots\ldots\ldots (7)$$

$$|V_3/V_1| = \sqrt{\left[\frac{1+(gX_2)^2}{1+(BgX_2+A)^2}\right]} \quad \ldots\ldots\ldots\ldots (8)$$

$$Z_i = R_3\,\frac{1+jBgX_2+jA}{1+gR_3(1+(1/\mu))} \quad \ldots\ldots\ldots\ldots (9)$$

$$A = \frac{X_1+X_2}{R_3} \quad \ldots\ldots\ldots\ldots (10)$$

$$B = (X_1/X_2)\left(1+\frac{X_1+X_2}{\mu X_2}\right) \quad \ldots\ldots\ldots\ldots (11)$$

The phase characteristic is the same as in the preceding cases, and optimum performance is again obtained with $B = 1$. However, the proportioning of the elements is quite different, as now $X_1 \simeq X_2$, against $X_1 \gg X_2$ derived previously.

The gain is much higher and approaches unity.

However, if the r.f. voltage applied to the grid is the same as in the non-resonant circuit, the output is only 2·2 times higher. The incidental amplitude modulation is the same as in the previous case.

The input impedance is approximately R_3 for $g = 0$, but at peak deviation it may drop by one order of magnitude below this value. This is a very undesirable feature of the circuit.

This mode of operation, with the additional simplification of $R_3 = \infty$ has been described elsewhere[2,3].

Effect of the Modulator Grid Capacitance

If the modulator is used at a high radio frequency, say 10Mc/s, the capacitances corresponding to the reactances X_1, X_2, and X_3, especially X_1, become very small and comparable with the hitherto neglected input capacitance C_4 of the modulator valve. If this is taken into account, the phase shift becomes:

$$\phi = \arctan \frac{gX_2[1+(BD/(1+D))+B]+A}{1+(BD/(1+D)) - B(gX_2)^2 - AgX_2} \quad \ldots\ldots\ldots\ldots (12)$$

where:

$$A = (X_3/R_3)\,.\,\frac{X_1+X_2}{X_1+X_2+X_3} \quad \ldots\ldots\ldots\ldots (13)$$

$$B = (X_3/X_2)\,.\,\frac{X_1}{X_1+X_2+X_3}\left[1+\frac{(1+D)(X_1+X_2)}{\mu X_2}\right] \quad \ldots\ldots\ldots\ldots (14)$$

$$D = (X_2/X_1)\,.\,\frac{X_1}{X_1+X_2} \quad \ldots\ldots\ldots\ldots (15)$$

If capacitances are used for the reactances, D becomes equal to $C_4/(C_1+C_2)$.

If A is assumed to be zero, working conditions identical with the previous case of $B = 1$ are obtained, if:

$$B = 1 + D = 1 + \frac{C_4}{C_1 + C_2} \qquad \text{.............. (16)}$$

General Discussion of the Results

The phase distortion in the circuit of Fig. 1 depends on the parameter B, optimum performance being obtained for $B = 1$, or slightly higher (equation (16)), but values between 0·5 and 2 are satisfactory. The condition $B = 1$ is also the one preferred with regard to incidental amplitude modulation which at this value disappears altogether. For the input impedance to remain unchanged over the modulation cycle, B must be 1 and at the same time

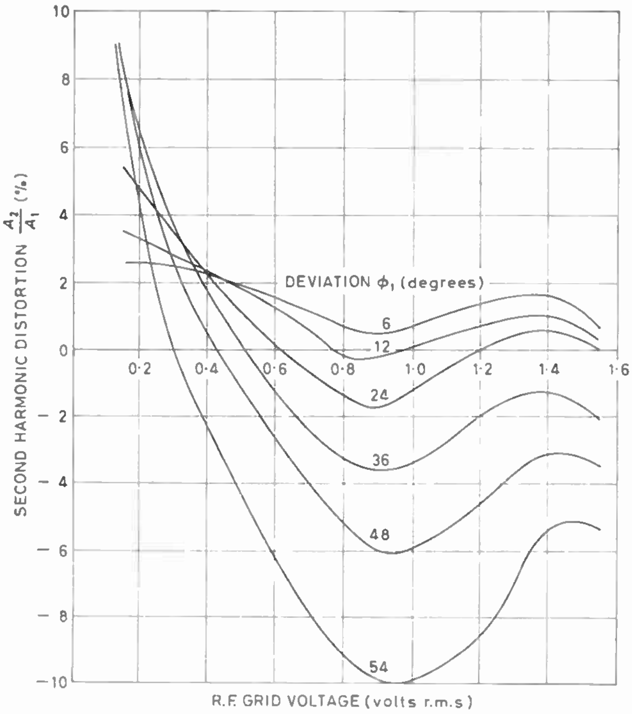


Fig. 4. Experimentally determined second harmonic distortion A_2/A_1, obtained with circuit Fig. 1 and an optimized g characteristic for various phase deviations, as a function of the r.f. grid voltage

$X_2 = X_3$. In practice this amounts to making X_2 and X_3 approximately equal and much smaller than X_1, and to using a high μ valve. Thus $B = 1$ gives the best performance with regard to all important performance requirements.

The reactance X_1 which enters into B, is not merely a coupling element to the generator, but plays an essential part in establishing favourable working conditions. If X_1 is treated as a coupling element and thus made very small ($B \rightarrow 0$), the performance becomes very poor.

A resistive load R_3 across X_3 produces an increase of distortion, but this is insignificant if the Q factor R_3/X_3 is greater than 10, a condition easy to meet.

Provision of an inductance for tuning out X_3 is unwarranted. Operation with X_3 tuned out does not give an improved modulator; on the contrary, it results in a circuit with very variable input impedance, the cost is higher and the circuit is frequency dependent. For these reasons it is inferior to the aperiodic circuit.

The reactances X_1, X_2 and X_3 may be capacitances or inductances, but as inductances can give rise to instability, preference will usually be given to capacitances.

Non-Linear *g*- Characteristic

For the performance assessment of the modulator a linear g/V_g characteristic was assumed. In practice, a valve will be used whose mutual conductance characteristic measured at r.f. as a function of the a.f. signal voltage, is concave upwards, so that some compensation of the modulation non-linearity is obtained.

In order to reproduce this characteristic in practice, a valve is chosen whose *g*- characteristic has rather a greater curvature than required, and negative feedback is used for its reduction so as to obtain the optimum shape. A mathematical treatment of shaping the *g*- characteristic by means of a cathode resistor is given by Boelens[4], but in practice this process is quicker carried out experimentally. All the derived equations remain valid for a non-linear *g*- characteristic.

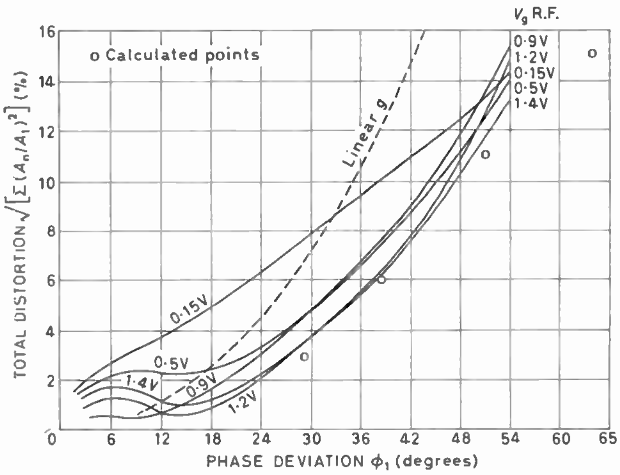


Fig. 5. Total harmonic distortion $\sqrt{[\Sigma(A_n/A_1)^2]}$ obtained with circuit Fig. 1 and the optimized g characteristic for various r.f. grid voltages, as a function of the phase deviation

Experimental Results

A circuit according to Fig. 1 was set up, for operation at 3·1Mc/s. All important components were adjusted experimentally to give the lowest possible modulation distortion at a phase deviation of 40°. With an EF95 and an h.t. voltage of 175V the optimum values were:

$C_1 = 5\text{pF}$, $C_2 = 47\text{pF}$, $C_3 = 33\text{pF}$, $R_k = 4{\cdot}7\text{k}\Omega$,
$R_{g2} = 150\text{k}\Omega$, $C_k = C_{g2} = 1\,000\text{pF}$.

The value of B taking into account circuit capacitances, chokes etc. was 1·15 which is in good agreement with 1·10 obtained from equation (16). No amplitude modulation could be detected.

Harmonic distortion was measured for various r.f. voltages, V_g applied to the grid, and for various phase deviations, after demodulation in a frequency discriminator. The second harmonic content A_n/A_1 is shown in Fig. 4. In comparing Fig. 4 with Fig. 2, it must be remembered that $A_2/A_1 = 2\,\phi_2/\phi_1$.

The shaping of the working characteristic by means of R_k was done at a r.f. grid voltage of 0·55V and a deviation of about 40° where the distortion is actually small. At other points, it depends markedly on the r.f. grid voltage, which thus becomes another significant parameter in setting up the optimum working conditions. In fact, another favourable working point is obtained with $V_{g(rf)} =$ 1·2 to 1·4V.

In view of the improvement obtained by a finite r.f. signal, the original restriction of an r.f. voltage small compared with the modulating signal, can be disregarded as non-essential in practice.

Third harmonic distortion is virtually independent of the r.f. grid voltage. It was 1 per cent for $\phi_1 = 12°$, 2·5 per cent for 24°, 6 per cent for 36°, 10 per cent for 48° and 12·5 per cent for 54°. The fourth harmonic was negligible.

The overall distortion $\sqrt{[\Sigma(A_n/A_1)^2]}$ is shown in Fig. 5 for some r.f. voltages, as a function of phase deviation. In addition, some points of calculated distortion obtained from the measured *g*- characteristic, are shown. They deviate significantly from the measured curves only at very high deviations. For comparison, the calculated distortion which would be obtained with a linear *g*- characteristic, is also shown.

Conclusion

In the design of the phase modulator under consideration, the principal condition is $B = 1 + D$. This has been derived for a linear *g*- characteristic, and confirmed experimentally for *g* non-linear. This condition leads to two practical working arrangements of which one ($X_1 = X_2$, $X_3 = \infty$) has been described previously. The second solution: $X_2 \simeq X_3$, $X_1 \gg X_2 + X_3$ is much to be preferred because it does not require any tuning element for X_3 and, in addition, has a substantially constant input impedance. Both solutions are distinguished by being free of amplitude modulation.

The phase deviation usable in practice is about ±60°, the total theoretical phase shift being 180°. The distortion can be reduced considerably, if a valve with a non-linear *g*- characteristic is used whose precise shape is adjusted by negative feedback at audio frequencies, at the r.f. input level at which the modulator is to work. In a practical design a waveform distortion of 10 per cent was reached at 48° of deviation.

Simple circuits, non-critical adjustment, absence of tuned circuits resulting in low cost, wide band characteristics, reliable operation, absence of amplitude modulation, are the strong points of the circuit; these undoubtedly outweigh the disadvantage of the somewhat high distortion.

Acknowledgment

The author wishes to thank British Telecommunications Research Ltd for permission to publish this article which deals with work done by the author while with that Company.

REFERENCES

1. Terman, F. W. Radio Engineers' Handbook, p. 582 (McGraw Hill, 1943).
2. Beleskas, S. M. Phase Modulation Circuit. *Proc. Nat. Electronics Conf.*, p. 654 (Nov. 1947).
3. U.S.A. Pat. No. 2 143 386.
4. Boelens, W. W. Valve Characteristic Giving Linear Modulation when a Feedback Resistor is Inserted in the Cathode Lead. *Philips Res. Rep.* 3, 227 (1948).

A New X-Band Radar

The X-band radar, type FA, No. 13, Mk.1 (manufactured by A.P.T. Electronic Industries Ltd), is designed as a versatile general-purpose precision auto-tracking radar to meet a wide range of military and civil operational requirements. It incorporates a number of special features which are not generally available in existing radar equipments and has provision for the incorporation of further special facilities according to the user requirements.

The equipment is housed in a four-wheeled trailer similar to that of the radar type AA, No. 3, Mk. 7 and is provided with an air-conditioning system which cools the equipment and maintains comfortable working conditions for the operator.

Some special and unusual features of the equipment are as follows:

(1) The transmitter may be operated with single-pulse transmission as for normal primary radar operation or with a coded double-pulse transmission suitable for secondary radar operation in conjunction with a transponder beacon.
(2) Parts of the receiver system are duplicated so that it is possible to receive primary radar signals and secondary radar (beacon) signals at the same time. In particular, it is possible to track a target under primary radar conditions while at the same time tuning the secondary radar system in readiness for secondary radar tracking.
(3) The changeover from primary radar to secondary radar operation is instantaneous and may be done at any time and as often as desired during the actual tracking of a target.
(4) The ranging system is provided with a correction to take account of the intrinsic delay in response of a transponder beacon. By means of the facilities mentioned above it is possible to compare the target range under primary radar conditions with the target range under secondary radar conditions and to make the two equal. This removes the main source of range error in normal secondary radar operation.
(5) Data output transmissions from the radar are available in polar co-ordinates by means of Synchro transmitters and in cartesian co-ordinates by means of 400c/s analogue voltages generated in the co-ordinate convertor system of the radar.
(6) Mounting arrangements are provided for one or two automatic plotting tables to be fitted in the radar cabin if required. The plotting tables provide a working area of 75cm by 75cm and are operated directly from the data and power supplies of the radar.

Applications for which this radar equipment is particularly suitable include the following:

(1) Tracking of aircraft or missile targets in a military role, either operationally or for trials work.
(2) Tracking and remote control of drone aircraft targets.
(3) Tracking of civil air traffic, either for air traffic control or for monitoring purposes.
(4) Use as a precision approach radar for landing aircraft in poor visibility. Use of an auto-tracking radar for this purpose is not common but has special advantages in some circumstances. The built-in co-ordinate conversion system of the type, FA, No. 13, Mk. 1 permits the reference point of co-ordinates to be changed very rapidly, thus allowing a fixed-position radar to serve a number of different runways.
(5) Tracking of meteorological balloons, for wind-finding. The use of an X-band radar for this purpose permits the use of smaller and cheaper balloon targets and also provides indication of storm centres, heavy precipitation, etc.

All the low-power and medium-power circuits in the radar use modern transistor techniques, thus reducing power consumption and increasing the reliability. Thermionic tubes are used only in the high-power sections of the transmitter and in a few circuits where a good 'noise factor' is essential.

The radar equipment is fitted into a four-wheeled trailer and is fully self-contained. The equipment consists of the following major groups or sub-assemblies.

The aerial mount is capable of continuous rotation in azimuth and of motion in elevation from −3° to +87° approx. Azimuth and elevation drives are by means of twin-servo-motor differential gearbox units, capable of smooth control over a very wide range of tracking speeds.

The aerial consists of a 60in paraboloid which is energized by a waveguide feed nutating at 28c/s. Either linear or circular polarization may be selected, at will.

The presentation unit is a rack assembly which contains all the power supply, control and indicating circuits, other than those which are specifically parts of the transmitter-receiver unit. The presentation unit has a central operator's control panel, all normal control actions being by push-button control.

The transmitter-receiver unit is a separate rack which houses the transmitter and its power supply circuits, the head amplifier stage of the receiver and all the associated power supply circuits, metering and monitoring circuits, etc.

The cabin installation consists of the master power switching panels, the air-conditioning plant, the motor-amplidyne set which drives the servo-motors in the aerial mount, and all necessary interconnexion cabling and wiring.

The CIRRUS Coincident Current Core Storage Unit

I. R. Butcher*

A coincident current core store with a capacity of 16 384 nineteen bit words and a conservative 6μsec cycle is described. The design objective was to provide a reliable store of moderate cost rather than high speed. The X and Y selection circuits are based on the load-sharing switch, providing efficient but conservative usage of the driving transistors. Operation over a temperature range of 10 to 50°C with wide operating margins is made possible by compensating the drivers.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

DUE to its high reliability and relatively low cost, the transistor driven core store in recent years has become the standard form of random access storage for digital computers. For large capacity and/or medium speed stores, the coincident current selection principle[1,2] is preferred with linear selection[3,4] giving increased speed at higher cost. The design of CIRRUS[5] required a main store of 16 384 nineteen bit words with a 6μsec cycle. The large size and moderate speed made the coincident current store a natural choice.

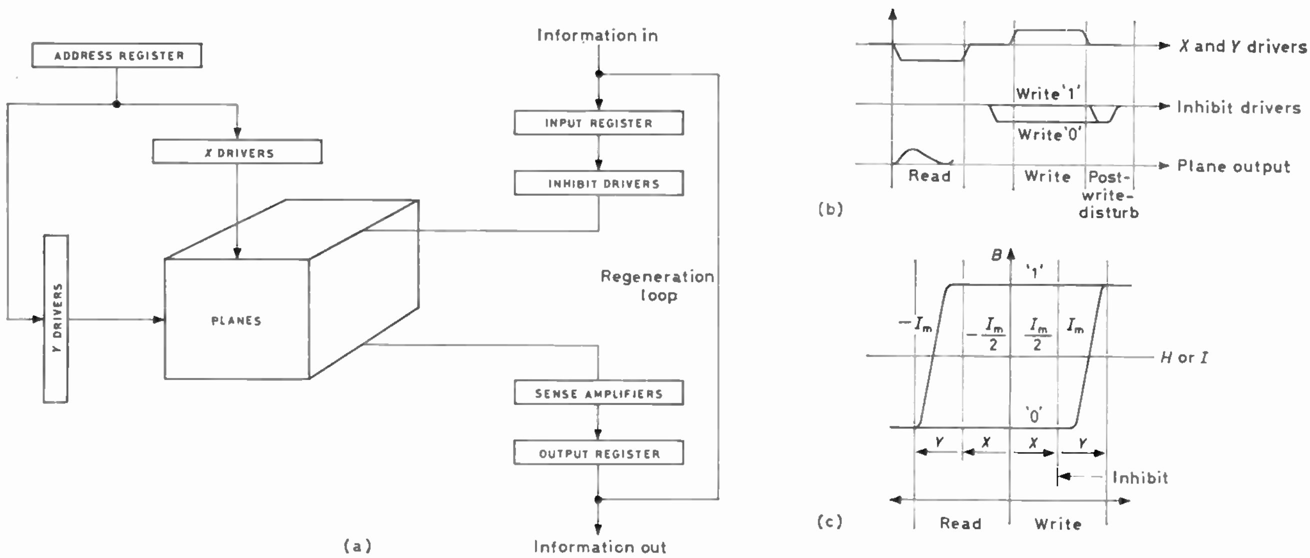


Fig. 1. The coincident current store
(a) General arrangement
(b) Timing cycle
(c) Hysteresis loop of storage core

The significant features of the store are shown in Fig. 1. The principle of both storage and selection depends on the shape of the hysteresis loop. In the absence of an applied field, the core will be in either of the two remanent states defined as the '1' and '0' states. As shown in Fig. 1(c) there exists a current I_m, such that this current will, if applied in the right direction, cause the core to change state, but a current of $I_m/2$ will leave the core state unchanged. This property is the basis of the selection principle. When wired into a storage array each core is linked by four sets of wires. One set links all cores along each X co-ordinate, another links all cores along each Y co-ordinate and the other two, the sense wire and the inhibit wire link all cores in the plane. A store is normally a three dimensional stack operated in a parallel mode. If a word has W digits, W planes are used and a word stored in the Z direction, occupying the same core position in each plane. The corresponding X and Y wiring of each plane is connected in series so that each plane receives the same drive currents. The store can then hold as many words as there are cores in a plane.

The store operates on a two phase cycle. During the read phase, currents of $-I_m/2$ are applied to the appropriate X and Y lines and the core on the intersection of these lines in each plane is set to the '0' state. No other cores can change state since the drive is less than the switching threshold. If a selected core had been in the '1' state, the resulting flux change will induce a voltage of 50 to 100mV in the sense wire linking the core. The flux change due to a '0' is much smaller, 5 to 10mV, and of shorter duration. The selected word will therefore appear in parallel on the W sense wires. For the write phase, currents of $I_m/2$ are applied to the X and Y lines. This, however, would set all the selected cores to the '1' state. If a '0' is to be written into a plane, the inhibit winding is energized with a current of $-I_m/2$. The selected core then receives a next current of $I_m/2$ and will remain in the

* *Research Laboratories of the Postmaster-General's Department, Melbourne, Australia.*

'0' state to which it was set during the read phase. Since reading is destructive, if the information read from the store is to be retained, it must be restored during the following write phase. When operated in this manner, the cores perform both storage and switching functions, since the well defined switching threshold allows part of the selection to be performed within the array itself.

The design objective of the CIRRUS store was to develop a reliable store of moderate cost. A sense amplifier and inhibit driver must be provided for each plane and little minimization other than developing efficient circuits is possible. Emphasis was therefore placed on developing economical X and Y selection circuits. The various techniques available for achieving the required 1 of n selection can be placed in one of two general categories, those using direct drive, with or without transformer coupling[6,7] and those using square loop magnetic switches[8,9]. In general, drive systems based on magnetic switches are cheaper than those using direct drive but switch efficiency tends to be low and the outputs are not as well defined. With the majority of drive systems, only one driver effectively contributes to the switch output and driver usage is rather inefficient. However, the load-sharing switch[10,11] is capable of combining and distributing the power from a number of sources, allowing the use of cheaper low power transistors. A selection system based on the load-sharing switch was chosen since this was one of the cheapest, provided efficient but conservative use of the drive transistors and produced well controlled outputs with low spurious signals.

TRANSFORMER \ DRIVER	1	2	3	4	5	6	7	8	9	10	11	12
0	+	−	+	−	+	+	+	−	−	−	+	−
1	+	−	−	+	−	+	+	+	−	−	−	+
2	+	+	−	−	+	−	+	+	+	−	−	−
3	+	−	+	−	−	+	−	+	+	+	−	−
4	+	−	−	+	−	−	+	−	+	+	+	−
5	+	−	−	−	+	−	−	+	−	+	+	+
6	+	+	−	−	−	+	−	−	+	−	+	+
7	+	+	+	−	−	−	+	−	−	+	−	+

Fig. 2. The load-sharing switch pattern

OUTPUT	READ						WRITE					
0	1	3	5	6	7	11	2	4	8	9	10	12
1	1	4	6	7	8	12	2	3	5	9	10	11
2	1	2	5	7	8	9	3	4	6	10	11	12
3	1	3	6	8	9	10	2	4	5	7	11	12
4	1	4	7	9	10	11	2	3	5	6	8	12
5	1	5	8	10	11	12	2	3	4	6	7	9
6	1	2	6	9	11	12	3	4	5	7	8	10
7	1	2	3	7	10	12	4	5	6	8	9	11

Fig. 3. Switch driver selection

The store itself is assembled from four blocks of 64 × 64 × 19 cores. The X and Y drive lines are connected to give a 128 × 128 array but the sense and inhibit windings are treated separately, so there are in effect four sense and four inhibit windings per plane. During construction of the store, planes with 128 × 128 cores arranged in four areas of 64 × 64 cores became available, but apart from allowing a substantial cost reduction would not require any system changes. The core used, Philips type 6C1, requires nominal drive currents of 250mA and switches in 0·9 to 1·0μsec to give 50mV output signal.

The X and Y Drivers

The X and Y selection circuits are based on a 12 input —8 output load-sharing switch. The switch is constructed from eight transformers each with 12 input windings of 8t and one output winding of 24t. The input windings are interconnected with the polarities shown in Fig. 2 and each of the 12 input lines is connected to a driver. Suppose drivers 1, 3, 5, 6, 7 and 11 are energized and deliver defined currents to the switch. Transformer 0 receives a positive excitation from each driver but all other transformers receive three positive and three negative excitations, and have no net drive. Output 0 will therefore be energized by the combined power of the six drivers and will deliver a positive output to the drive line attached to it. Similarly, by energizing drivers 2, 4, 8, 9, 10 and 12, output 0 will deliver a negative pulse and all others ideally will produce no output. By choosing the appropriate six drivers, any of the eight outputs can be selectively energized with the combined power of six drivers to produce either a positive or negative output. This principle can be extended to switches with any reasonable number of outputs. For those of interest, $n = 2^m$ outputs, the most efficient switch requires $n + 4$ inputs[11]. However a switch with the required 128 outputs is impractical. The selection principle is extended from 8 to 128 outputs by employing 16 switches as shown in Fig. 4. The outputs of a common set of drivers are steered to the appropriate switch by means of the 16 sink circuits and the diode network. An active (selected) sink has an output level of −20V which forward biases the diodes on the selected switch. The switch then adds and distributes the power of the six active drivers to the selected output to generate the required read and write drives. All other diodes are reverse biased by the voltage level, 0V, of the non-selected sinks and the other switches produce no output. A particular output is therefore selected by selecting the appropriate six drivers and one sink.

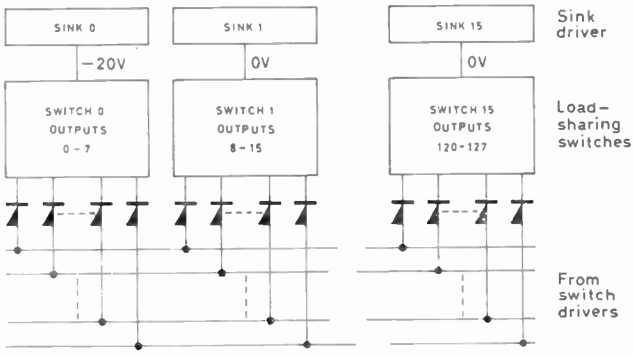


Fig. 4. Switch selection

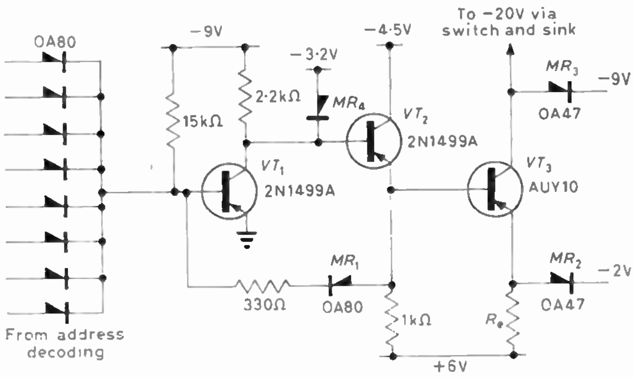


Fig. 5. The load-sharing switch driver

The basic driver circuit is shown in Fig. 5. If all inputs are negative, VT_1 will conduct and hold VT_3 off. If any input goes positive, VT_1 will be turned off and the base of VT_3 will be taken to approximately −3·5V. Under these conditions VT_3 conducts, reverse biases MR_2 and delivers a current of $\alpha/R_E(9{\cdot}2 + V_{MR4} - V_{be2} - V_{be3})$ amperes to the switch. Allowing for reasonable variations in α and the forward voltages of the diodes, the current is defined to an accuracy of $\pm 2\frac{1}{2}$ per cent. MR_3 ensures that VT_3 maintains an operating point well out of satura-tion by conducting if a voltage transient takes V_{c3} more positive than −9V. The load-sharing switches have a turns ratio of 1:3, requiring six input currents of 125mA to generate the required 250mA output current. In practice the need for damping resistors across the switch to control the voltage developed by the inductive load increases this to 140mA per driver.

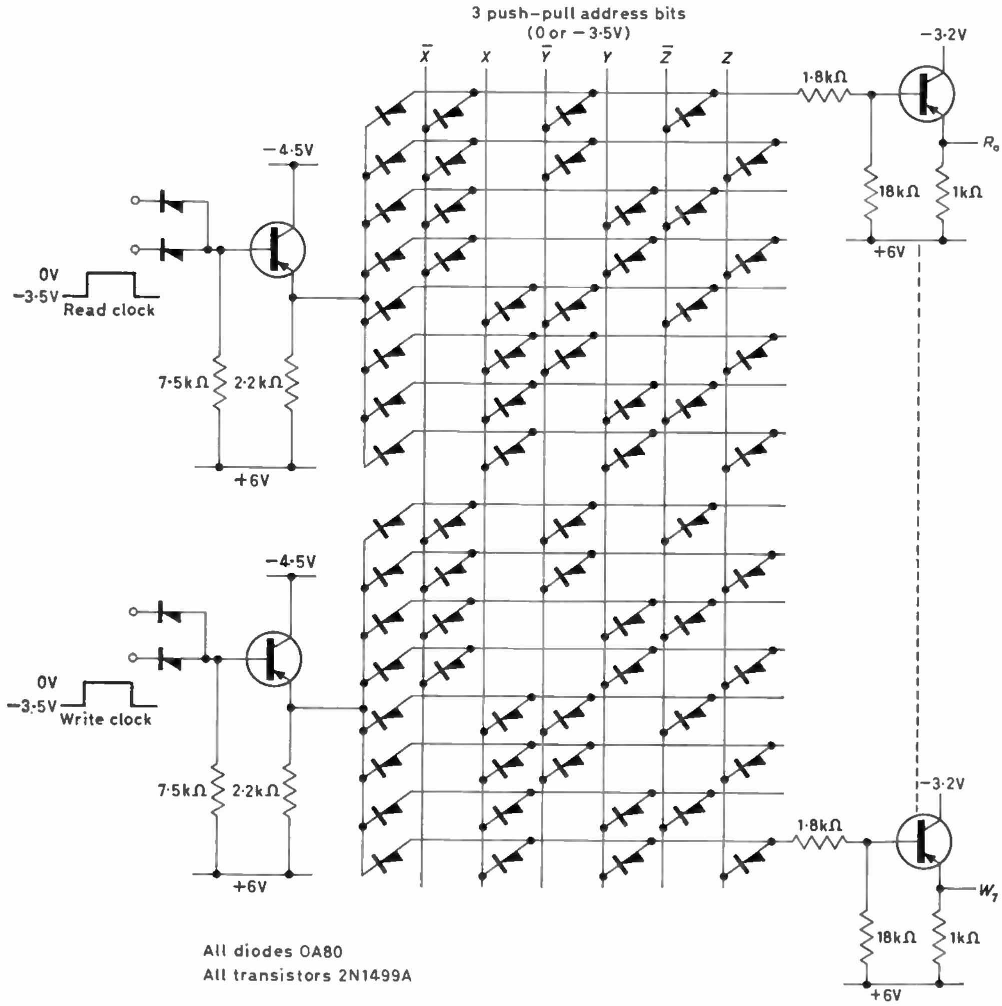


Fig. 6. The driver selection circuits

Fig. 6 shows the driver selection circuits. The decoding packages are standard units which provide a 1 of 8 decoding from three push-pull address bits with an overriding input for control or timing functions. In the absence of a clock waveform, all outputs are at −1V. If either clock waveform rises to 0V, one of the 16 outputs will be taken to +1V, and will turn on six drivers by means of the OR diodes in the driver. The appropriate decoding has been listed in Fig. 3.

The sink circuit is shown in Fig. 7. If the input is at −1V, VT_1 is on, VT_2 and VT_3 are off, and the output is at 0V. When the input is taken to +1V, VT_2 and VT_3 conduct and the output is at −20V, since the base of VT_4 is tied to the −20V supply through a saturated transistor. In this condition VT_4 is capable of absorbing the pulse of 780mA applied when the drivers conduct during a read or write operation. The selection of 1 of 16 sinks is achieved with two decoding packages. In this case a fourth address bit is applied in push-pull to the control gates so that one of the packages is logically prevented from allowing an output to rise to +1V.

Inhibit Drivers

The basic inhibit driver (Fig. 8) is very similar to the load-sharing switch driver. The input logic allows the driver to generate the required 250mA output if both the information input and inhibit clock are at 0V or if the post-write-disturb clock is at 0V. It will be noted that this produces an inversion of normal core operation, i.e., the core is set to '1' during read. A switching core therefore represents a '0' output instead of the conventional '1'. As mentioned earlier, the inhibit winding is split into four sections of 64×64 cores each. During either write or post-write-disturb, it is only necessary to energize that section of the plane which contains the selected core. To avoid having four inhibit drivers/plane, a 1 of 4 steering circuit is attached to each inhibit driver as shown in Fig. 9. By saturating the appropriate steering transistor the inhibit current can be steered to the correct section of the inhibit winding. The diodes in the base circuit of the OC140's are necessary to prevent damping the fall time of the inhibit current. At this time the collector voltage of the AUY10 and the unselected OC140's rises to

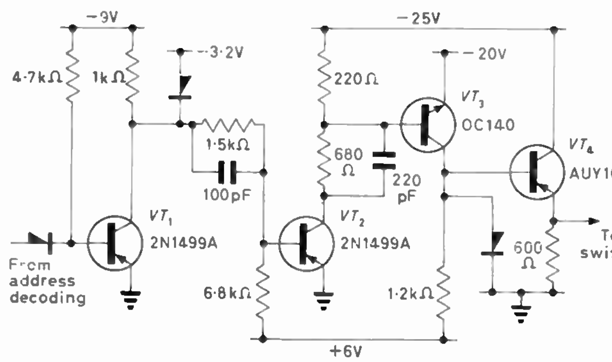


Fig. 7. The sink driver

approximately −30V forward biasing the *c-b* diode of the OC140's. The diodes prevent the current flow that would otherwise result. The 120Ω resistor and 0·01μF capacitor are necessary to control the back e.m.f. generated by the inhibit winding. The selection of the appropriate set of steering transistors is achieved by techniques identical to those described for the load-sharing drivers and the inhibit driver. A 1 of 4 decoding from the two most significant address bits selects 1 of 4 possible block current switches to generate a 380mA output. This is applied to the appropriate block input of each inhibit driver and due to the 620Ω resistors in series provides approximately 20mA to each of the 19 steering transistors. The steering transistors for the other three blocks have 5V reverse bias applied since an inactive current switch is very nearly open-circuit.

Output Circuits

An inspection of the layout of the four blocks of memory suggests that two sense amplifiers can handle the signals arising from the four sense windings by grouping into diagonal pairs. If the selected core is in block 0 (Fig. 10) then the sense winding of block 3 produces no signal at all so these two can use the same amplifier. Blocks 1 and 2 will produce a small signal during the rise and fall of the drive currents but these will have decayed by strobe time, and will have no effect on the total plane output. In practice this worked quite well although the output was reduced due to the loading of the other plane. The signals coupled to the sense winding by the *X* and *Y* drive currents were trivial and the outputs were very clean during the read phase. The major trouble experienced was with the noise coupled from the post-write-disturb or inhibit pulse. The extra loading of the other plane, increased the natural decay time of the transient and increased amplifier recovery time. It was still possible however to achieve the desired cycle time under the worst case as discussed in a later section. The resulting sense amplifier is shown in Fig. 11.

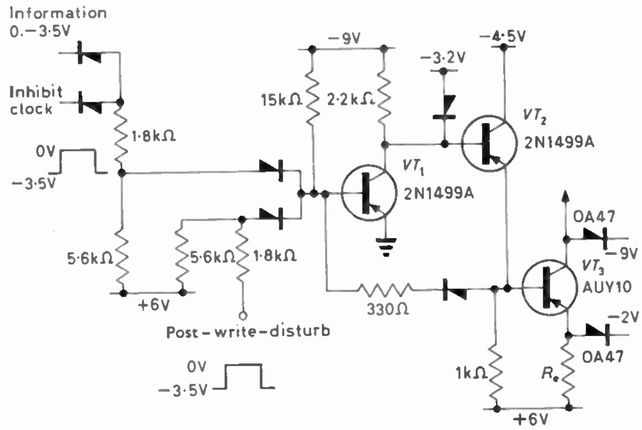


Fig. 8. The inhibit driver

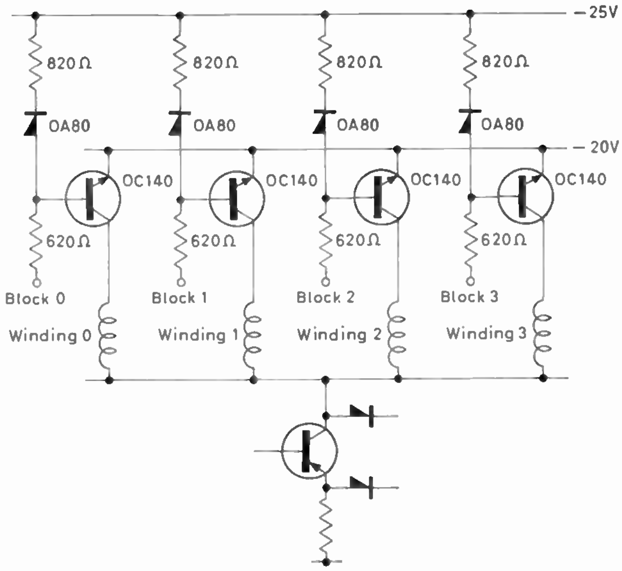


Fig. 9. The inhibit steering circuit

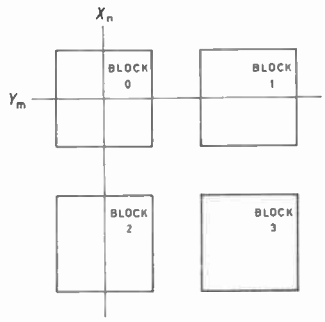


Fig. 10. Grouping of sense windings

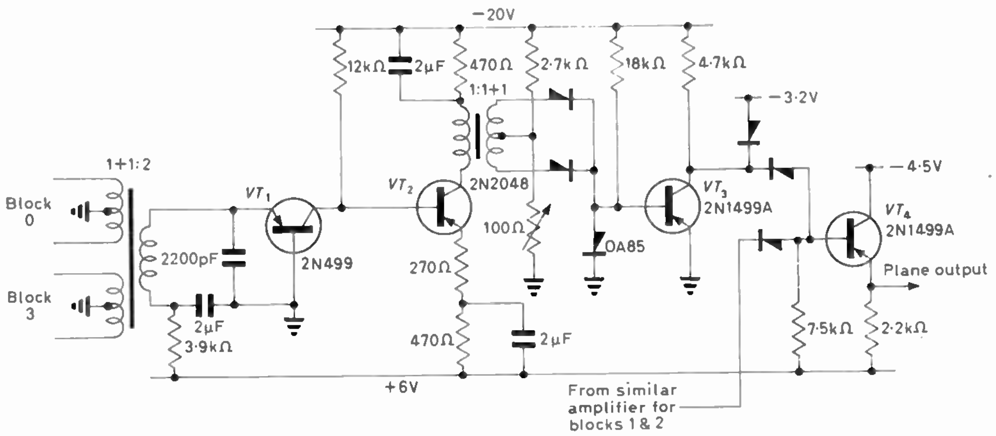


Fig. 11. The sense amplifier

Each input is applied to a centre tapped primary to eliminate the common mode signals arising from capacitive coupling to the drive lines. A common mode rejection ratio of >200:1 is achieved by careful transformer design. The first stage uses the grounded base configuration largely for impedance matching, and amplifies the plane output to approximately 2V. The second stage provides the necessary power gain and some voltage gain to drive the rectifying network and threshold circuit on the base of VT_3. The rectification is required to handle the bipolar output signal arising from the winding configuration of the sense winding. A switching core will produce a positive output of approximately $1\frac{1}{2}$V on the base of VT_3 taking the output to $-3{\cdot}5$V, the voltage level representing a '0'. The amplifier for sections 1 and 2 of the sense winding is identical up to VT_3. The outputs are then combined in a negative OR gate to provide the plane output.

The combination of grounded base and degenerate grounded emitter configuration for the first two stages provides an amplifier with very good high frequency response and a gain largely independent of transistor parameters. The use of the 2N499 in the first stage was necessary to provide rapid recovery from overload.

No provision is made for strobing in the amplifier since this is achieved by sampling the output with the $0{\cdot}2\mu$sec clock which inserts the information into the output register.

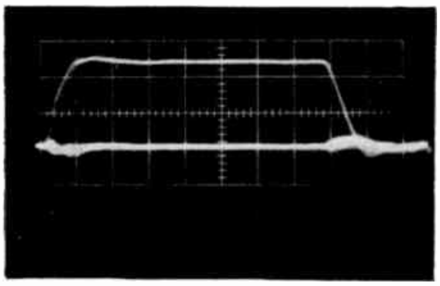


Fig. 12. Drive currents 100mA/cm, 0·5µsec/cm

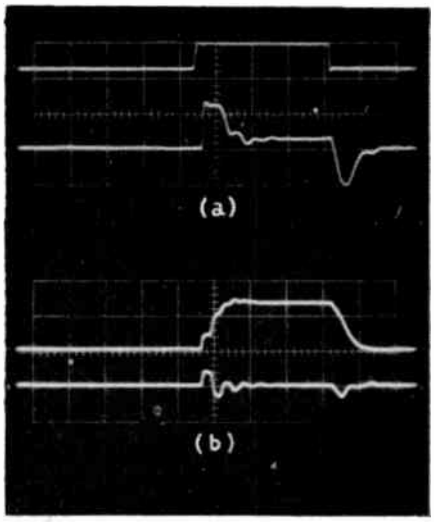


Fig. 13. Inhibit driver waveforms
Inhibit clock—5V/cm
Collector voltage—10V/cm
Current in selected block—200mA/cm
Current in non-selected block—200mA/cm
0·5µsec/cm

This form of sensing does not make full use of the improved discrimination provided by the strobe technique since the response of the threshold transistor is too slow to faithfully follow the amplitude variations in the applied signal. However, the store output signals during the read period (Figs. 14 and 15) were clean enough to allow discrimination on the basis of signal amplitude and no trouble was experienced with this technique.

Temperature Compensation

Due to the relatively low Curie temperature of the majority of square loop ferrite materials, the coercive force and therefore optimum drive currents have a temperature coefficient of approximately $-0{\cdot}5$ per cent/°C.

Since most cores have a knee current of 60 to 65 per cent of the nominal drive current a memory can be operated over a temperature range of about 40°C provided the drive currents can be held to very close tolerance. Adequate temperature compensation however, ensures that the core will at all times be used under optimum conditions and yields a potentially more reliable system.

Tests on the core used indicated a temperature coefficient of $-2{\cdot}5$mA/°C over the required temperature

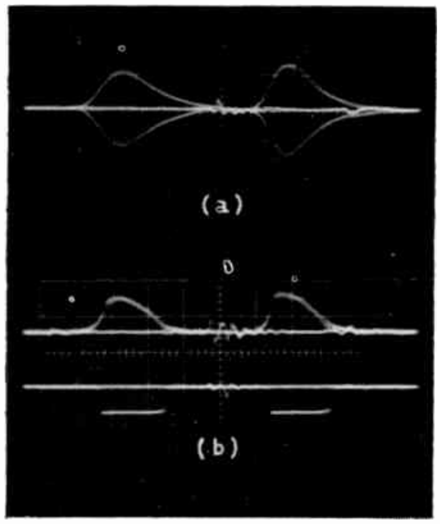


Fig. 14. Sense amplifier waveforms
V_{C1}—2V/cm
V_{b3}—1V/cm
Output—5V/cm
1µsec/cm

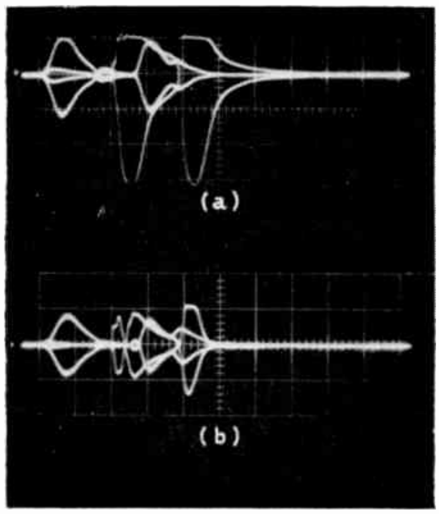


Fig. 15. Plane output for noisy pattern
(a) Balanced drive currents
(b) Unbalanced drive currents
V_{C1}—2V/cm
1µsec/cm

range of 10 to 45°C. With the current switching technique used in the driving circuits the drive currents can be varied by adjusting the +6V supply. Consequently the current defining resistors in the emitter circuit were returned to a separate +6V regulator, the voltage of which was adjusted by means of a thermistor network.

With the compensation thus provided the core output and operating margins were held essentially constant over a temperature range of 0 to 50°C.

Experimental Results

The two most important aspects of the driving circuits, the uniformity of the output currents and the ratio of selected to non-selected output are shown in Fig. 12. In both cases, the departure from the ideal behaviour is due to mismatch of the load-sharing drivers either in amplitude or timing. The performance of a driver is summarized in Fig. 13. An inhibit driver is chosen since this is the more heavily loaded unit. The back e.m.f. presented by the 4096 cores on the inhibit winding is shown reflected on to the collector waveform. The clamping action of the diode to -9V can be seen during the rise of the current waveform. The diode regulates the rate of rise of the current through the inductance of the plane in such a manner as to maintain the transistor out of saturation. The excess current is bypassed to the -9V supply. If the transistor is allowed to saturate, the accurate definition of collector current is lost and a considerable overshoot occurs.

Typical waveforms through the read amplifier for a '1' and '0' are shown in Fig. 14. These were taken with patterns of all '1's and '0's recorded in the plane. Under these conditions the inductive noise coupled from the drive winding to the read winding is minimized. The remaining noise source, capacitive coupling, has been eliminated in the input transformer. The worst case occurs when every core linking the read winding in a positive sense is in one state and every core linking in a negative sense is in the other state. The sense wires of the planes used were wired in the conventional diagonal pattern. The method of threading is such that a current of 500mA applied to the sense wire will set the core into states approaching the worst noise case. This pattern was then read out and it was found that the pattern could be derived logically from the address bits by the expression $(A_6 \overline{A_7} + A_7 \overline{A_6})\, A_1 + (A_6 A_7 + \overline{A_6 A_7})\, \overline{A_1}$. A noisy pattern could then be readily established in any plane. The noise from this pattern is much greater (Fig. 15(a)) and amplifier recovery due to the inhibit transient extends into the next read period. This is due more to the long time-constant associated with this noise than to its magnitude and is accentuated by the loading of the other plane. It was found, however, that an unbalance of the drive currents was very effective in reducing the noise. The optimum values, determined experimentally were:

$$I \text{ read} = 490\text{mA}$$
$$I \text{ write} = 530\text{mA}$$
$$I \text{ inhibit} = 260\text{mA}$$

The noise voltages with the new values of drive currents are shown in Fig. 15(b) for the worst case. The signals during read are particularly clean and for a typical random pattern in the plane are little different from the all '1' and all '0' signals. The unbalance of current can conveniently be achieved by changing the $-3{\cdot}2$V clamp voltage in the switch drivers. A flip-flop associated with the store in the control unit changes state at the appropriate times and provides the required voltage levels of $-2{\cdot}6$V and $-3{\cdot}2$V by means of two level-shifting networks. With the drive currents adjusted in this manner the tolerance is slightly reduced but the performance proved to be quite satisfactory over a wide range of operating conditions. Experience also indicated that the post-write-disturb pulse was unnecessary, so this was eliminated allowing another $\frac{1}{2}\mu$sec for amplifier recovery.

Construction

The store including regulators is constructed in a frame approximately 3ft × 2ft × 6in. This does not include any of the static registers which due to the machine structure are part of the general hardware. All low level circuits are assembled in packages 6in × 2½in and the high power output circuits are mounted directly on sections of heat sink which are treated as a plug-in unit. Some 600 transistors are used only 80 of which have a dissipation greater than 100mW. The total power consumption for the store is about 200W. The unit is required to operate up to an ambient temperaure of 45°C so two fans are employed to eliminate the possibility of local hot spots.

Conclusion

The objective of the design was to develop a store making the most efficient use of hardware in an effort to reduce the overall cost. In the store described this was reduced to less than 35 per cent of the total memory costs allowing for the lower cost of the larger planes now available. All voltages and currents could be varied by at least 10 per cent before failure occurred and the margins in timing were such that the cycle could be reduced to 5μsec under worst case conditions without failure. As yet no data is available on long term reliability but the results to date indicate that this will be quite satisfactory.

Acknowledgment

The work described in this article was part of a larger project carried out in the Electrical Engineering Department of the University of Adelaide with financial assistance from the Postmaster-General's Department and the Weapons Research Establishment. The author is indebted to his colleagues, particularly Dr. M. W. Allen for helpful discussions and criticism and to the Engineer-in-Chief of the Postmaster-General's Department for permission to publish.

REFERENCES

1. Forrester, J. W. Digital Information Storage in Three-Dimensions using Magnetic Cores. *J. Appl. Phys.* *22*, 44 (1951).
2. Allen, C. A., Councill, E. C., Bruce, G. D. A 2·18 Microsecond Megabit Core Storage Unit. *I.R.E. Trans. Electronic Computers.* *102*, 233 (June 1961).
3. Renwick, W. A Magnetic Core Matrix Store with Direct Selection using a Magnetic Core Switch Matrix. *Proc. Instn. Elect. Engrs.* *104B*, Suppl. 7, 400 (1957).
4. Rhodes, W. H., *et al.* An 0·7 Microsecond Ferrite Core Memory. *I.B.M. J. Res. Developm.* *5* 174 (July 1961).
5. Allen, M. W., Pearcey, T., Penny, J. P., Rose, G. A., Sanderson, J. G. CIRRUS, an Economical Multiprogram Computer with Micro Program Control. (Submitted to *I.R.E. Trans. Electronic Computers*).
6. Padwick, G. C., Cain, A. L. Transistor Circuits for a Ferrite Store. *Proc. Instn. Elect. Engrs.* *106B*, Suppl. 16, 675 (1959).
7. Ashley, A., *et al.* Core Memory Systems. *Sylvania Technol.* *12*·4, 140 (1959).
8. Rajchman, J. A. Static Magnetic Matrix Memory and Switching Circuits. *RCA Rev.* *13*, 183 (June 1952).
9. Richards, R. K. Digital Computer Components and Circuits. (D. Van Nostrand. 1957).
10. Constantine, G. A Load-sharing Matrix Switch. *I.B.M. J. Res. Developm.* *2*, 204 (July 1958).
11. Chien, R. T. A Class of Optimal Noiseless Load-sharing Matrix Switches. *I.B.M. J. Res. Developm.* *4*, 414 (Oct. 1960).

Closed-Circuit Television for Propeller Testing

Variable-pitch ejectable propellers, which provide a stand-by electrical or hydraulic power supply in an emergency on VC 10 and Trident aircraft, are being tested at Dowty Rotol Ltd's Gloucester plant by an ingenious technique using a stroboscope in conjunction with a closed-circuit television system, both supplied by EMI Electronics Ltd.

Using the new technique, a stroboscope 'freezes' the propeller as it rotates on the test rig at a known speed. The light from the stroboscope is reflected along the line of sight of the camera by a polished stainless steel sheet with a central hole through which the camera looks. A type 8 camera views the blade in its plane of rotation and a picture of the blade root is clearly seen on the television receiver outside the test cell.

The blade root has a scale marked on it which lines up with a fixed pointer on the hub and from this arrangement the blade angle can be assessed within an accuracy of ±1 per cent. In this way engineers can check whether, under flight conditions, the blade will rotate at the correct speed.

Some Precision Direct Coupled Transistor Amplifiers and Their Approximate Design

(Part 2)

By C. W. B. Grigson*

(Voir page 503 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 510)

A Precision Current Source giving any Desired Current Between 0 and 20A

Some Comments on Stabilized Supplies

Stabilized supplies, whether designed to give a constant voltage or a constant current, are just another case of a d.c. amplifier. The amplifier is connected between the reference, which is the input, and the load. The unstabilized supply fixes a node voltage inside the feedback loop, and effects of it on the load are reduced by the loop

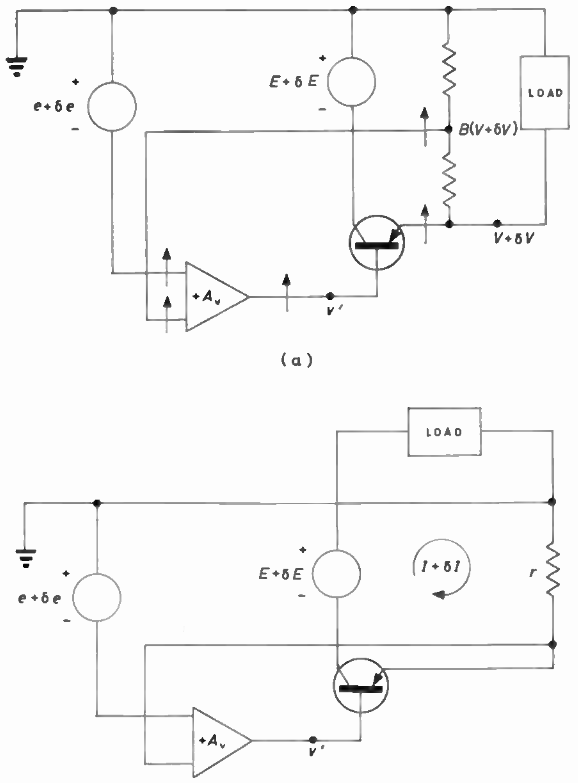


Fig. 11. Sequence diagrams for stabilized sources regarded as d.c. amplifiers
(a) voltage
(b) current source: e is the voltage reference, E the unstabilized main power supply, V or I the stabilized output

gain. Fig. 11(a) shows the sequence diagram of a constant voltage supply from this point of view. Fig. 11(b) shows the circuit to be described. A_v is a ring-of-three tailored to suit the input impedance of the current amplifier. The latter is usually a power transistor or a compound group of transistors. The unstabilized supplies are connected to the collectors of this stage. Fig. 11(a) shows a single transistor, for this case a nodal analysis gives:

$$\delta V = (\delta e/B) + \frac{\delta E\,(r_b/\beta r_d)}{AB} \quad \ldots\ldots\ldots\ (32)$$

* *Engineering Department, University of Cambridge.*

here δe is the drift or ripple in the reference, δE that in the unstabilized supply, δv that in the load. It is assumed the voltage loop gain $AB \gg 1$.

Equation (32) applies equally to total quantities V, e and E, provided that their values are within a range in which no part of the amplifier is cut off or saturated. One sees at once the dependence of V or δV on the reference e or δe. And that δe must be as small as possible: therefore e should be derived from a Zener diode and resistance chain connected to the stabilized output itself. Moreover e can be made variable, this gives a convenient adjustment of V. Often e is derived from a Zener diode energized by a separate unstabilized supply and V is varied by altering B. Fig. 11(a) shows also a way of making the series transistors safe from overload: one adds a series resistance R, ($R \ll r_d$ of the transistor). This is a simple alternative to the electronic cut-outs sometimes used and is feasible when V is not too large.

Fig. 11(b) shows the corresponding sequence diagram for a constant current supply. For this, see Appendix 1.

$$\delta I = -(\delta e/r) + (r_b/\beta r_d)(\delta E/A_v r) \quad \ldots\ldots\ (33)$$

provided $A_v A_g r \gg 1$, where A_g† is the conversion gain between load current I and node voltage v'. A similar equation applies to total quantities I, e, E. For such constant current supplies I may be much more conveniently adjusted by altering e than by altering r; since it is quite difficult to make a stable variable resistance capable of handling large currents and of staying cool.

The variational equation (33) shows, again, that δE is multiplied by the very small factor $r_b/\beta r_d$ as well as the factor $1/(A_v r)$.

Description and Performance of 20A Supply

Fig. 12 shows the circuit details of a 20A supply which has the sequence diagram just discussed. A_v is the ring-of-three which is similar to the previous ones, and is designed to give up to 25mA at 6V. A_v supplies A_c, a compound common emitter circuit in which one power transistor drives 17 others in parallel. The latter were selected for uniformity of collector current at a given base-emitter voltage, the scatter of I_c being ± 10 per cent. Any constant current up to 20A can be provided at a maximum load voltage of 20V and an output impedance of 300Ω or more. 100c/s current ripple at 2A output was 3·5mA peak-to-peak, and at 20A output it was 20mA peak-to-peak.

The circuit was built and tested by J. H. Hinton* who found the following experimentally determined additions to be necessary, Fig. 12, $C = 1\mu$F prevents a high frequency oscillation at low values of output current; $R_1 =$

TABLE 2

I	2	5	10	15	20	(A)
Z_{out}	490	470	440	380	300	(Ω)
δV	17	15·5	6·2	9·0	5·6	(V)
δI	3·5	33	14	23·6	18·7	(mA)

* *As part of the instrumentation of a high frequency plasma Ph.D. project in this laboratory.*
† *A_g is written as μ in the appendix.*

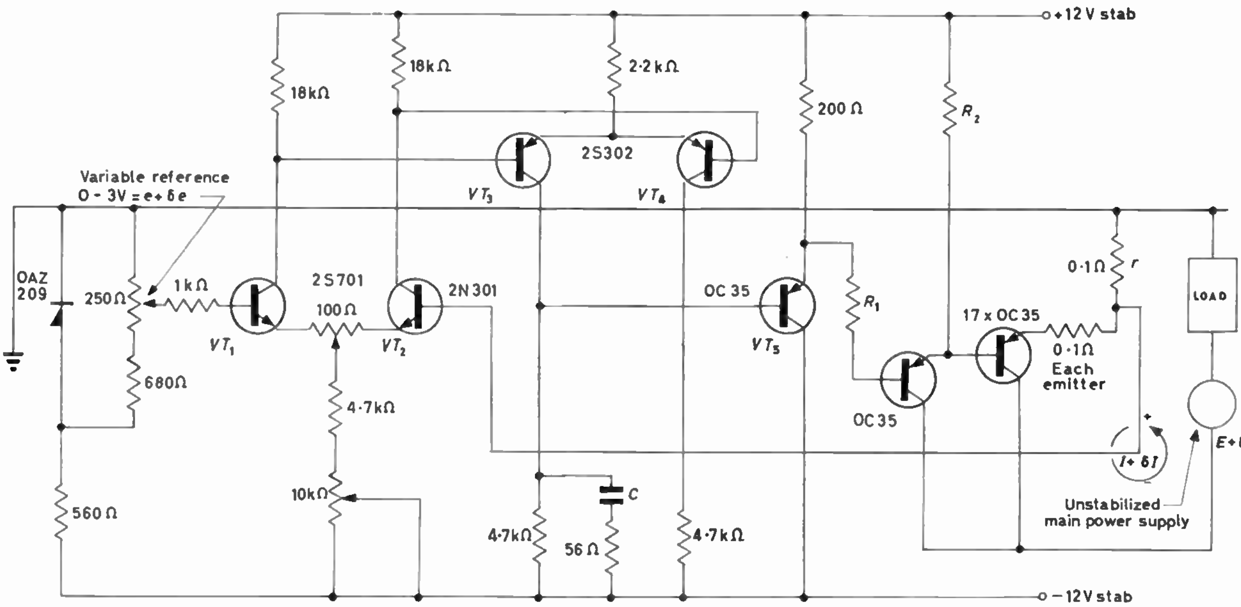


Fig. 12. 0-20A constant current source 300Ω output impedance

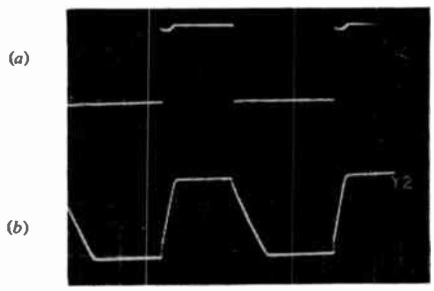


Fig. 13. Response of constance current source to a square-wave reference voltage at 50c/s

(a) input 1·5V peak-to-peak

(b) output 15A peak-to-peak, quiescent current 13A

33Ω limits the current drawn when the output stages are turned hard on; $R_2 = 1{\cdot}2\text{k}\Omega$ injects reverse bias, without it the output current could not be reduced to less than 0·3A.

Table 2 shows the output impedance at different current levels. Z_{out} was obtained as the ratio of load voltage change δV to load current change δI when part of the load was short-circuited.

If, instead of the constant input voltage from the Zener circuit, an a.c. voltage is used, the amplifier should reproduce the waveform of the input as a current. Fig. 13 shows the current waveform in response to a square wave of input voltage. The limiting rate of rise or fall, of uniform slope, is set by the constant tail current of the second long-tailed pair all of which, as a limit, can be routed through VT_3, to charge C.

$i/C = dV/dt|_{max} = 3\text{V/msec}$ which is the value found.

High Capacity D.C. Amplifiers

If d.c. amplifiers are to deliver much power then the output stage must work in class-B, ideally in the complementary symmetry push-pull circuit. But the latter requires matched npn-pnp pairs, and large npn transistors are expensive. The account given is limited to amplifiers with non-complementary d.c. coupled class-B stages, and the design procedure outlined deals, again, with rings-of-three. For this type of amplifier unmatched pnp germanium power transistors are used, and may be operated in parallel to give considerable outputs: it looks as though there is no essential limit to the power possible.

The difficulty in the design of this kind of wholly pnp class-B amplifier is that it is not balanced. One phase of the output circuit works in common-emitter, the other in common collector. Therefore not only must the drive signals be transferred across different d.c. levels, but they see different impedances. Since the phases are inherently unbalanced, the design does not require matched output transistors and successful circuits need not use them.

A glance at the collector characteristics of power transistors will show that if high currents are wanted then low collector voltages only are possible, otherwise the operating point will cross the collector-dissipation curve. In an OC35 in class-B the operating point might swing from $I_c \times V_{ce} = -0{\cdot}5 \times -8\text{AV}$ to $-8 \times -0{\cdot}5\text{AV}$. If higher load voltages are needed their peak currents must be small and the number of parallel transistors must increase sharply. The maximum load voltage is not in any case high, because of the limitation:

$$V_{ce(max)} \simeq 2V_p$$

The most costly part of amplifiers of this kind is the d.c. supply for the output stage. It must be low impedance, it must have three rails at about $+V_p$, 0 and $-V_p$, with a current capacity of I_p, where V_p, I_p are peak values of the output signal.

Fig. 14 shows the general type of circuit, Fig. 15 current and voltage waveforms. The phases are labelled CE and CC from the phase splitter onwards along the signal lines. A first-order design will be developed. All transistors will be assumed to have the same average value of current amplification factor β, to have collector branch resistance r_d large compared with the other relevant impedances, and that peak base-emitter voltages are 400mV greater than the standing V_{be}. It will be clear how to modify the design to allow for differing β values; values of r_d could be estimated from collector characteristics and of V_{pbe} from input characteristics. Such refinements add nothing to design certainty since the calculations are dominated by β, which is unknown in value to a factor of three. It is

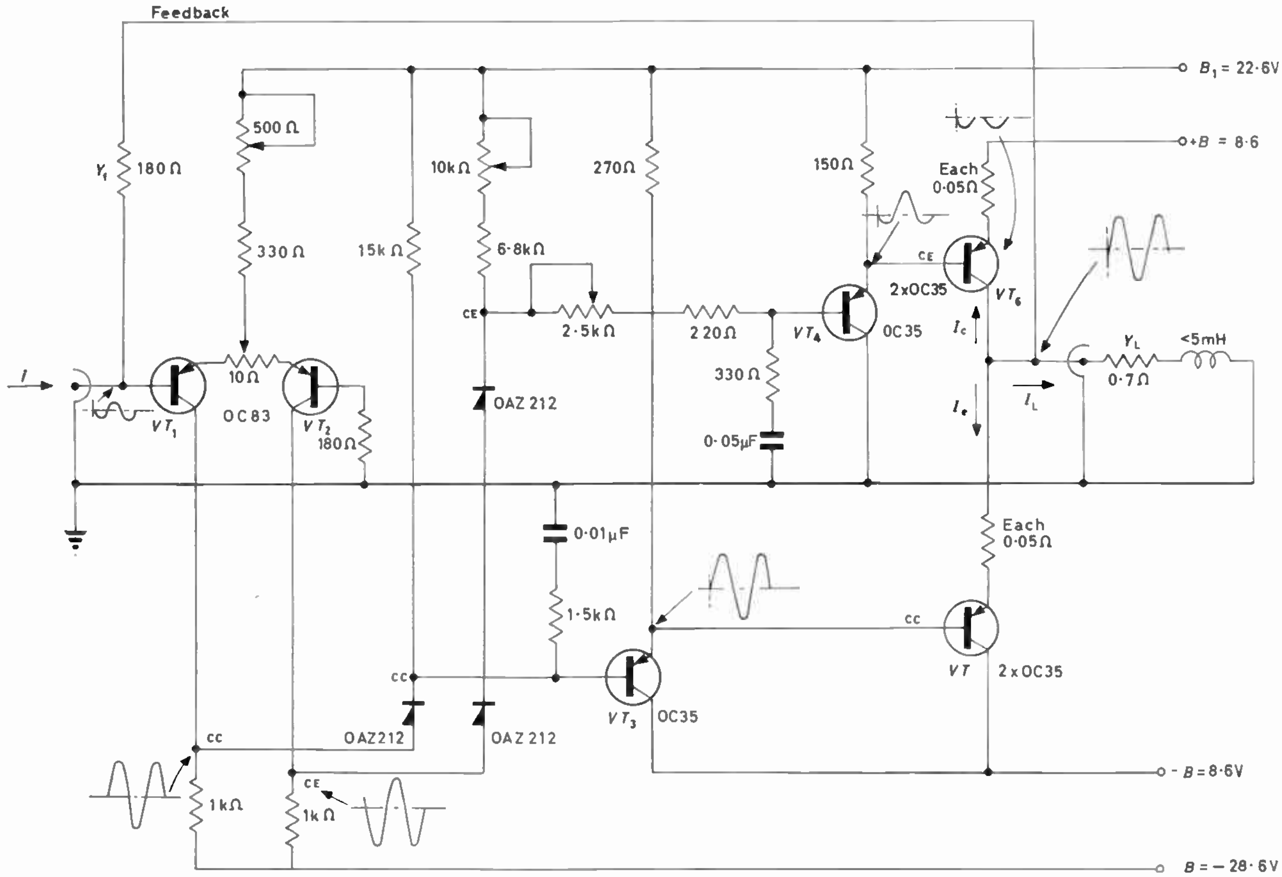


Fig. 14. Class B ring-of-three d.c. amplifier capacity 16A peak-to-peak into 0·7Ω, 20A peak-to-peak into 0·25Ω

contended, once again, that a more complicated design analysis is worthless. The design method given is adequate to permit calculation of all components except those of the high frequency cut-off network. Development of the circuits must include some laboratory work, the setting up of the d.c. levels and killing of the h.f. oscillations. These will occur since a lot of feedback should be used. Current gains for the open ring-of-three are of the order of 100 000, and with feedback as low as 200. The feedback stabilizes not only the gain, but the thermal drift of the power stage.

Inspection of the circuit (Fig. 14) and the waveforms of current (Fig. 15(c,d)) show that for the output transistors:

$$-I_{\text{load}} = I_{\text{emitter}}\Big|_{\text{CC phase}} + I_{\text{collector}}\Big|_{\text{CE phase}}$$

the quiescent current in each transistor I_q is usually between 1/10 and 1/5 of I_p.

The voltage waveforms show that the CC phase, VT_5, needs a signal $\simeq$ V$_p$ in magnitude, quiescent level $V_q \simeq$ earth, VT_6, in the CE phase, requires a voltage signal in peak amplitude, d.c. level $\simeq +B$.

VT_5 and VT_6 are driven by VT_3 and VT_4, emitter-followers in class-A. These, in turn, are supplied from VT_1 and VT_2, a long-tailed-pair phase splitter which is preferably balanced. If so VT_1 must handle an a.c. signal of $\simeq V_p$, the quiescent collector voltage of VT_1 must be $\simeq -V_p = -B$, and the line B_2 of voltage at least $-2B$, in order to handle this swing without ever being cut off. If the pair is to be handled VT_2 must handle a similar signal. Therefore Zener diodes must be included to shift d.c. levels from the collector of VT_1 and VT_2 (at $-B$ quiescent) to that of the bases of VT_3 and VT_4 (respectively $\simeq 0$ and $\simeq +B$).

An impedance changing network must also be included between VT_2 and VT_4. The most direct way of changing the impedance might seem to be to add series resistance of the right size to the base circuit of VT_6, but this is unwise since to reduce any tendency to thermal drift or runaway the resistance in the base of common emitter connected power transistors should be kept as low as possible. For the same reason VT_4 and VT_3 are operated entirely in class-A so that the bases of VT_6 and VT_5 see always the low variational output resistance of emitter-followers. The line voltage B_1 is large enough for the pair VT_1 and VT_2 to be fed through a large resistance and to increase the resistance in the emitters of VT_3 and VT_4. In practice B_1 and $B_2 \simeq |2B|$; lines B_1 and B_2 each require a relatively small current capacity.

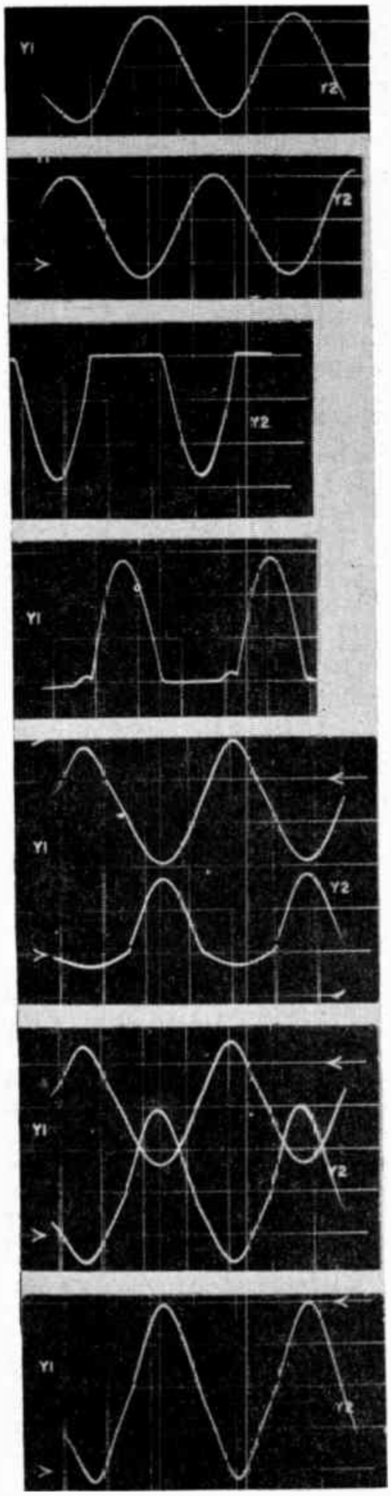
The Output Stage

I_p and V_p are fixed by the load resistance R and the load power demanded. From an inspection of collector characteristics one decides a quiescent current $I_q \gg \beta I_{CO}$, and sufficient to minimize crossover distortion. I_q is small. From the input characteristics one obtains V_p', the base-emitter voltage necessary to swing the collector current from I_q to I_p. $V_p' \simeq -0{\cdot}8$V with considerable scatter. (For the OC35 the figures are $-0{\cdot}8$ median, $-1{\cdot}4$ max $-0{\cdot}4$ min for V_{be} for swings from low to large currents.

Let suffices q, p refer to quiescent and peak signal states respectively. The bases must both have currents:

$$I_{qb} = I_q/\beta - I_{CO}^{*} \quad \ldots\ldots\ldots\ldots\ldots \quad (34)$$

* *The distinction between β and β + 1 is here ignored.*



(a) Input current 2·8mA/division

(b) Load current 7A/division

(c) Collector current in one of the two transistors VT_6 1·4A/division

(d) Emitter current in one of the two transistors VT_5 1·4A/division

(e) Emitter voltage VT_3 5V/division

(f) Emitter voltage VT_4 5V/division

(g) Collector voltage VT_1 5V/division

(h) Collector voltage VT_2 5V/division

(i) Input voltage, base VT_1 0·2V/division

Fig. 15. Waveforms in Class-B d.c. amplifier

$$I_{pb} = I_p/\beta \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (35)$$

In the common-collector phase:

Base d.c. level, for germanium power transistors

$$V_{qb} = 0 - 0{\cdot}3 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (36)$$

Base peak voltage

$$V_{pb} = V_p + V_p' + I_pR_e = V_p \quad \ldots\ldots\ldots \quad (37)$$

The variational input impedance is:

$$R_{in} = \frac{V_p + V_p' + I_pR_e}{I_p/\beta} \simeq (\beta V_p/I_p) = \beta R_L \quad \ldots\ldots \quad (38)$$

For the common-emitter phase corresponding quantities are:

$$V_{qb} = B - 0{\cdot}3 \quad \ldots\ldots\ldots\ldots\ldots \quad (39)$$

$$V_{pb} = V_p' + I_pR_e = 0{\cdot}8 + I_pR_e \quad \ldots\ldots\ldots \quad (40)$$

$$R_{in} = \frac{V_p' + I_pR_e}{I_p/\beta} = \beta\,[0{\cdot}8/I_p + R_e] \quad \ldots\ldots \quad (41)$$

These expressions follow from the waveforms of Fig. 15 and give the quiescent conditions to be provided by the drivers. The assumptions made about base-emitter voltages are tantamout to regarding the base-emitter diode as nearly ideal, with a break point at −0·3V and a low forward resistance.

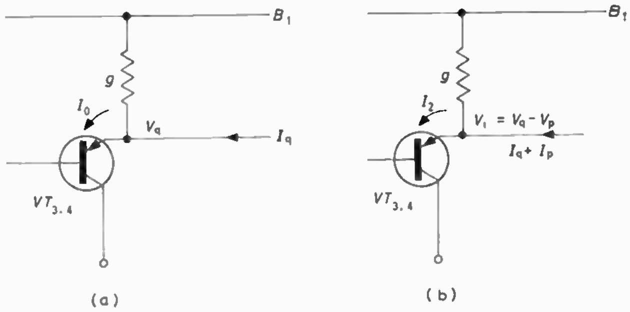


Fig. 16(a). Quiescent conditions
(b). Peak drive conditions during active half-cycle

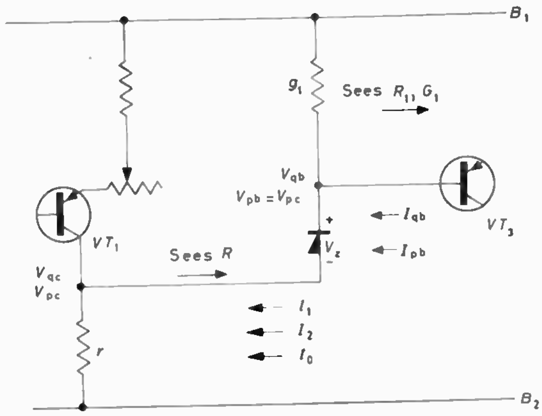


Fig. 17. Quiescent and peak conditions in the common-collector side of the phase-splitter

The Driver and Phase Splitter of the Common-collector Phase

The driver transistors must supply a distorted waveform of current to the output stage, because the latter works in class-B. But although the active part of the cycle is the part of major interest, it must be ensured that during the passive half cycle the driver is never cut off and remains in class-A. Let I_1, I_0, I_2 be the least, the quiescent, and the maximum current in the driver, VT_3, Fig. 16 I, I_1 should be much greater than the base-open leakage current:

$$I_1 \gg \beta I_{CO}$$

I_0 is chosen by the designer, choice of I_0 leads to the value of g (Fig. 16(a)).

$$I_0 - I_q = (B - V_q)g \quad \ldots\ldots\ldots \quad (42)$$

Maximum current is then given by:

$$I_2 = I_q + I_p + (B - V_1)g \quad \ldots\ldots\ldots \quad (43)$$

where $V_1 = V_q - V_p$ and I_q, I_p, V_q, V_p are conditions at the base of the output transistor VT_5 given by equations

(34) to (37). V_1 is the lowest voltage at VT_5 base; if V_2 is the highest, $V_2 = V_q + V_p$ and:

$$I_1 = (B - V_2)g \quad \ldots\ldots\ldots\ldots \quad (44)$$

When VT_5 base is at V_2, VT_5 is cut off and no current flows, so that least transistor current is equal to the current in g only.

The current in VT_3 is unequal in the maximum and minimum swings

$$I_2 - I_0 = I_p + V_p g$$
$$I_0 - I_1 = I_q - V_p g$$

knowing $V_1, V_2, V_q, V_p, I_2, I_1$ and I_0 the base conditions for VT_3 are obtained.

For the base of the common-collector driver:

$$V_{qb} = V_q - 0{\cdot}3 = -0{\cdot}6 \quad \ldots\ldots\ldots\ldots \quad (45)$$
$$V_{pb} = V_p' + V_p \quad \ldots\ldots\ldots\ldots \quad (46)$$
$$I_{qb} = I_0/\beta - I_{CO} \quad \ldots\ldots\ldots\ldots \quad (47)$$
$$I_{pb} = \frac{I_2 - I_0}{\beta} \text{ or } \frac{I_0 - I_1}{\beta} \quad \ldots\ldots\ldots \quad (48)$$

$$\text{Input resistance} = R_1 = \beta/(g + G) \quad \ldots\ldots\ldots\ldots \quad (49)$$

Here V_q, V_p refer to voltages at VT_5 base, V_p' is the voltage swing, consistent with the current swing $I_2 - I_0$, of V_{be}. It is usually sufficiently acurate to ignore V_p' compared with V_p.

G is the input conductance looking into VT_5, and is the reciprocal of equation (38).

The currents and voltages, equations (45) to (48), must be maintained by the common-collector side of the long-tailed pair, Fig. 17. If the Zener diode is to hold its specified voltage there is a minimum current I_1 which may flow in it.

$$(B_1 - V_{qb} - V_{pb})\, g_1 + (I_{qb} - I_{pb}) = I_1 \quad \ldots\ldots \quad (50)$$

V_{qb}, V_{pb}, I_{qb} are obtained from equations (45) to (47), I_{pb} from the second of equation (48).

The maximum Zener diode current is then:

$$(B_1 - V_{qb} + V_{pb})\, g_1 + I_{qb} + I_{pb} = I_2 \quad \ldots\ldots \quad (51)$$

I_{pb} is the first of equation (48), corresponding to the active half-cycle. Quiescent Zener diode current is given by:

$$(B_1 - V_{qb})\, g_1 + I_{qb} = I_0 \quad \ldots\ldots\ldots\ldots \quad (52)$$

The peak signal current in the transistor VT_1 must be:

$$I_p = (1 + (R/r))\,(I_2 - I_0) \quad \ldots\ldots\ldots \quad (53)$$

where R (Fig. 17) is the resistance, which is closely R_1, seen looking towards the diode, and r is the collector load. $R \simeq R_1$ because g_1 is usually quite small. The collector quiescent current must at least equal this. $I_q \geqslant I_p$.

Thus the current in r

$$= \frac{V_{qc} - B_2}{r} = I_q + I_0 \geqslant (1 + (R/r))\,(I_2 - I_0) + I_0$$

$$\therefore\ r \leqslant \frac{V_{qc} - B_2 - R\,(I_2 - I_0)}{I_2} \quad \ldots\ldots\ldots\ldots \quad (54)$$

This gives the maximum value for r. Finally quiescent levels give:

$$V_z = V_{qb} - V_{qc} \quad \ldots\ldots\ldots\ldots \quad (55)$$

If the pair is to be balanced equations (50) to (54) give the corresponding values of I_2, I_1, I_0, r, R and V_{pc} which must hold for the common-emitter side of the phase splitter.

The Driver, Phase Splitter and Transfer Network in the Common-Emitter Phase

VT_4 must have values of minimum current I_1 decided by the designer, of emitter load g and maximum current given by equations similar to (42) and (43). The voltage swings are much smaller than those of VT_3 so that $1/g$ and the quiescent transistor current can be smaller. Base conditions for VT_4 are:

$$V_{qb} = B - 0{\cdot}6 \quad \ldots\ldots\ldots\ldots \quad (56)$$
$$V_{pb} = V_p' + V_p \quad \ldots\ldots\ldots\ldots \quad (57)$$
$$I_{qb} = I_0/\beta - I_{CO} \quad \ldots\ldots\ldots\ldots \quad (58)$$
$$I_{pb} = I_2/\beta \quad \ldots\ldots\ldots\ldots \quad (59)$$

The 0·6 in equation (56) is twice the V_{be} for germanium transistors in quiescent conditions. V_p in equation (57) is the figure given by equation (40). The variational input impedance is:

$$R_2 = V_{pb}/I_{pb} \quad \ldots\ldots\ldots\ldots \quad (60)$$

and $R_2 \ll R_1$, because the voltage swing is reduced.

For balance of the pair VT_1 and VT_2, the Zener diode currents (Fig. 18) must be the same as those given by

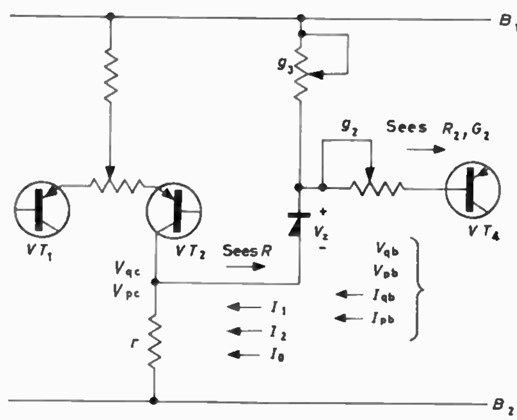


Fig. 18. Quiescent and peak quantities in the common-emitter side of the phase-splitter

equations (50) to (52) and the resistance looking into the Zener diode must equal R_1. The diode and transfer network must change the d.c. level from V_{qb} to V_{qc}, a change which approximately equals $2B$, and must alter the impedance from R_1 to R_2. Moreover the transfer network must contain adjustments sufficient to cope with the scatter of β values in VT_4. A number of networks are possible, but perhaps the simplest is that of Fig. 18.

If G_2 is the admittance looking into VT_4

$$\frac{g_2}{g_2 + G_2} = V_{pb}/V_{pc}$$

or:

$$g_2 = \frac{V_{pb}}{V_{pc} - V_{pb}}\, G_2 \quad \ldots\ldots\ldots\ldots \quad (61)$$

D.C. conditions fix the Zener diode voltage:

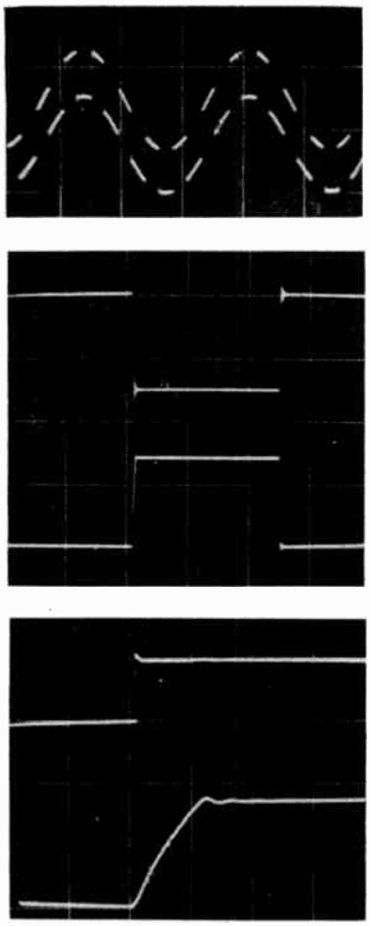
$$V_{qb} - I_{qb}/g_2 - V_{qc} = V_z \quad \ldots\ldots\ldots \quad (62)$$

D.C. balance at the node requires that:

$$(B_1 - V_z - V_{qc})\, g_3 = I_0 - I_{bq} \quad \ldots\ldots\ldots \quad (63)$$

I_0 is given by equation (52).

It has already been pointed out that the current demands of VT_4 are smaller than those of VT_3, so that the condition for minimum diode current can be satisfied. There is an a.c. mismatch with this transfer network, because although the voltage swings V_{pc} at the node are the same for both VT_3 and VT_4 the current swing required at VT_4 base may be less. The quiescent current balance is ensured by choosing g_3 correctly from equation (63), but g_3 goes to $+B_1$



(a) 100c/s sine plus 600c/s square 7A/division

(b) 100c/s square 2V/division

(c) Response 7A/division

(d) Unit step 2V/division

(e) Response 7A/division

Time scale 65μS/division

Fig. 19. Response of Class-B amplifier to various waveforms

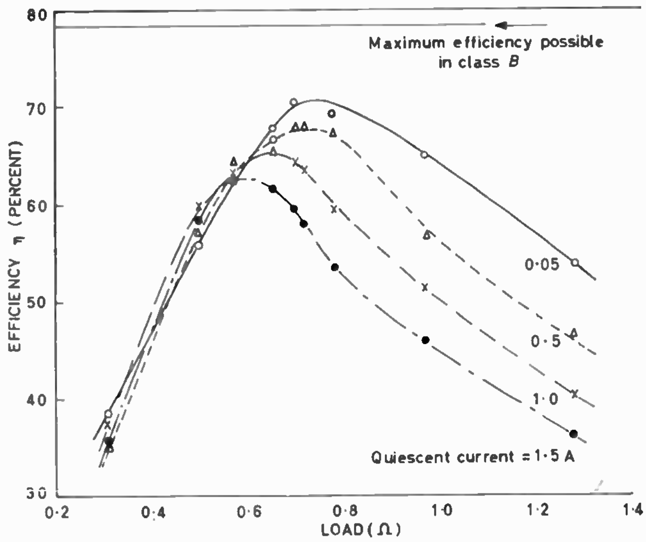


Fig. 20. Variation of efficiency with load

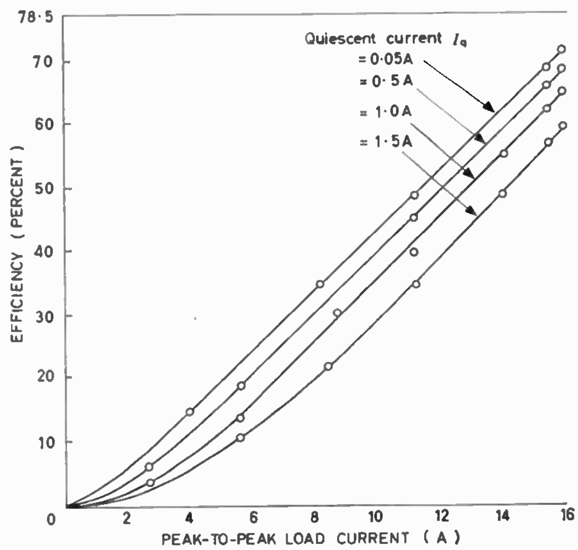


Fig. 21. Variation of efficiency with output current

and is generally too small to ensure an a.c. match simultaneously. The transfer network can be modified by including a conductor to earth and it is then possible to satisfy the a.c. match requirement exactly. However, it is not worthwhile to do so because the setting-up procedure, which allows adjustment to the scatter of β values, is then harder. The a.c. mismatch of the simpler circuit is small if the effects of the differing emitter loads of VT_3 and VT_4 are ignored, and if g_1 and g_3 (Figs. 17 and 18) are small, there is no mismatch.

On the common-emitter side the variational impedance at VT_6 base is approximately:

$$\beta\, V_{pb}/I_p,$$

at VT_4 base it is $\beta^2\, V_{pb}/I_p$,

and at VT_2 collector it is:

$$\frac{V_{pc}}{V_{pb}} \cdot (\beta^2\, V_{pb}/I_p) = V_{pc}\,\beta^2/I_p$$

TABLE 3

QUANTITY	ESTIMATED OR DESIGNED VALUE (V)	ACTUAL VALUE (V)
VT_6 base $= VT_4$ emitter	+8·3	+8·3
VT_5 base $= VT_3$ emitter	−0·3	−0·44
VT_4 base	+7·9	+8·1
VT_3 base	−0·6	−0·63
VT_2 collector	−9·3	−11·2
VT_1 collector	−9·3	−10·3
VT_1, VT_2 emitter	0	−0·18
Junction of g_2 & g_3	+5·6	+4·5
Zener voltage CE phase	15·2	15·5
Zener voltage CC phase	8·7	9·7

Here the value of V_{pb} from equation (40) is used throughout.

On the common collector side the impedances are approximately at VT_5 base:

$$\beta\, V_{pc}/I_p$$

at VT_3 base: $\beta^2\, V_{pc}/I_p$ = impedance at VT_1 collector.

Consider next the effect of ignorance of β value. On the common collector side it is uncritical, because of the common collector connexions. On the common-emitter

TABLE 4

Current gain A_i (12A p-p)	$-1{\cdot}3\times10^5$
Current gain A_i low level	$-1{\cdot}9\times10^5$
Voltage gain A_v (4A p-p)	−41
Voltage gain A_v (0·4A p-p)	−75
Input admittance y (12A p-p)	2×10^{-3} mho
Input admittance, low level	$8{\cdot}3\times10^{-4}$ mho
Output admittance Y (12A p-p)	1·4 mho
y with feedback, low level	0·39 mho
Y with feedback	50 mho

side, however, the βs of both VT_6 and VT_4 strongly affect the currents required at the base of VT_4. These affect the values of g_2 and, slightly less strongly, g_3, from equations (60) and (62). Therefore g_2 and g_3 are made variable. In setting up one sets g_2 to its central value and then adjusts g_3 to give the correct V_{qb}, one then puts on a.c. signal and adjusts g_2 to give the correct ratio of V_{pb} to V_{pc}. This will usually necessitate readjustment of g_3, but the process converges.

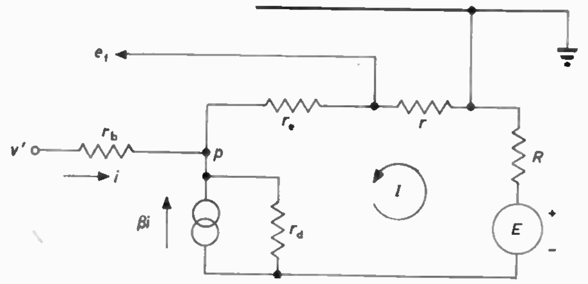


Fig. 22. Part of constant current source including equivalent common emitter T of the series transistor. (See also Fig. 11(b))

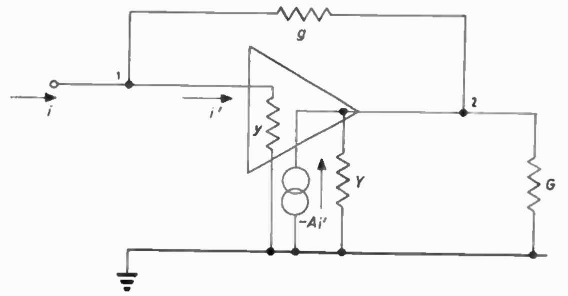


Fig. 23. Current amplifier, gain —A, input admittance y, output admittance Y. G is the load conductance, g the feedback conductance

Performance of a 20A Peak-to-Peak Amplifier and Conclusion

The approximate method just given was used to design a 20A d.c. amplifier working into a 0·7Ω load using ±8·6V main supply voltage. The maximum swing obtainable without distortion was 16A peak-to-peak on this load, but swings of 21A, peak-to-peak, were obtainable on 0·25Ω loads. The power transistors were not selected, indeed means of measuring their βs at appropriate current levels were not available. The design method was successful in that no components needed changing. The values of the high-frequency cut-off capacitors were found by experiment as usual, and the large-signal performance was improved at higher frequencies by adding series resistors (the 0·05μF in series with 330Ω, or 0·01μF in series with 1 500Ω in Fig. 14). The response was accurately flat for small signals (15A, peak-to-peak) from d.c. up to 20kc/s, then rose to a maximum of 1·5× midband gain at 40kc/s, and was 3dB down at 60kc/s. At high currents (14A peak-to-peak) waveform distortion appeared at 1·5kc/s.

Table 3 shows the estimated and actual voltages at the nodes of the circuti. Agreement is adequate. Table 4 summarizes the main properties of the amplifier, using the notation of Fig. 23.

The current gain relation, equation (71), was accurately obeyed, as shown by the measurement of current gain with feedback A_f for various loads at high level ($9V_{pk\text{-}pk}$ output), Table 5.

TABLE 5

Measured current gain A_f	75	124	181	275	235	372	410
Ratio G/g	76·5	127	184	281	239	381	417

Smaller current gains than 75 could not be obtained without oscillation; however in the latter case the ratio of current gain with feedback to gain without was about 6×10^{-4}, a high degree of feedback.

The measured input and output admittance were much smaller than the values predicted by equations (78) and (79).

The amplifier may also be used as a *voltage* amplifier, though with loss of precision, by applying a voltage to VT_2 base. 100 per cent voltage feedback is then provided via Y_f, the 180Ω feedback resistor for operation as current amplifier, which joins the hot end of the load to the other, difference, terminal of the pair VT_1 and VT_2. Fig. 19(a) shows the result of applying a current signal to VT_1 base and simultaneously a voltage signal to VT_2 base. The current wave is a sine wave at 100c/s, the voltage wave a square wave at 600c/s. Fig. 19(b-e) shows the response to a current square wave and to a current unit step.

The amplifier is an efficient power convertor with correct value of load, when a maximum efficiency of 72 per cent can be obtained before the onset of distortion. Fig. 20 shows the variation of efficiency with load for different values of quiescent current I_q in the output transistors. It is possible, because of the high degree of feedback, to operate in true class-B, with $I_q < 50$mA for either phase. This explains the efficiencies of 70 per cent, which compare with the 78·5 per cent theoretically possible.

Fig. 21 shows the variation of efficiency with output current. The curves stop at the point where clipping of the waveform begins. For true class-B operation the curves should be straight lines, and this is closely the result when $I_q < 50$mA.

One may conclude that the design method given is adequate to produce efficient and precise high level d.c. amplifiers, with relatively little effort both in design calculation and in laboratory work.

Acknowledgments

Thanks are due to the following bodies: Messrs. Metals Research, Cambridge, and the B.W.R.A. Abington for laboratory assistance; the A.W.R.E., Aldermaston, and D.S.I.R. for components. Mr. C. Thomas of B.W.R.A. built and tested the low drift 'straight' voltage amplifier. At the Engineering Laboratory, Cambridge, Messrs. H. L. Lowe and A. Balodis built the current measuring and precision current source amplifiers; Mr. J. H. Hinton put the latter into operation and tested it; and Mr. K. Y. Ketteridge helped with testing the class-B amplifier.

Appendix

(1) Derivation of the Ripple Relation for a Constant Current Source

The system is shown in Fig. 11(b), and Fig. 22 shows the output transistor in its common-emitter T equivalent.

Approximation one is made that, as far as the flow of I is concerned the shunting of $(r_e + r)$ by r_b is neglible (r_b goes to system earth via the low output impedance of the voltage amplifier). Then the current in r:

$$I = \frac{E}{R+r'+r_d} - \beta i \frac{r_d}{R+r'+r_d} - i \frac{r_d+R}{R+r'+r_d} \qquad \ldots\ldots\ldots\ldots (64)$$

where $r' = (r_e + r)$.

Approximation two is that the last term, which is $1/\beta$ of the second term, is negligible. Let $z = R+r'+r_d$.

The voltage at node p (Fig. 22) is:

$$V_p = -Ir'$$

for the base branch:

$$\frac{v' - v_p}{r_b} = i$$

$$\therefore\ i = \frac{v' - (\beta i r_d - E)/r'z}{r_b}$$

$$\therefore\ I = \frac{v + Er'/z}{r_b + r'\beta r_d/z} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (65)$$

Insert equation (65) in the second term of equation (64), and with approximation two:

$$I = (E/z) - (\beta r_d/z) \cdot \frac{v' + Er'/z}{r_b + r'\beta r_d/z}$$

$$\therefore\ I = \frac{r_b E - \beta r_d v'}{z r_b + r'\beta r_d} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (66)$$

But:

$$v' = (e - e_t)A_v = [e + IrA_v] \quad \ldots\ldots\ldots\ldots \quad (67)$$

write equation (66) as $I \equiv \lambda E - \mu v'$

$$= \lambda E - \mu e A_v - \mu I r A_v$$

$$\therefore\ I = \frac{\lambda E - \mu e A_v}{1 + \mu r A_v}$$

If:

$$\mu r A_v \gg 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (68)$$

this becomes:

$$I = (\lambda E/\mu r A_v) - (e/r)$$

but:

$$\lambda/\mu = r_b/\beta r_d$$

hence:

$$I = -(e/r) + (r_b/\beta r_d)(E/A_v r) \quad \ldots\ldots\ldots \quad (69)$$

differentiating this gives equation (33).

The inequality equation (68) is easily satisfied. Inserting the value of μ and z:

$$\frac{\beta r_d}{(r_d + r' + R)\, r_b + r'\beta r_d} \cdot A_v r \gg 1$$

$$\therefore\ \frac{A_v r}{r_b/\beta + r'} \gg 1 \quad \text{or} \quad \frac{A_v r}{r_b/\beta + r_e + r} \gg 1$$

r, r_b/β, r_e are of the same order, and A_v may be 1 000. Hence the left-side of equation (68) may be 300.

(2) Derivation of Current Gain, Input and Output Admittances for the General Current Amplifier with Feedback

With the notation of Fig. 23, nodal equations are:

$$i = v_1(g + y) - v_2 g$$

$$-Ai' = -v_1 g + v_2 G_{22} \text{ where } G_{22} = g + G + Y$$

also:

$$i' = v_1 y$$

Therefore:

$$v_1 = -v_2\,(G_{22}/(Ay - g)) \quad \ldots\ldots\ldots\ldots \quad (70)$$

and:

$$i = -v_2\left[\frac{(g + y)\,G_{22}}{Ay - g} + g\right] \quad \ldots\ldots \quad (71)$$

Hence the current gain with feedback:

$$A_t = i_o/i = v_2 G/i = -\frac{G}{\dfrac{(g+y)G_{22}}{Ay-g} + g}$$

which may be written in the simpler form.

$$A_t = -\frac{G/g}{1 + G_{22}/Q} \quad \ldots\ldots\ldots\ldots\ldots \quad (72)$$

where the conductance Q is large and is an abbreviation for

$$Q = \frac{g(A - g/y)}{1 + g/y} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (73)$$

The input admittance felt by the input current equals

$$y_{in} = i/v_1 = -v_2\left[\frac{(g + y)\,G_{22}}{Ay - g} + g\right] \Big/ \left[-v_2\frac{G_{22}}{Ay - g}\right]$$

$$= y(1 + (g/y))\,(1 + Q/G_{22}) \quad \ldots\ldots\ldots\ldots \quad (74)$$

It is therefore much increased by feedback, and the system approaches a perfect current amplifier more nearly.

The output admittance is the quotient of short-circuit output current by open-circuit output voltage:

$$y_{out} = i_{o(sc)}/v_{2(oc)}$$

On short-circuit, $G = \infty$, i divides between y and g giving $i' = i \cdot \dfrac{y}{g + y}$

$$\therefore\ i_{o(sc)} = -Ai' + i\,\frac{g}{g + y}$$

$$= -i\,\frac{(Ay - g)}{g + y}$$

On open-circuit $G_{22} = g + Y$,

$$v_{2(oc)} = \frac{-i}{\dfrac{(g + y)(g + Y)}{Ay - g} + g}$$

Therefore:

$$Y_{out} = Y\left(1 + \frac{g + Q}{Y}\right) \quad \ldots\ldots\ldots\ldots\ldots \quad (75)$$

Thus the feedback greatly increases the output admittance, to give, paradoxically, a characteristic of a *voltage* source.

In the limiting case when A, the open-current gain, becomes very large:

$$Q = \frac{gA}{1 + g/y} \text{ mho} \quad \ldots\ldots\ldots\ldots\ldots \quad (76)$$

$$A_t = -G/g \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (77)$$

$$Y_{in} = ygA/G_{22} \text{ mho} \quad \ldots\ldots\ldots\ldots \quad (78)$$

$$Y_{out} = \frac{gA}{1 + g/y} \text{ mho} = Q \quad \ldots\ldots\ldots \quad (79)$$

In the case of the high capacity d.c. amplifier, working with large signals:

$A = 130\,000$
$g = 5{\cdot}53 \times 10^{-3}$ mho
$y = 2 \times 10^{-3}$ mho
$G = 1{\cdot}43$ mho
$Y = 1{\cdot}45$ mho
$Q = 191$ mho $G_{22}/Q = 0{\cdot}015$ $G/g = 259$
$A_t = 255$, $Y_{in} = 1{\cdot}0$ mho, $Y_{out} = 191$ mho

Corrections to Part 1

It is regretted that several errors occurred in Part 1. On page 456 the second line in column two should read '.... I_{oc} must satisfy equation (17). On page 458 equation (28) should read

$$v/i = -(1/BG)\left[1 - \frac{G + g}{A_v BG}\right].$$

On page 459 the third line below equation (31) should read '.... parallel with g_1....'.

In Fig. 1 the bottom end of capacitor C should be connected to the junction of VT_3 collector and VT_5 base. In Fig. 9 the collector of the output transistor should be connected to the B_2 line.

A Voltage Stabilizer for Cold-Cathode Tube Circuits

By A. J. Oxley*, M.A., Ph.D.

A single cold-cathode trigger tube is used in the stabilizing circuits described, for outputs of 150 to 1000V. One version intended for supplying Dekatrons and trigger tubes gave up to 20mA at 470V, with an output impedance of 250Ω and stabilization ratio of 35:1.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

MOST Dekatron and cold-cathode trigger tube circuits make light demands on a power supply. The current consumed is small and some mains ripple is tolerable, so gas-filled diode stabilizers can be used in parallel with the load. These are available for a small variety of voltages (e.g. 90V, 108V, 150V) and can be connected in series, but a more flexible system using a trigger tube as a stabilizer has been described[1]. It is shown in Fig. 1.

Each positive swing at the input raises *A* and *B* until the trigger tube conducts when, assuming the running voltage *v′* is less than the level at *B*, diode *D* cuts off and

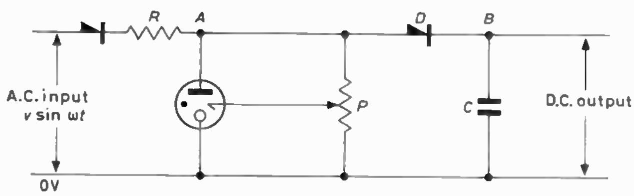


Fig. 1. The original stabilizer (earthed cathode)

A is held at *v′* until the input falls too low to maintain the discharge. The maximum level reached at *B*, when the tube fires, depends on the setting of potentiometer *P* but within limits is independent of the load current or input voltage *v*. Resistor *R* must be small enough for the desired level at *B* to be reached for the load current concerned, but not so small that the rating of the trigger tube is exceeded (in some cases a resistor between *A* and anode may help). So the circuit will stabilize the peak voltage at *B*, which will thereafter fall as the load current discharges capacitor *C*, until it is recharged at the next cycle.

It was pointed out[1] that two such circuits could be stacked together to obtain a stabilized output from a voltage doubler. Results are reported here for a new circuit (Fig. 2) in which the output of a voltage doubler is stabilized by a single trigger tube. This is made possible by earthing the anode of the trigger tube instead of its cathode, a change which also makes extra smoothing of the output possible. A suitable tube is the Ericsson type GPE120T now available; the circuit was used for supplying Dekatrons and other trigger tubes.

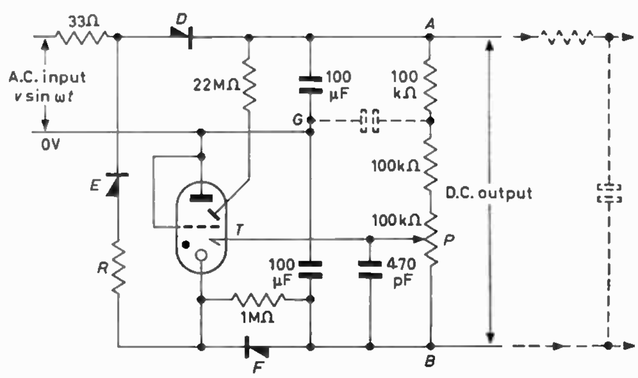
The GPE120T has a primer anode (connected via 22MΩ to *A*) to make triggering more reliable and a shield, which can be connected to the anode as in Fig. 2 (though a resistor between them might be an advantage, as the current would then be more equally shared with the anode). The 1MΩ resistor maintains the priming discharge during the positive half-cycle of the input, and allows a diode of lower p.i.v. rating to be used at *F*. The 470pF capacitor has little effect on stability, at least with a relatively low resistance divider chain feeding the trigger, and there is no marked difference if it connects *T* to the cathode or *G*, or is omitted altogether. Potentiometer *P* adjusts the output voltage.

It will be seen that the peak voltage at *A* is not stabilized, being allowed to fluctuate with fluctuations of the input. As *B* is carried downwards by conduction through diodes *E* and *F* the potential of the trigger with respect to the cathode rises until the tube fires, raising the cathode to −105V (the running potential) and cutting off diode *F*. This point will be reached earlier if the input voltage increases, but for all reasonable inputs and loads should occur at the same trigger-cathode potential, and in consequence at the same *A-B* potential.

Thus the one tube stabilizes the total voltage between *A* and *B*, and a similar system as in Fig. 3 could be used with a voltage quadrupler if required. Expressed as a percentage of the output, the instability due to slight fluctuations in firing level of a tube is about the same for each circuit, or for a stacked two (or four) tube circuit, because the instabilities for several tubes are additive. The tube is liable to break down with 250V between shield and cathode regardless of the trigger potential, so the maximum output in Fig. 2 is (v + 250) volts, though up to (v + 400) is possible if the shield is on a divider between anode and cathode. Precautions must be taken above (v + 225) volts to avoid current flowing to the trigger when the tube glows. The minimum output is (v + 105) volts, fixed by the running potential, but a resistor could be inserted next to *D* to achieve a lower output. Unless necessary, though, all the resistance (except for the surge-limiting 33Ω) should be put at *R* to reduce the loading on the tube.

Some characteristics of the circuit were determined as follows:

Fig. 2. Earthed-anode stabilizer applied to voltage doubler



* *University of Oxford, Department of Nuclear Physics.*

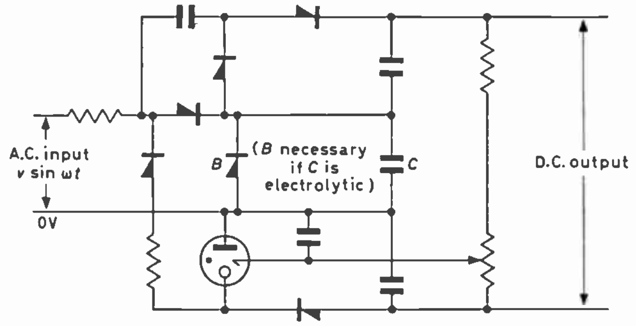


Fig. 3. Earthed-anode stabilizer for voltage quadrupler

(a) With $R = 3k\Omega$, external load $82k\Omega$, input 240V a.c., P was set to give 470V output.

(1) Stabilization was effective over the input range 210V to 255V, which caused an output variation of about 2·5V, a stabilization ratio of nearly 40:1.

(2) After resetting to (a) the load was removed, causing a rise of about 1·5V, equivalent to an output impedance of some 250Ω. The maximum stabilized load current was 10mA.

(3) Again resetting to (a), P could be adjusted to give stabilized outputs in the range 435V to 495V.

(4) With the conditions of (a), the circuit was left running four days. With high stability resistors in the trigger divider chain, the output was within a range of 1V whenever checked.

Other tests made on different versions of the circuit indicated that a small capacitor between trigger and cathode generally increases the maximum stabilized output voltage attainable, although there is no marked effect on stability at lower voltages. The tube draws a peak current of about 80mA and a mean of 15mA in (a), and is rated for a peak/mean of 125/25mA for long life or 250/50 for reduced life. So it is permissible to reduce R and tests were done with $1{\cdot}5k\Omega$ as follows:

(b) Input 240V a.c., load $27k\Omega$, P set for 470V output.

(1) Input range 215V to 260V.

(2) Stabilization effective for loads of 0 to 20mA.

(3) Output range 435V to 510V,

resetting to (b) before each.

Overall, the performance is comparable with that of diode stabilizer tubes, with the advantage of adjustable output, but more ripple (in Fig. 2 the output has a sawtooth waveform of 100c/s and 0·2V/mA output current, smoothing being entirely due to the capacitors). Extra resistance-capacitance smoothing can be introduced as indicated by the broken lines in Fig. 2, in which case the top of the trigger divider must go to the positive output terminal (instead of to A) to stabilize against load variations: smoothing requires an earthed-anode stabilizer, in single or voltage-multiplying circuits. There are also some random jumps of less than 0·5V at the output. The trigger tube circuit is often more efficient, and where it is adequate, its low cost is attractive.

REFERENCE

1. *Electronic Appl.* *21*, 25 (1960-61).

A Data Transmission System for Marshalling Yards

Modern hump marshalling yards use a variety of automatically controlled devices to enable wagons to be sorted quickly and efficiently. The routing and speed of wagons from reception to sorting sidings and the order of sorting can be electronically controlled from a central control tower. For these controls to route the maximum number of wagons possible it is essential for them to obtain quickly accurate information from the reception sidings, giving the details and number of wagons and the sorting sidings to which each wagon is to be routed. This information is required in a printed form for the operators to understand and in a coded form or punched paper tape for instructing the electronic equipment controlling the points, retarders, etc.

The distance from the reception sidings to the control tower can be up to 1 000 yards and many systems have been tried to transmit accurately the information required. Systems utilizing pneumatic tubes, telephones or radio links, which are at present in use, have two disadvantages. Errors can occur through ambiguity in understanding at the control tower the written or spoken words and even when these are understood a further operation is necessary to encode this information into punched paper tape for instructing the electronic controls. Errors can occur in translating unless some form of verification is carried out.

A new approach has now been made, which overcomes these problems, in a data transmission system which uses standard add/listing machines connected to tape punches and tape readers. Such a system has recently been ordered from Addo Ltd for one of British Railways' new marshalling yards at Tinsley, near Sheffield.

Trains arriving at the reception sidings are first uncoupled into 'cuts.' Each 'cut' consists of one or more wagons which are for the same destination. The object of a marshalling yard is to transfer all the cuts for a particular destination into one sorting siding so that they can be made into a train and sent via a main line route to their destination. When a train is uncoupled a 'tally man' checks each cut and marks on a cut card the sorting siding to which each cut is to be transferred, the number of wagons in each cut and the type of wagon. A cut card is made out for each train arriving in the reception sidings and the information on the cut card then has to be transmitted to the control tower.

In a typical arrangement of transmitters and receivers there are three transmitting stations in the reception sidings and each consists of an Addo-X single register, three column, shuttle carriage machine, mounted on a table in a kiosk. The machines are fitted with output units and associate equipment to enable information which is printed to be transmitted electrically over a cable to the receiver. The receiver consists of an identical machine fitted with both an input unit, which enables it to receive and print the data from each transmitter, and an output unit which is connected to an Addo tape punch which ensures that all the information printed is automatically encoded into punched paper tape. The receiver is housed in the control tower and is sound proofed against noise.

When the tally man has completed a cut card he proceeds to the nearest transmitter kiosk to transfer his information into the Addo machine. Each transmitter has an indicating lamp which when lit allows the machine to be operated. The lamp is not lit and the machine is locked against operation whenever another transmitter is in use. The Addo machines have a simplified, symmetrical keyboard consisting of 10 digit keys, 0 to 9, and three motor keys. The tally man enters his cut card information into the Addo using one motor key for the heading data, i.e. train no. and date, another motor key for the cut data and when all the data is transferred from the cut card he operates the third motor key to signify the end of message.

All the information entered into the Addo is printed in the same form as on the cut card and therefore can easily be verified by the tally man as he is entering the information into the transmitting machine. The receiving machine also prints the information in the same form and automatically produces a punched paper tape which contains both the transmitted data and the control codes required by the electronic control equipment. The receiving and transmitting machines are both programmed so that the number of wagons in each cut is entered in the register of the machines. When the third motor key is operated at the end of each cut card the total number of wagons for that train is printed on both transmitting and receiving machines. Also the tally roll on both machines is automatically spaced up for about one inch to enable it to be removed from the machines.

Voltage and Current Substandards

By M. Pacak*

An electronic voltage or current stabilizer is described, which is equipped with an efficient amplifier and several other features ensuring exact and reproducible setting of the stabilized output quantity, determined almost exclusively by a reference divider, and a reference voltage. A stepwise or continual adjustability over the range of about 1 000V or 1A to zero or below (reverse polarity) may be attained. The stability and setting precision of 10^{-4} or better enable the instrument to be used as a voltage or current scaler for diagram plotting, instrument calibrating, etc.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

APART from electronic d.c. voltage or current stabilizers characterized by simple classical design and especially by continuous ajustability of the output value in a rather limited range, stabilizers of a new type have recently appeared with the output value adjustable in a relatively wide range, generally over several decades. The main difference compared with the older type is stepwise switching of the desired output value by means of a decade or other dividing system, and high relative accuracy of individual step values. These stabilizers have so far been built as voltage supplies only, but it is possible to change the basic circuit easily to a constant current source. If sufficient precision can be attained, say 0·1 per cent or better, then an evident application of these supplies is the calibrating of d.c. voltmeters or amperemeters. Therefore the denomination 'substandards' is legitimate for stabilizers of this kind.

Several arrangements of substandard circuits have been examined by the author and the following basic facts have been established. The general arrangement is shown in Fig. 1. The same concept may be used for voltage (a) and current (b) substandards, respectively, the only difference being in the reference circuit. The precision and stability of the reference resistors R_1, R_2 or R_I, and of voltage source V_r present the limits of substandard quality. Therefore wire-wound resistors with the precision required from the substandard or better must be used here, and a stable dry cell or battery is the most suitable reference source.

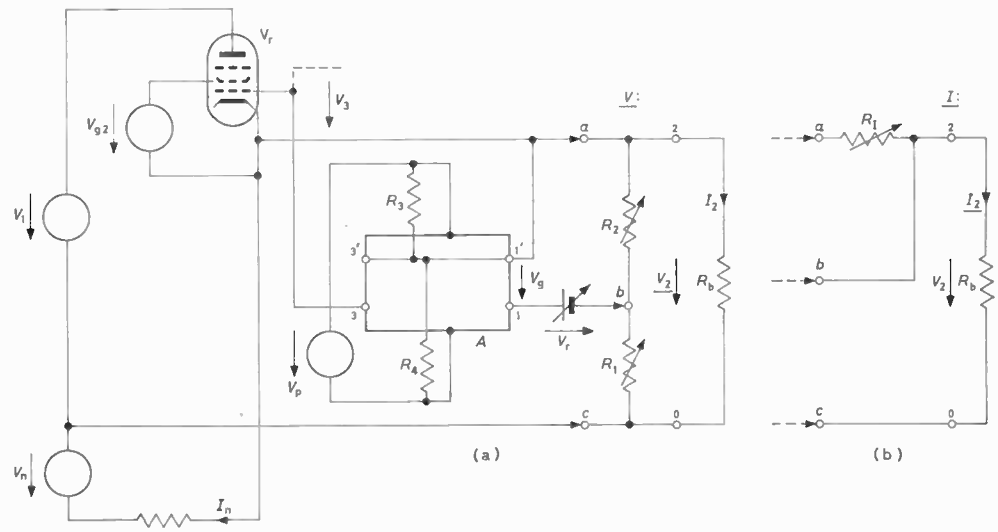


Fig. 1. General concept of substandard-type electronic d.c. voltage (a) or current (b) stabilizer with precautions made for wide range and precise output value adjustment.
The arrangement of auxiliary sources V_n and V_{g2} may be seen here

The following formulae for the output quantities may be deduced from Fig. 1:

$$V_2 = (V_r + V_g)\,(1 + (R_1/R_2)\,);\ I_2 = (V_r + V_g)\,(1/R_I) \quad \text{(1a; 1b)}$$

Should the stabilized value V_2 or I_2 be defined by the respective reference circuit parameters only, the amplifier input voltage V_g must be reduced to practically zero. As far as its constant component is concerned, this may be attained by using a symmetrical amplifier input. The signal and noise component of V_g may be reduced by ensuring high amplification (10^4 to 10^6 in general) and by choosing a low noise input valve and suitable circuit arrangement.

Assuming the input voltage V_g zero, the relations (1ab) reduce to

$$V_2 = V_r(1 + (R_1/R_2));\ I_2 = V_r(1/R_I) \quad \text{(2a; 2b)}$$

Obviously, the output values V_2 or I_2 may be adjusted by changing either the reference circuit resistivities, or the reference voltage. The lowest V_2 attainable equals V_r, which therefore has to be made equal to the lowest value of V_2 required, say 1V. If still lower values of V_2 are desired, V_r must be varied proportionally. In this way the output value may be changed over from an arbitrary starting point to zero or even to negative polarity, if auxiliary current I_n is driven through the regulator valve. If the output value is adjusted by means of changing resistors R_1, R_2 or R_I, the reference voltage V_r remaining constant, the relative value of output error or fluctuation is also practically constant. If, on the contrary, a change of reference voltage V_r is used for adjusting the output quantity, the absolute value of output error is constant and is equal to the equivalent amplifier input noise signal multiplied by the ratio $k = (R_1 + R_2)/R_2$.

As the output voltage of a substandard stabilizer varies widely, the amplifier anode circuits must be supplied by a separate source V_p. Large amplification and wide output dynamic range are necessary. Therefore the symmetrical amplifier is the best arrangement, ensuring also the compensation of most disturbing effects and eliminating interstage couplings caused by the V_p supply internal resistance.

* *Inst. Phys. Chemistry, Czechoslovak Academy of Sciences, Prague.*

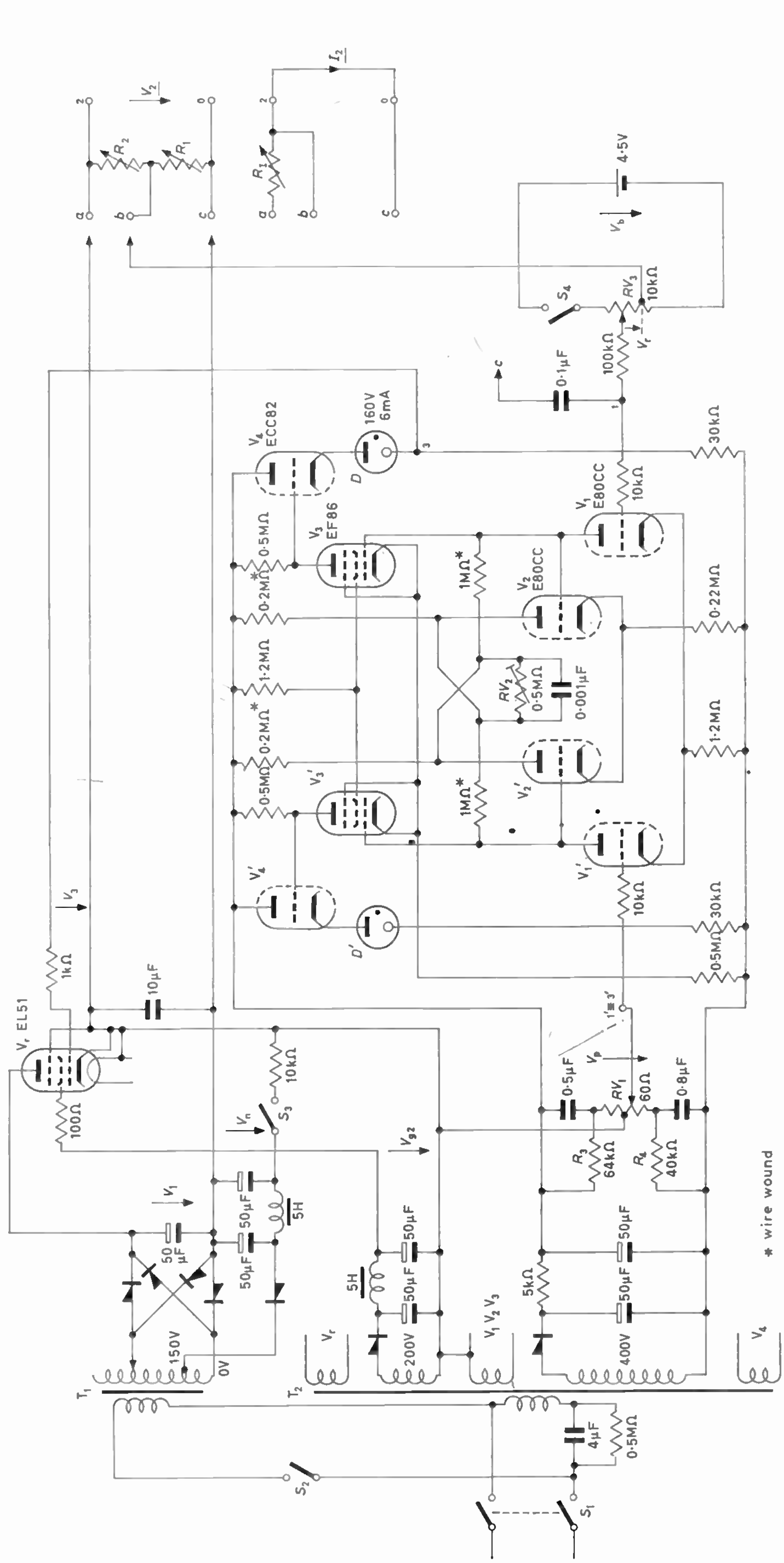


Fig. 2. Circuit diagram of a substandard stabilizer for d.c. voltage or current. Large amplification and dynamic range, low amplifier output resistance must be achieved. Consistently symmetrical amplifier circuit simplifies the supplying problems and compensates the most disturbing effects

For a symmetrical amplifier the voltage V_p need not be stabilized.

Two other auxiliary supplies are necessary for ensuring the proper substandard properties. Ballast current source V_n, driving a defined value I_n through the regulator valve V_r even if the output current I_2 falls to zero. Regulator screen grid supply V_{g2}, enabling valve V_r to function as a pentode. None of these supplies need be stabilized. A pentode regulator valve (i.e. with independent screen grid supply) permits a substantially greater V_2 range, the control signal V_3 remaining smaller than if, as usual, the regulator screen grid is connected to its anode. This is true even if V_{g2} varies over some 10 per cent, this variation being still of about one order smaller than regulator anode-cathode voltage variation, which is at least the full range of V_2 required, i.e. several hundred volts or more. In this way 1000V of V_2 range may easily be attained, without switching the V_1 supply, presuming adequate regulator power capacity. The effect of V_n or V_{g2} fluctuations is negligible, as they enter the feedback loop at a high signal level and the feedback amplification is large.

Practical applications of the statements given above are embodied in the universal (i.e. voltage- or current-) substandard, Fig. 2, corresponding basically to Fig. 1. The main transformer voltage may be switched to the appropriate value with respect of the largest V_2 required. Polarization current I_n may be switched off by S_3, when it is not necessary. Anode supply V_p is connected to the main circuit by means of a voltage divider R_3, R_4 and tapped potentiometer RV_1, enabling fine compensation of V_p fluctuations.

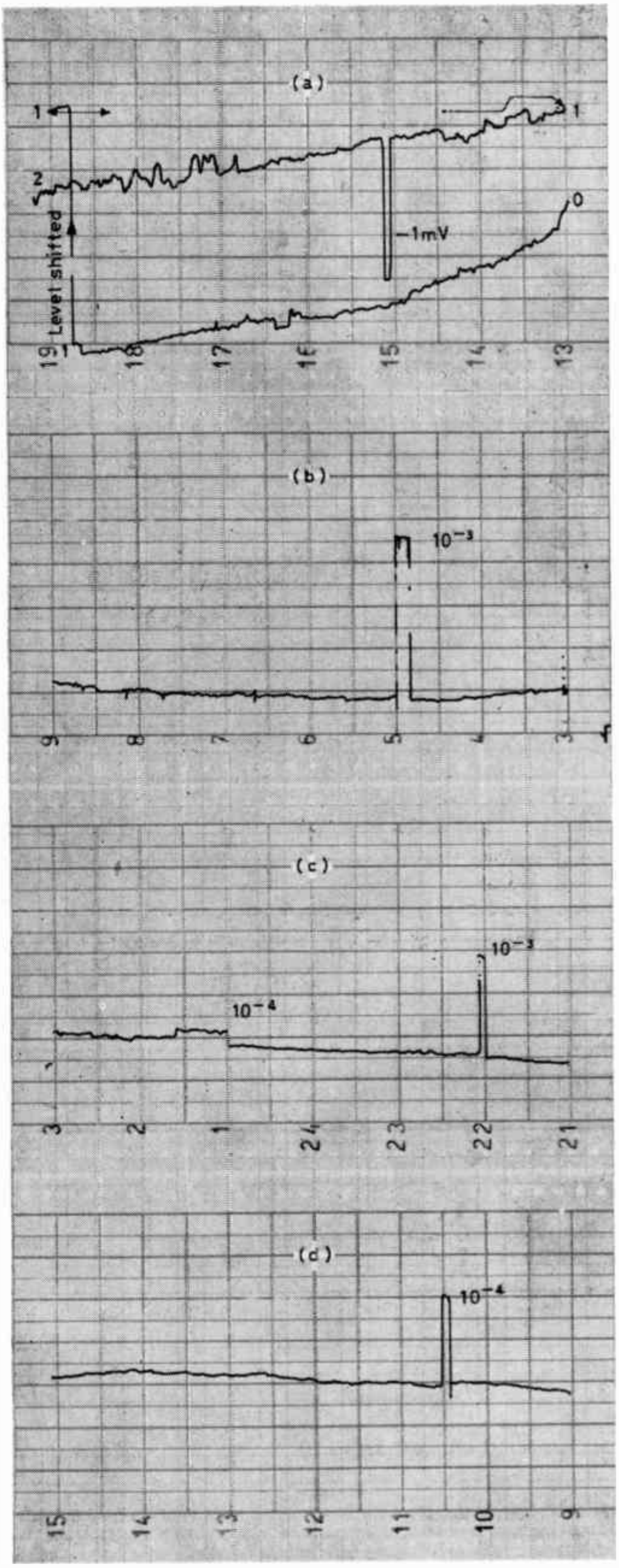


Fig. 3. One hour stability tests of substandard stabilizer in Fig. 2, taken with chopper-type recording millivoltmeter

Transformer T_2 supplying all voltages with exception of V_1 and V_n is built as a simple (non-compensated) magnetic stabilizer, reducing the relative mains voltage fluctuations by the factor of 7. Therefore the instrument may operate correctly for the mains fluctuating between 190 and 250V. The output voltage may be adjusted by means of operations discussed above between zero (or even a small negative value, say 10 per cent of the maximal V_2 value preset by reference resistors) and several hundred volts, according to preset value of V_1. The output current may be changed similarly from zero or from a small negative value to several hundred milliamperes according to rectifier and regulator current capacity. With the arrangement applied here the output values may safely cover the range of 1000V and 1A.

The symmetrically arranged amplifier consists of four stages. Double triode V_1, V_1' operates as a common cathode input stage. Its input current is practically zero, which is necessary because the reference circuit resistance may vary substantially when switching for individual steps. This is because at first the following stage V_2, V_2' has positive feedback, its input resistance being in general rather small. If the critical condition is adjusted by variable resistor RV_2, infinite amplification properties are attained[1]. The low-power pentodes V_3, V_3' provide the main amplification and the dynamic range necessary. The first three stages use common cathode resistors securing simple interstage coupling, good signal symmetry and reduced heating change effects.

The double triode V_4, V_4' operates as a cathode-follower pair with neon tubes D, D' as coupling elements for d.c. level transposition and low output resistance. This is important especially when several parallel connected regulator valves are to be controlled together (for larger current capacity). The positive feedback stage proposed here is rather advantageous, as it requires only two pairs of precise and stable resistors marked* (Alma wire-wound resistors are used in the instrument examined below).

The adjustable reference voltage supply as shown in Fig. 2 makes it possible to adjust V_r from the full positive value through zero to 10 per cent negative. For the voltage substandard two resistance decades or precision step voltage divider may be used as R_1, R_2. In the current substandard, a resistance decade or better a conductance decade with a linear scale of the values $1/R$ may be used, the latter enabling direct proportional current steps to be set.

In a substandard-type stabilizer constructed essentially in accordance with Fig. 2 the following results were obtained. Absolute short-time zero voltage stability of 0·2mV peak-to-peak, and drift value of 0·6mV/h were realized (Fig. 3(a)). When using a 4·5V dry battery as reference voltage, ballasted by approximately 1mA, relative short-time variations of $0{\cdot}5 \times 10^{-4}$ and relative drift 2×10^{-4}/h were registered (Fig. 3(b), (c)). When using a reference voltage of 50V (which of course increases the lowest V_2 attainable to the same value), the relative variations and drift of about one order smaller, $0{\cdot}5 \times 10^{-5}$ and 3×10^{-5} (Fig. 3(d)). Consequently, when using a larger value of V_r, the favourable properties of the substandard concept result in high-grade stabilization, reaching probably the best values attainable with direct-coupled amplifiers. The current version of this instrument offered in general similar results. When changing the mains voltage stepwise from 190 to 250V, the output value changed by about 3×10^{-4} with 4·5V reference source. Residual a.c. output voltage under 1mV was obtained for arbitrary output value. When switching the output values over the whole range, the stationary value of amplifier input signal did not exceed 0·2mV. Consequently, with a reference voltage of 1V the relative error value of individual steps does not exceed the tolerance value of reference divider or resistor by more than 2×10^{-4}, satisfactory substandard precision being verified in this way.

For lower precision and power requirements the amplifier and supply circuits could be simplified in some respects. The same design principles may be used for transistor sub-standards with exception of the input circuit, which needs to be designed differently. The transistor stabilizer is convenient for output values of several tens of volts and/or several amperes.

REFERENCE

1. PACAK, M. A Voltage Stabilizer Principle. *Electronic Engng.* 30 210 (1958).

Communications in a Steel Mill

A new £32M plant comprising a steel-making department, a bloom, slab, and billet mill, and a continuous narrow hot strip mill, engaged in the production of steel strip has just been completed at the works of The Park Gate Iron & Steel Co. Ltd, one of the T.I. group of companies. A feature of the new plant is the use of three English Electric-Leo computers to assist in production planning for the whole works and also to assist in production control in the primary mill. One computer is used solely to optimize the yield cut at the flying shear in the bloom, slab and billet mill.

Fully adequate and reliable transmission of information between process stations is essential to smooth flow running, and the English Electric Co. Ltd, as the main electrical contractor to the new plant, in collaboration with the consultants, the International Construction Co. Ltd, called upon the Marconi Co. Ltd, and The Marconi International Marine Co. Ltd for the planning and installing of visual and audio intercommunication facilities respectively.

To enhance the smooth running of the production processes a Marconi tabular cathode-ray tube information display system, controlled by English Electric-Leo computers, is installed in the primary mill. This tabular display equipment is ideally suited for this type of application being thoroughly proved and capable of full 'on-line' operation. The function of the tabular display equipment at Park Gate is to display instructions to the different operators in the mill. The instructions that appear on the display screens are in plain language form and are continually up-dated as information is fed back to the computer from the various parts of the plant.

Mounted on the production controller's desk are repeat displays of the bloom and billet mill schedules in addition to a number of input keys and switches. From this position the production controller is able to monitor the mill operation and to instruct the computer to modify any of the information being displayed to the mill operators.

The Marconi tabular display system is capable of being viewed in bright daylight, it will display data, in alpha-numeric form, on a cathode-ray tube at a writing speed of 50 000 characters per second. A total of 50 different characters can be used for display and three different character sizes are available to provide variations in emphasis in the displayed information. The equipment uses transistors throughout with the exception of the cathode-ray tube and e.h.t. rectifiers.

A closed-circuit television camera channel provides the bloom mill operator with a view of the reverse side of the mill which would normally be out of sight. This enables the operator to detect 'fishtailing' or other rolling faults that may cause damage to the rolls.

In the primary mill a secondary information system is also provided by a Marconi Marine audio-communication network. Special remote control facilities for overhead crane drivers are incorporated in the Marconi Marine system.

The main amplifier racks of the audio system, one assembly situated in each mill, embody the latest techniques in amplifier design and have been specially built for Park Gate. The four amplifiers in each rack are among the most powerful solid-state amplifiers ever built, each delivering 100W into a 50V line.

Visual information display is not installed in the strip mill, and communication between control and process points is by the audio system. Production control is carried out from the main control pulpit where the operator has a microphone/loudspeaker unit linking him with all the various sub-centres of production along the length of the mill.

From the strip mill stockyard the slabs are carried by overhead cranes to the reheating furnace and, as in the primary mill, Marconi Marine have provided communications facilities linking these various stages of production. Three wall mounted weatherproof microphone boxes with associated 42in horn loudspeakers have been fitted at convenient positions in the stockyard, while the two overhead cranes have loudspeaker/microphone units operating on the inductive loop principle. Further loudspeaker/microphone units are fitted in the reheating furnace discharge desk and at the furnace receiving pulpit.

After heating, the steel passes through the strip mill emerging as steel strip of the required thickness and width. At convenient intervals along the mill are fitted five telephone handsets with four associated loudspeakers, providing a communication link to the loudspeaker/microphones in the main pulpit where a portable microphone is also provided. As strip the steel passes automatically to the coiling section to be wound and strapped. From the loudspeaker/microphone unit in the coiler pulpit the operator can communicate with the strip mill, loudspeaker/microphone units at the strapping position, coil batching station, and coil weighing position, as well as with the two coil bay cranes, the cabs of which are again connected into the sound system on the inductive loop principle.

Desk loudspeaker/microphone units have been provided in the engineer's, mill manager's, shift manager's, and foreman's offices providing communication to the stockyard, discharge area and strip mill. As in the primary mill communications facilities are also provided in the motor room and lubrication cellars. In the motor room a loudspeaker/microphone is fitted in an acoustic booth, while a loudspeaker/microphone desk unit has been installed in the motor room attendant's office and a weatherproof loudspeaker/microphone wall unit in the lubrication cellar. The fitters' and electricians' areas are provided with direct communication to the main control pulpit, motor room and lubrication cellars, each area having two loudspeaker/microphone units.

The entire system, devised in close consultation with Park Gate and English Electric engineers, provides a communications network designed to assist in the efficient flow of work throughout the plant, and in addition to point-to-point communication incorporates an elaborate personnel calling system. In order to provide the highest possible insurance against the breakdown of communication through component failure, facilities for cross-switching of main amplifiers is to be incorporated.

A Transistorized High-Voltage Regulator Using A.C. Control

By I. Izumi* and M. Kokubu*

A method is described which is to regulate at an a.c. low-voltage level and convert to a required high-voltage output with the use of an impedance, in series with the transformer primary. It is similar to a d.c. to d.c. convertor type regulator or a magnetic amplifier type regulator, but it uses no d.c. to d.c.convertor and its recovery is not restricted by oscillator- or line-frequency.

A practical circuit for 200 to 800V, 0 to 5mA, and a stabilization ratio of the order of 100 from d.c. to several hundred cycles per second is described.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

THE design of a high-voltage transistor regulator is complicated by the limited collector-emitter voltage ratings of transistors. This limitation can be overcome by using several transistors in a series stack, but difficulties arise when the regulator is used for stabilizing voltages in excess of several hundred volts.

Another method uses a d.c. to d.c. convertor where a high-voltage output is sensed and regulation is performed at a low-voltage, high-current point while the regulated output is transformed to the desired high-voltage, low-current level. This method is able to stabilize a voltage higher than several thousand volts and choice of a fairly high oscillator frequency permits good smoothing of a rectified output with small capacitors. This technique seems awkward and complicated in a mains operated instrument, because a.c. power is rectified, then converted to a.c., and transformed high voltage is rectified again.

Here a method is proposed to regulate at high voltage level. The operating principle is similar in some respects to that of the magnetic amplifier type regulator: both control a.c. mains voltage so as to maintain the output voltage constant. The chief difference between the two is the frequency response of the feedback loop.

This article describes the operating principle of this regulator and some design examples of high voltage regulators for radiation detectors such as chambers and proportional or scintillation counters.

Principle of Operation

Fig. 1 shows the basic circuit of this regulator. The a.c. input voltage is transformed to a high-voltage and then rectified and filtered. A portion of the d.c. output voltage is sensed by the voltage divider R_4, R_5 and R_6, and delivered to the regulating amplifier to compare with a reference voltage. If the output voltage tends to increase, the output of the regulating amplifier will decrease, increasing the current through R_2 from the transformer secondary lower windings and reducing the voltage across the transformer primary, to restore the output voltage to its original value. It should be noted that regulation is not achieved without a change in voltage drop across R_1, in series with the transformer primary.

D.C. Feedback Loop Gain

The d.c. loop gain G_o is determined as follows:

$$G_o = \beta A_o G_1 \qquad (1)$$

where $\beta = R_6/(R_4+R_5+R_6) = V_{ref}/V_o$

A_o = voltage gain of the regulating amplifier

G_1 = output voltage change due to the output voltage change of the regulating amplifier = dV_o/dV.

It is shown in the Appendix that G_1 is expressed as:

$$G_1 = \frac{n_3 \cdot 1/R_2}{1/R_2 + n_3^2/(R_L+R_3) + n_1^2(1-2/\pi \cdot \sin^{-1}\rho)/R_1} \cdot \frac{R_L}{R_L+R_3} \qquad (2)$$

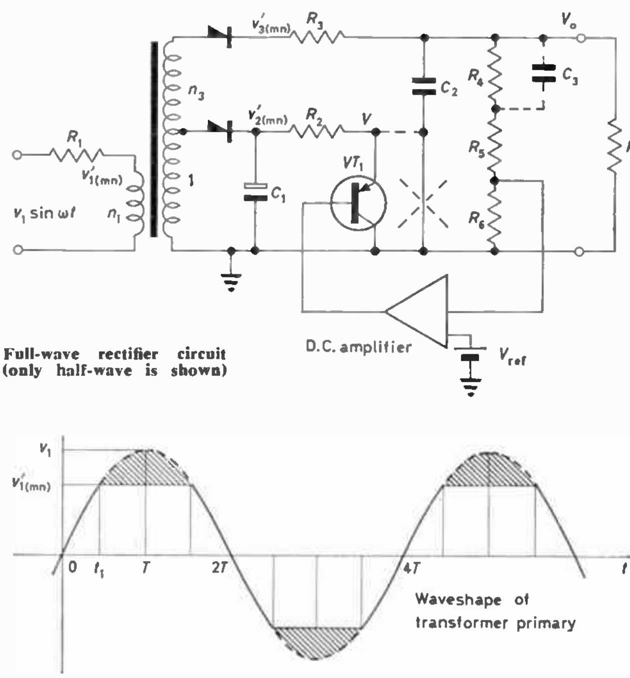


Fig. 1. Basic circuit

G_1/n_3 becomes zero dB in the ideal case where the value of R_2 becomes zero or the values of R_1 and R_L become infinity, but G_1/n_3 will lie between -3dB and -10dB in a typical circuit. An experiment by a model circuit where ρ, R_1, R_2 and R_L are widely changed, shows good agreement with equation (1).

Although β becomes smaller as higher voltage regulators are designed, loop gain G_o will change little because n_3 becomes larger according to the output voltage.

Frequency Characteristics of the Loop Gain

The loop gain $G(s)$ is determined by the time-constants τ_1 and τ_2 of two low-pass filters at a sufficiently lower frequency than 50c/s.

* *Mitsubishi Atomic Power Industries, Inc., Omiya, Japan.*

$$G(s) = \frac{G_o}{(1 + \tau_1 s)(1 + \tau_2 s)} \qquad \ldots\ldots\ldots\ldots (3)$$

Ripple content is not improved by this regulator and it is difficult to obtain good feedback stabilization with the use of a higher order of filter network because the filters are in the feedback loop. Therefore a large value high voltage capacitor must be used to provide sufficiently low ripple. Moreover, transient recovery time is restricted by the filters. The d.c. to d.c. convertor type regulator is better in this respect, because a fairly high oscillator frequency obviates these disadvantages.

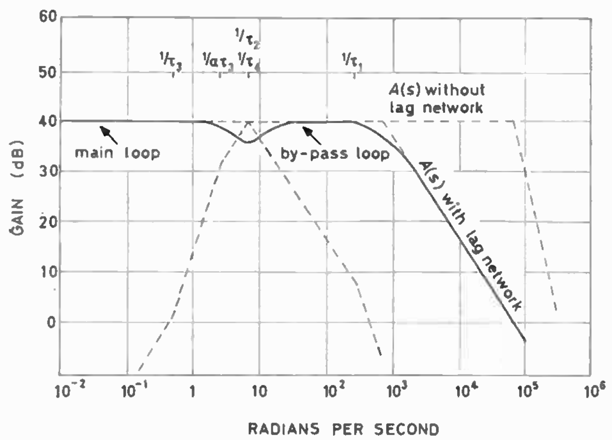


Fig. 2. Frequency response of open-loop gain

A higher frequency bypass loop can be obtained, however, in order to avoid these disadvantages by connecting C_2 and C_3 as shown by the dotted line in Fig. 1. This bypass loop gain $G'(s)$ is:

$$G'(s) = A(s) \cdot \frac{R_6}{R_5 + R_6} \cdot \frac{\alpha(1 + S\tau_3)}{(1 + \alpha S\tau_3)} \cdot \frac{S\tau_4}{1 + S\tau_4} \qquad \ldots (4)$$

where $\tau_3 = R_4 C_3$

$\alpha = (R_5 + R_6)/(R_4 + R_5 + R_6)$

$\tau_4 = C_2 \, . \, R_L R_3/(R_L + R_3)$

The overall loop gain is the parallel combination of $G(s)$ and $G'(s)$. Flat characteristics with frequency from d.c. to the cut-off frequency of the regulating amplifier can be obtained by the proper selection of time-constants τ_1, τ_2, τ_3 and τ_4, and by satisfying the following equation.

$$G_o = A_o \, . \, R_6/(R_5 + R_6) \qquad \ldots\ldots\ldots\ldots (5)$$

In this case the transistors, rather than the filters, play a determining role in the feedback stability of the regulator.

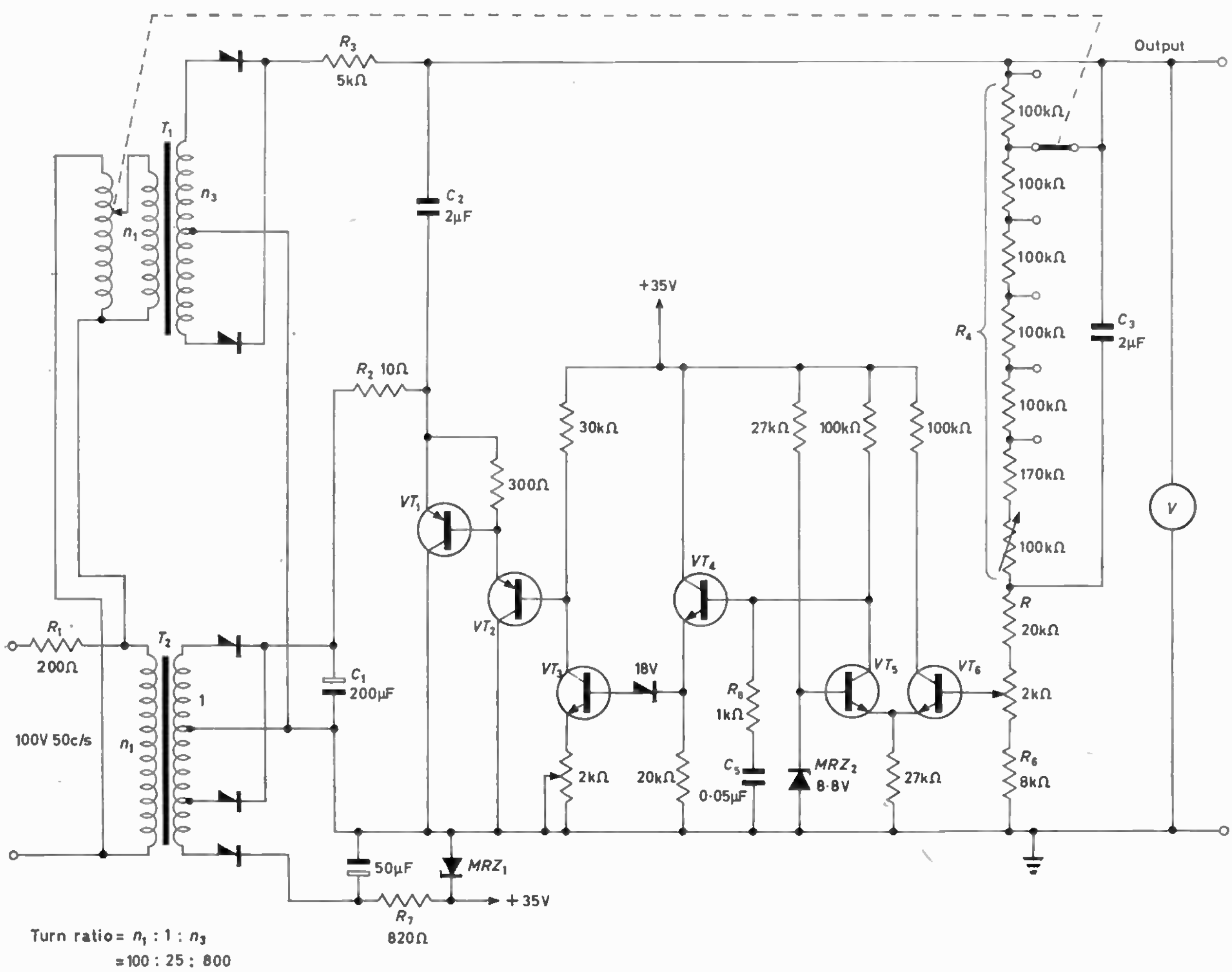
Recovery from a load transient is complete, to within regulation limits, in less than 0·1msec. It is mainly restricted by the bandwidth of the closed feedback loop. This regulator is not inferior to the d.c. to d.c. convertor type regulator in this respect and much superior to the magnetic regulator.

Regulator Performance

Regulator performance can be expressed in terms of S, R_o and K_T in the regulation equation.

$$\Delta V_o = S\Delta v_1' \, . \, V_o/v_1' + R_o \, . \, \Delta I_L + K_T \Delta T \, . \, V_o/V_{ref} \qquad \ldots . (6)$$

Fig. 3. Arrangement of 200-800V supply



where S = stability factor = $1/G_o$

R_o = output resistance

$$= 1/G_o \left\{ R_3 + n_3^2 \cdot \frac{R_1 R_2}{R_1 + n_1^2 (1 - 2/\pi \cdot \sin^{-1}\rho) R_2} \right\}$$

K_T = equivalent input temperature-induced drift voltage of the regulating amplifier which can be reduced less than 1mV/°C.

Circuit Description

A practical circuit for 200 to 800V, 0 to 5mA, stabilization ratio of approximately 100, is shown in Fig. 3. It comprises:

TRANSFORMERS AND RESISTORS R_1 AND R_2

Transformer secondary windings are determined by the required output voltage and the dynamic range of the regulating amplifier. A value of R_2 as low as is practical should be used to maximize d.c. loop gain. On the other hand by the choice of the optimum value of R_2, substantial reduction in maximum power dissipation of power transistor VT_1 is obtained. The a.c. by-pass loop gain cannot be obtained in the absence of R_2. The value of R_1 may be selected as $R_1 \geq n_1^2 (1 - 2/\pi \cdot \sin^{-1}\rho) R_2$. In this case G_o/n_3 becomes less than −6dB when the load current is negligible.

RECTIFIERS AND FILTERS

The output filter is necessary to reduce the ripple to an acceptable level in view of the stabilization ratio expected from this circuit.

REGULATING AMPLIFIER

The required voltage gain of the regulating amplifier will be approximately 10dB higher than the desired loop gain. It is also necessary to obtain considerable current gain. The first stage comprising of VT_5 and VT_6 is an emitter-coupled difference amplifier operated at a very low collector current to obtain low drift and high input impedance. The reference voltage is obtained from a compensated Zener diode MRZ_2 with temperature coefficient of less than 0·002 per cent/°C.

Transistor VT_4 is an emitter-follower which reduces the base current requirements of VT_3 in the collector of VT_5, thereby permitting VT_5 to develop a large voltage gain. Transistor VT_3 may be viewed as a grounded-emitter, which is driven by the emitter of VT_4 through a Zener diode. Transistor VT_2 acts as an emitter-follower in transferring the regulating signal from VT_3 to transistor VT_1, which also operates as an emitter-follower.

Feedback stabilization is obtained with C_5 and R_8. Zener diode MRZ_1 is operated by one of the windings of T_2 and provides a +35V potential for the regulating amplifier.

VOLTAGE DIVIDING NETWORK

The output voltage is divided by R_4, R_5 and R_6 to a level comparable with the reference voltage. The divider current is set to 1mA. The dividing ratio of R_5 and R_6 is obtained from equation (5). Low-temperature-coefficient resistors should be used in the divider, especially to reduce short-term drift. The value of C_3 is determined by consideration of the desired lower cut-off frequency of the by-pass loop.

Results and Conclusions

Performance of this regulator is as follows:

Output voltage	200 to 800V
Load current	0 to 5mA
Output resistance	less than 100Ω
Stabilization ratio	approx. 100:1
Ripple, 100c/s content	less than 10mV r.m.s.
Recovery time (to within regulation limits)	less than 0·1msec.
Temperature coefficients (0°C to 50°C)	less than 0·01 per cent/°C
Long-term stability (in a constant environment)	better than 0·1 per cent/day
Short-term stability	better than 1mV/5sec
Noise	less than 1mV p.p.

(in the bandwidth of 0·1 to 4c/s)

There are no difficulties in modifying it to satisfy other output requirements. A circuit for 1kV to 3kV, 0 to 1mA can be constructed with a few changes of the circuit shown in Fig. 3 and similar results are obtained.

This regulator is similar to a shunt regulator in view of low efficiency and inherent overload and short-circuit protection. Low efficiency is not thought to be a serious disadvantage when the equipment is mains operated and power levels of a few watts are involved.

The development of this regulator clearly demonstrates that a simple high-voltage regulator can be designed with a little addition to ordinary low voltage shunt regulator techniques and is expected to have good performance and reliability.

APPENDIX

CALCULATION OF GAIN G_1

Referring to Fig. 1, the following equations can be obtained on the assumption that the transformer and the rectifiers are ideal and the values of C_1, C_2 are large enough.

The power P_1 transmitted from primary to the secondary

$$= \text{the power } P_2 \text{ consumed at the secondary} \quad \ldots\ldots (1)$$

$$P_1 = 1/T \int_{t_1}^{T} v_1'{}_{(mn)} \cdot \frac{(v_1' \sin \omega t - v_1'{}_{(mn)}}{R_1} \cdot dt \quad \ldots\ldots (2)$$

$$P_2 = \frac{v_2'{}_{(mn)} - V}{R_2} \cdot v_2'{}_{(mn)} + \frac{v_3'{}_{(mn)}{}^2}{R_L + R_3} \quad \ldots\ldots\ldots\ldots (3)$$

where $v_1'{}_{(mn)} = v_1'\rho$ and other symbols are defined in Fig. 1. From equations (1), (2) and (3) the following relation is obtained where v_1, v_3 etc. are converted to secondary lower windings.

$$V = v_{2(mn)} \left\{ 1 + \frac{n_3^2 R_2}{R_L + R_3} + \frac{2 n_1^2 R_2}{\pi R_1} \left((\pi/2) - \frac{\sqrt{(1 - \rho^2)}}{\rho} - \sin^{-1}\rho \right) \right\} \quad \ldots\ldots\ldots\ldots (4)$$

Differentiation of equation (4) gives:

$$G_1 = dv_{2(mn)}/dV = \frac{n_3 \cdot (1/R_2)}{(1/R_2) + \dfrac{n_3^2}{R_L + R_3} + (n_1^2/R_1)(1 - (2/\pi)\sin^{-1}\rho)} \quad \ldots\ldots (5)$$

Acknowledgments

The authors wish to thank Mr. H. Asai for his continued support and discussion and Mr. M. Okano for his discussion.

REFERENCES

1. SCHWARTZ, S. Selected Semiconductor Circuits Handbook, parts 8 and 9 (John Wiley & Sons Inc. 1960).
2. RITSON, F. J. U., FOSS, R. C. Transistor Power Supplies with Limited Overload Current. *Electronic Engng.* *34*, 526 (1962).

A Cold-Cathode Valve Binary Delay Line

By D. Q. Mayne*

A simple cold-cathode valve binary delay line, which utilizes two switched power supplies, is described. The two power supplies enable a binary signal to be transferred from an A-line to a B-line and, after a delay of one unit, to be transferred back to the A-line.

(Voir page 577 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 584)

A BINARY delay line is shown in Fig. 1. A and B are supplied from switched power supplies whose waveforms are illustrated in Fig. 2. The parameters of the

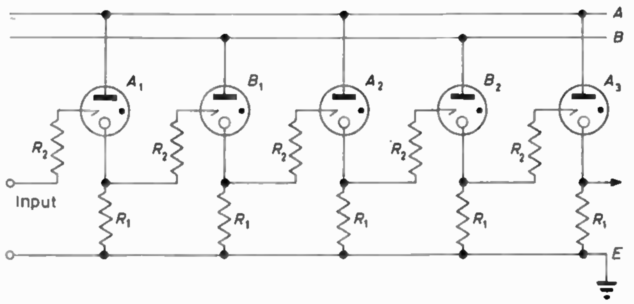


Fig. 1. Binary delay line

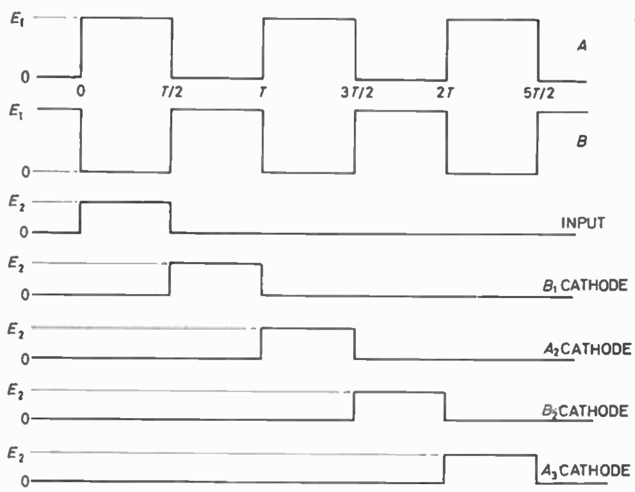


Fig. 2. Delay line waveforms

circuit are such that E_2, the cathode voltage of a valve which has been struck, is sufficient to cause the trigger of the next valve in the chain to strike. Thus if the input signal shown in Fig. 2 is applied to the trigger of A_1, A_1 will strike during the interval 0 to $T/2$. During this interval, therefore, the trigger of B_1 will strike, thus ionizing B_1. At time $T/2$, A_1 is switched off and so is the trigger of B_1. However the anode of B_1 reaches the value of E_1 volts before B_1 has time to deionize, and so B_1 strikes. This causes the trigger of A_2 to strike, and at time T, when B_1 is switched off, A_2 strikes. Thus the input signal is propagated down the delay line, appearing with a delay of T sec at the cathode of A_2, and a delay of $2T$ sec at the cathode of A_3, etc.

A delay line employing GTR120W trigger valves was constructed. The relevant properties of this valve are:

Maximum anode voltage to prevent self ignition in all valves .... +310V

Anode-cathode running voltage at 4·5mA .... 95 to 140V

Minimum trigger voltage to cause breakdown in all valves 170V

Deionization time 3msec

E_1 was chosen to be 310V. E_2 thus varies from 215 to 170V, and some preselection of valves to exclude those with a high running voltage might be necessary. Alternatively trigger bias might be added as shown in Fig. 3. R_1 was chosen to be 47kΩ so that the running current would be approximately 4·5mA, and R_2 220kΩ. The minimum value of T was found to be about 6msec, due to the deionization time of 3msec.

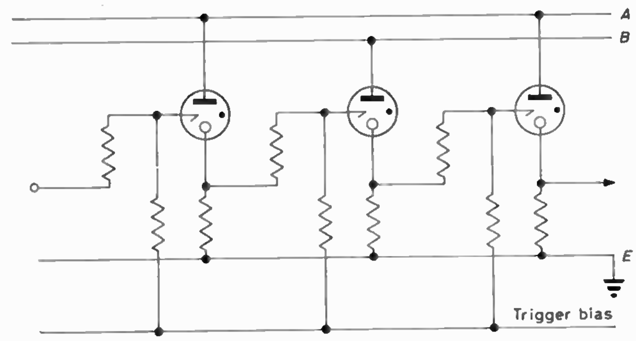


Fig. 3. Modified delay line

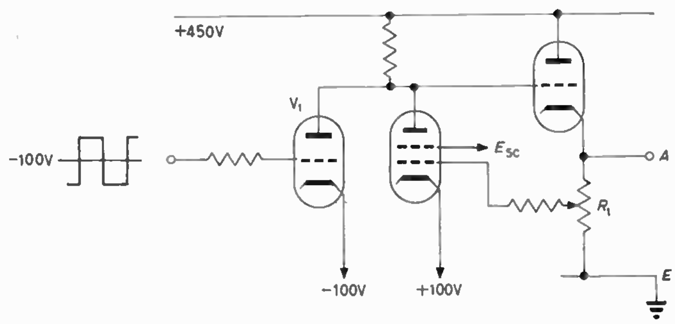


Fig. 4. Switched power supply

By controlling the trigger bias of an individual valve, this valve can be made to act as a gate. Thus the circuit needed to obtain the two important operations of delaying and gating is very simple, though at the expense of requiring more complex power supplies.

Power Supplies

The A power supply, shown in Fig. 4, consists of a conventional voltage regulator whose output voltage is controlled by a setting of R_1. A square wave fed to V_1 cuts off the regulator for half-cycle intervals when the voltage at A falls rapidly to zero. The voltage at A is thus a square wave, as desired. The output of A can be used to switch a similar regulator on and off to supply line B.

* *Imperial College of Science and Technology.*

LETTERS TO THE EDITOR

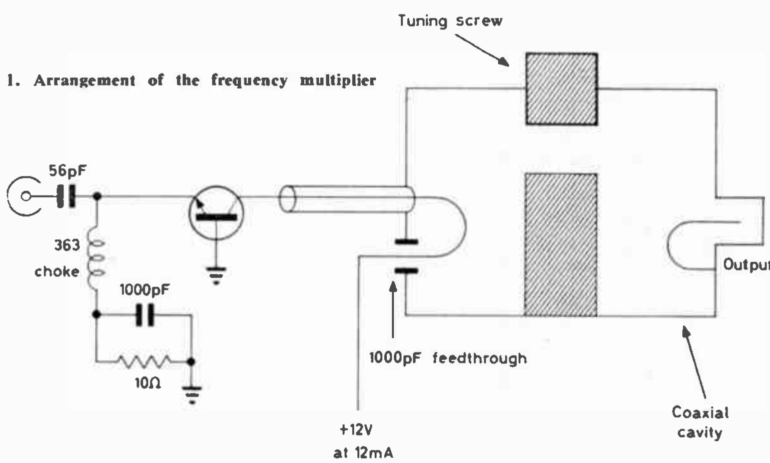
(We do not hold ourselves responsible for the opinions of our correspondents)

A Transistor Frequency Multiplier

Dear Sir,—Your readers may be interested in a transistor frequency multiplier which was developed here recently and which could have alternative applications.

The requirement was for a phase-stable, ×4 multiplier using an input frequency of 363Mc/s. Some tests done on a Varactor multiplier showed that the phase change and output power were critically dependent on the input power. Alternative multiplier systems were therefore investigated.

Fig. 1. Arrangement of the frequency multiplier



The arrangement we have used is that of a grounded base 2N918 transistor with zero nominal bias (Fig. 1).

An input power of 15mW gives 6mW out at 1452Mc/s, and a typical power characteristic has the shape shown in Fig. 2.

The phase changed by 13° for 1dB change of input power and this was reduced by stabilizing the input power with a d.c. control amplifier. This measures the voltage across the 10Ω resistor; the phase changes are then about 5° for 3dB change of input. The 3dB bandwidth is set by the coaxial cavity and its loops and is 80Mc/s, i.e., input frequencies between 353 and 373Mc/s. The cavity can also be tuned to the third and fifth harmonics of 363Mc/s.

Other arrangements have been used in the laboratory which gave 25mW out at 1452Mc/s for an input power of 68mW.

A system such as this could have applications up to 2000Mc/s and, with future transistor types, it should be possible to obtain higher frequencies as harmonics up to the 8th of 363Mc/s have been observed with the present arrangement.

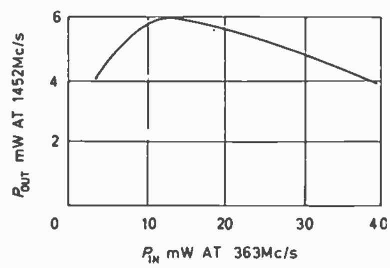
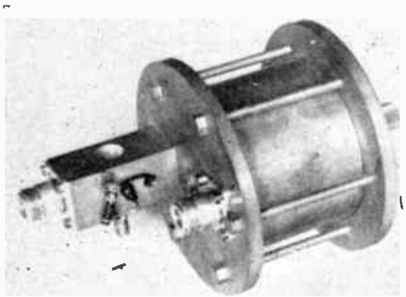


Fig. 2. Power characteristic

The complete unit



Yours faithfully,

N. Owen,

Cavendish Laboratory,
University of Cambridge.

L.F. Noise in Transistor Direct-Coupled Amplifiers

Dear Sir,—I read with interest Mr. Pinto's article 'Transistor Low Drift D.C. Amplifiers'[1]. I have covered similar ground in a recent investigation, and agree with his remarks on pages 304, 305, and in the first column of page 306, which form a useful commentary on the present state of the art. The 'Mirror Image' amplifier is an interesting application exploiting the possibilities of balanced circuits using silicon planar transistors. There is, however, one fundamental limitation to these circuits which I have not seen described, namely, random zero drift due to transistor noise. Before silicon planar transistors were introduced, a balanced direct-coupled transistor amplifier might be expected[2] to have a temperature coefficient of 100μV/°C. Then transistor noise was often masked at frequencies below, perhaps, 10c/s, by the effects of temperature changes. Now that temperature coefficients of 1μV/°C can be achieved, transistor l.f. noise is the dominant source of amplifier drift, unless temperature fluctuations of several tens of degrees Celcius are experienced.

Mr. Pinto describes the use of an oven to temperature stabilize, to ±1°C, an amplifier having, presumably, the 2μV/°C temperature coefficient mentioned. One would expect output fluctuations equivalent to ±2μV, or 4μV total. Actually Mr. Pinto observed ±20μV for short-term and ±40μV for long-term measurements. The discrepancy is almost certainly mainly due to transistor l.f. noise. Some fluctuations may be caused by temperature differences between the balanced pairs of transistors due to convection 'draughts', but these effects are likely to be small in an oven forming an isothermal enclosure.

I have made recordings of the output fluctuations produced by an amplifier using balanced stages with the double version of the C111, the 2C111. R_g was 200Ω, I_c was 20μA, no input was applied, except for calibration. Two samples were tried in the input stage, selected for V_{be} of the halves of each transistor to be within 10mV. The temperature coefficient resulting was less than 5μV/°C. Both transistors produced noise records of the same character and amplitude when compared over periods of a few minutes. For one sample, in a period of 5 hours the output fluctuated between limits equivalent to a range of about 50μV at the input. During most of the period the ambient temperature changed less than 0·2°C. The figure of 50μV seems to agree with Mr. Pinto's long-term observation of 80μV. Naturally close agreement cannot be expected as the observed figure will depend on the measuring time and on variations between transistors. The difference in I_c and R_g between the amplifiers would have a slight effect. It would be interesting to know over what period Mr. Pinto's measurements were made.

The C111 is not recommended by the manufacturers as a low noise type, and other types tried in my circuit, with no other component changes produced smaller fluctuations, for example, 15μV in 5 hours.

It is apparent that an important property of transistors for use in balanced direct coupled amplifiers is the l.f. noise. Temperature coefficients of 0·2μV/°C obtained by Middlebrook and Taylor[3], are useful if large changes of ambient temperature are expected, as, for example, in space vehicles, to keep the total zero drift small. With the use of temperature stabilization it does not appear worth spending manufacturing time in minimizing the temperature coefficient of the circuit, unless transistors with much improved l.f. noise become available.

It is to be hoped that transistor manufacturers may be able to reduce the level of the l.f. noise, and, particularly, may be able to curtail the low frequency part of the l.f. spectrum.

Yours faithfully,

S. J. M. WHITFIELD,
London, N.W.11.

REFERENCES

1. PINTO, J. J. Transistor Low Drift D.C. Amplifiers, *Electronic Engng*, *36*, 304 (1964).
2. CHAPLIN, G. B. B., OWENS A. R. Some Transistor Input Stages for High-Gain D.C. Amplifiers, *Proc. Inst. Elect. Engrs. 105B*, 249 (1958).
3. MIDDLEBROOK, R. D., TAYLOR, A. D. Differential Amplifier with Regulator Achieves High Stability, Low Drift. *Electronics*, *34*, 56 (28 July 1961).

The Author Replies:

DEAR SIR,—Mr. Whitfield's comments on noise in transistor low drift d.c. amplifiers are useful and informative, especially in the context of drift compensation, when drifts of less than 1μV may be achieved.

Low frequency or 'flicker' noise in transistors is generally considered to have two main components[1]: surface noise, which is caused by fluctuations of energy levels at junction surfaces and is dependent on the current, so that for minimum noise the transistor is operated at an optimum collector current; leakage noise, which is caused by variations of leakage current across the junctions and is dependent on the voltage applied, being almost negligible for good transistors at low voltages. Noise in transistor amplifiers is also very much dependent on the source resistance, and is minimum for most transistors when this is approximately $1\,000\Omega$. Mr. Whitfield is quite right in suggesting that, when drift compensated direct-coupled amplifiers are being designed for extremely low drifts, noise becomes increasingly important. Collector voltages and currents and the source resistance must be chosen for optimum noise performance, bearing in mind also the choice of these factors with regard to drift.

The use of selective amplification (reduction of bandwidth) is a method that may be used to advantage in the reduction of noise. The 'Mirror Image' amplifier discussed in my article was developed for use with thermocouples and, like most low drift d.c. amplifiers, the required speed of response was extremely slow. The large capacitors connected across the output collectors of this amplifier have the effect of integrating random noise. The time-constants involved could be increased even further to make the amplifier sensitive only to very slow changes, without effecting the drift performance and considerably reducing the effects of noise.

Mr. Whitfield's observations on noise in balanced amplifiers using the transistors mentioned agree with my experience. I have, however, succeeded in reducing random fluctuations by the integration technique and have observed drifts of less than 5μV over periods of five to eight hours. The C111 is not a low noise type, but the 'Mirror Image' amplifier was thoroughly investigated for use with general purpose planar transistors and the drifts quoted in my article refer to tests done on about a dozen randomly selected samples.

I would like to mention that the amplifier with a drift of $2\mu V/{}^\circ C$ was being built using a single-chip double transistor (Texas Instruments type BN209) for the first stage. The average drifts of $\pm 20\mu$V for the temperature stabilized circuit refer (this was mentioned in the article) to the amplifier using unselected planar transistors.

Yours faithfully,

J. J. PINTO,
Kent Precision Electronics Ltd,
Tonbridge, Kent.

REFERENCE

1. BENETEAU, P. J. Designing Low Noise Transistor Circuits. *Electronic Design*, (3 August, 1960).

Identification of Aliased Traces in L.F. Sampling Oscilloscope Ranges

DEAR SIR,—During work carried out in this Department, a Lumatron model 120 sampling oscilloscope has been used. This instrument has an extra time-base facility: the 'chopped' mode, with time-base ranges of 1msec/cm and 10msec/cm. These ranges use the sampling circuits and amplifiers driven at a constant 50kc/s p.r.f., and generate a real-time x sweep ramp rather than the staircase of the sampling ranges. The reason for their inclusion is to permit the viewing of l.f. phenomena of the order 10c/s to 10kc/s on the same instrument as used for viewing high speed pulses.

While using these ranges, an ambiguity was noted; if while in the 'chopped' mode a signal of frequency near to the sampling frequency or one of its harmonics was applied to the input, an 'aliased' low frequency version appeared on the screen.

Thus an operator, faced with a steady trace which appeared to be, say, a 1kc/s sine wave, had no immediate way of knowing if the input signal was a sine wave of 1kc/s or one of frequency $(nf_s \pm 1\text{kc/s})$ where f_s is the sampling frequency.

The method to be described makes it immediately obvious if an 'aliased' trace should be displayed.

Fig. 1 shows the variation of trace frequency with sampling frequency for a given input signal frequency, and immediately suggests a way in which the trace may be identified as a 'real' version of the input signal or as an 'aliased' version.

In the region where $f_s > 2f_i$, the output frequency is constant and equal to f_i. In the region where $f_s < 2f_i$, the output frequency f_o is given by $f_o = f_i - nf_s$, where n is the nearest whole number to (f_i/f_s). It can be seen that here f_o varies between 0 (d.c.) and $\frac{1}{2}f_s$ as f_s varies.

Thus if a slight amount of frequency modulation is introduced into f_s, the output frequency will be unaffected provided $f_s > 2f_i$, but if $f_s < 2f_i$ there will be a resultant frequency modulation of the output frequency f_o. This is the basis of identification.

The clock circuit producing the constant frequency is an RC relaxation

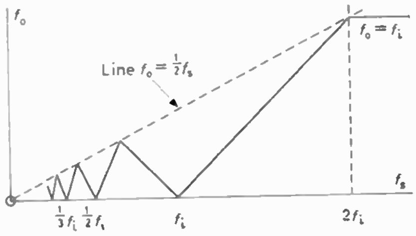


Fig. 1. Variation of trace frequency with sampling frequency

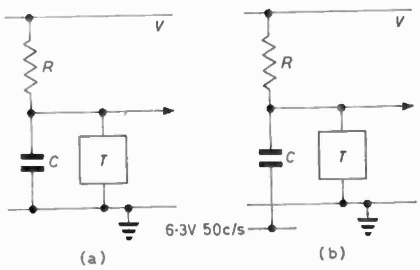


Fig. 2. Clock circuit and introduction of f.m.

oscillator of the type in Fig. 2(a). T is an active device for regularly discharging C.

A suitable amount of f.m. was introduced into the clock frequency by altering the circuit to that of Fig. 2(b). 6·3V 50c/s is found typically to give a maximum deviation of the order of 1kc/s for a basic clock frequency of 50kc/s.

With the f.m. introduced into the sampling frequency the trace is unaffected if the input signal frequency is $< f_2$ and readings can be taken confidently. Any aliased trace now contains 50c/s f.m. and can no longer be made to give a steady uniform trace, so that as soon as such a distorted trace appears on the screen the operator knows he must go to a higher time-base range in the sampling mode in order to make any measurements. Since the modified clock circuit is not used in the normal sampling ranges, no changes occur in these ranges.

This method, applied to the sampling oscilloscope in use by the author, provides a useful feature requiring no extra components and only one wiring change to implement it.

Yours faithfully,

J. F. FULFORD,
Marconi Instruments Limited,
St. Albans, Hertfordshire.

BOOK REVIEWS

Digital Storage Systems

By W. Renwick. 212 pp. Demy 8vo. E. & F. N. Spon Ltd. 194. Price 50s.

THIS book offers a review on Digital Storage Systems as they are used in present technology. Following a brief historical introduction and the establishment of the "language" of the book, the ensuing chapters deal in more detail with the various forms of storage. Although all of them are used for digital storage, some of them could be used as analogue stores. The forms of storage are divided into groups, each group forming a chapter. These are titled delay line storage, electrostatic and ferroelectric storage, magnetic surface recording, magnetic core storage, other magnetic storage elements, non-magnetic random access storage and non-erasable storage. Optical stores are included in this last group.

Each chapter explains the basic principles of the systems considered and points out some advantages and disadvantages. Additionally there is a fairly extensive bibliography attached to each chapter which should form a good start for anyone wanting to go deeper into the subject. The penultimate chapter deals with access and control circuits in random access stores and goes briefly into comparative costs. The last chapter reviews the storage systems that are presently available and considers their limiting factors. It also expands briefly on future trends.

There appears to be a numerical error in the figure of merit for a magnetic core store on p. 176 which should be 54dB on the figures quoted.

On p. 182 there is a statement that no optical store has been developed for dictionary translation but this has been reported in the literature in 1958.

This book provides a general introduction to the very wide field of digital storage systems and the bibliographies should give useful leads to deeper reading.

R. COUZENS.

Magnetic Tape Recording

By H. G. M. Spratt. 368 pp. Med. 8vo. 2nd Edition. Heywood & Co. 1964. Price 33s.

SINCE modern magnetic tape recording was born with the invention of the Magnetophon in Germany towards the end of the last war, the art, science and technology has progressed a great deal. Nowadays, magnetic tape is used for recording virtually every type of signal which was originally handled by other methods. Additionally, many new forms of recording are now possible with tape, which proved impracticable with earlier devices, these are notably wide-band recording and the recording of digital data.

It is welcome therefore, to see a second edition of Mr. Spratt's book the first edition of which was published some five years ago. The present text has been revised to include developments in ferrites and the like, pulse duration modulation and digital recording.

The book deals for the most part in a practical way not only with the principles of magnetism, magnetic recording and the associated equipment but also in fair detail with the materials used for and the actual manufacture of tape. There are also chapters on the testing of tape and of the recorders themselves.

The main criticism must be the fact that despite the title, the book deals mainly with recorders for speech and music. Scant attention is given to the very wide applications of magnetic recording in computer work, instrumentation, machine control and television waveform recording.

A. HARRIS.

Basic Pulse Circuits

By R. Blitzer. 436 pp. Med. 8vo. McGraw Hill, 194. Price 68s.

IT is often stated that everybody can write one book—based on his life story. Similarly it appears to be believed that every lecturer can write one text-book, presumably the one based on his lectures. Unfortunately that is only rarely true, because one's own lectures are given in a certain context: They are preceded and followed by other lectures and circumvented by them thus making them unsuitable for other courses. The book under review is an example of this: no doubt an excellent series of lectures given by Mr. Blitzer as a part of an advanced technicians' course has been turned into a second rate book which this reviewer cannot recommend for any course or reader.

The first fifty pages are devoted to matrix solution of d.c. "Network circuits" (*sic*!). This could well have been omitted completely for all the purpose it serves the rest of the text. The next chapter plunges straight into high frequency compensation of *RC* coupled amplifiers. The student is not only expected to be familiar with amplifiers but also with solution of a.c. networks. Symbols are used irregularly (db for dB, G_M for g_m); terms are used loosely ("capacitance C_{PD} increased by a factor called the Miller effect"); or wrongly (the frequency at which amplification is 3dB below maximum is referred to as "highest desirable frequency to be amplified"); undisciplined thinking and slipshod definition abound (Q . . . "is often called the resonant rise of voltage . . ." or ". . . if a second identical amplifier were added in cascade . . . the gain—would multiply"—we know what all this means but does it justify such clumsiness?)

Strict and unique relations are quoted for shunt and series compensation without giving the reader any inkling why or whether the equations quoted give the best results in all circumstances. This takes some forty pages and is followed by equally dogmatic chapters on linear and non-linear wave shaping and so on to multivibrators and "time base oscillators and generators" (*sic*!). Most of the common circuits are described qualitatively in great detail in both their valve and transistor forms but the information given is barely adequate to design any of them.

Methods of mathematical binary to decimal, binary to octal, octal to decimal etc., etc., conversions get full airing followed by some more manufacturers' blurb type of circuit descriptions. There is a reasonable chapter on gates. After a little more random reading about applications and some more "miscellaneous circuits" the book ends with a chapter on transient analysis about which this reviewer can only be kind by saying nothing. The price: £3 8s.; this money will be well spent on another book.

M. H. N. POTOK.

Introduction to Integrated Semiconductor Circuits

By A. J. Khambata. 233 pp. Med 8vo. John Wiley & Sons. 1963. Price 57s.

FEW electronic engineers will be unaware of the explosive impact the subject of integrated circuits is making on systems design, and many regard with apprehension the ultimate consequences for the circuit designer.

Mr. A. J. Khambata is Senior Development Engineer, Microcircuits Department, UNIVAC Division of Sperry Rand Corporation, and his wide experience has made him singularly well qualified as author of this compact but very comprehensive book. In his preface, he suggests that the book is intended for "technical and management people in the electronics industry who are on the threshold of becoming involved in the new and fascinating field of Integrated Circuits."

For a work of this type, the author has succeeded in keeping the book remarkably up to date, and a comprehensive bibliography will assist the reader in tracing extensions of the text.

In Section 1 of the book, a review is given of the historical background of microminiaturization, and the complexity of the subject is shown by the fact that it has been necessary to subdivide into seven principal categories, and to restrict the book to three. These include integration by thin film, semiconductor technology, and hybrid combinations of the two technologies.

To regard integrated circuits simply as a new sub-assembly of components is to misinterpret the significance of the new technique. New organizational relationships will be necessary, involving close co-ordination between systems designer and microcircuit designer, and Mr. Khambata examines this aspect with some care.

After reviewing the general technological steps of thin film and semiconductor practice, the reader is led gently towards circuits of increasing complexity. A particularly useful feature is a comprehensive and critical discussion of digital logic circuits.

Packaging of integrated circuits is a subject on which the circuit designer is usually vociferous. The author discusses the various packages currently available, and explains the predominance of the JEDEC TO-5. The importance of this aspect would seem to justify rather more emphasis than has been given to it, as the inter-relation of the type of package with the method of system assembly is of considerable moment to the user.

In a valuable forty-four page appendix, fully integrated standard product line circuits are discussed in detail, ten of the well known American manufacturers' products being analysed, followed by three multi-chip circuit sources; one thin film supplier is quoted, which reflects the general emphasis of the book.

Mr. Khambata can be assured of a wider audience than he had planned, and it is to be hoped that successive editions of this book will kept equally abreast of the subject.

I. G. CRESSELL.

Mathematical Techniques in Electronics and Engineering Analysis

By J. W. Head. 264 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 45s.

The object of this book is to make available to engineers and other users of mathematics, widely applicable techniques which will make their work easier. The subjects discussed are mainly those which have arisen in the course of the author's work in the Research Department of the B.B.C. Engineering Division.

The book is divided into two parts. The first half consists of general mathematical techniques—the solution of equations; series and partial fractions; differentiation and integration; versors, vectors and trigonometry; and various labour-saving devices. The second half deals with more specialized applications of mathematics to problems in electrical and electronic engineering—operational calculus; matrices applied to two-part problems of transistor circuits; conditions for minimum variation in a function; analytical geometry and impedance calculations; and stability criteria.

Matrices—Their Meaning and Manipulation

Edited by W. G. Bickley and R. S. H. G. Thompson. 168 pp. Demy 8vo. The English Universities Press. 1964. Price 21s.

This book is based on lectures given to third year and postgraduate engineering students of the Imperial College and opens with samples of the types of problem for the solution of which matrix algebra and the associated arithmetic are appropriate. Matrix algebra is then developed as a language in which linear equations and transformations can be concisely expressed and manipulated. Later chapters are concerned with the solution of sets of linear algebraic equations with the inversion of matrices, and with eigen-values and eigen-vectors. In these chapters the principles of the methods for the arithmetical solution of large-scale problems are explained and exemplified.

Field-Coupled Surface Waves

By J. R. Melcher. 190 pp. Med. 8vo. M.I.T. Press. 1964. Price 38s.

This monograph describes the behaviour of some simpler kinds of surface-coupled continuum-electromechanical systems—interactions which are basic to more sophisticated experiments now under way at M.I.T.

The electrohydrodynamic systems described point to a variety of significant research topics.

Emphasis is placed on the basic surface-interaction coupling mechanism rather than its applications. However, the dynamics described will interest those conducting research in such diverse fields as the weather, two-phase heat transfer in the presence of EHD fields, separation or confinement of fluids, magnetohydrodynamics electrohydrodynamics, and electro mechanical energy conversion.

Handbook of Microwave Measurements Vol. 1, 2 and 3

Edited by M. Socher and J. Fox. 2 364 pp. Demy 4to. John Wiley & Sons. 1964. 3rd. Edition. Price 300s.

The third edition of the Handbook of Microwave Measurements embodies a complete revision of earlier editions as evidenced by the completely new format and the considerable additions to the textual material and topics covered. Also, the form of presentation has been substantially revised and measurement problems have been treated in a more systematic manner. The revision was initiated, and supported in part, by the U.S. Army Electronics Research and Development Agency.

Propagation of Radio Waves

By B. Chatterjee. 115 pp. Demy 8vo. Asia Publishing House. 1963. Price 25s.

This book deals with several aspects of radio propagation in different frequency ranges. It begins with a discussion of some basic considerations in radio wave propagation. Chapters are then devoted to an analysis of different aspects of surface and space wave propagation including the effects of ground reflection. The structure and properties of the ionosphere and the elementary theories of scatter propagation are also discussed. The theories of sky wave propagation, as caused by ionospheric reflection, are specially treated as they are mainly responsible for long distance radio communication and broadcasting.

Fixed Capacitors

By G. W. A. Dummer. 270 pp. Demy 8vo. Sir Isaac Pitman & Sons. 1964. Price £2 5s.

In this book every type of capacitor is described, including recently developed evaporated dielectrics, for microelectronics. Data on fundamental characteristics, measurement techniques, specifications, common faults and future developments are all given and, in addition, a classified bibliography of some 700/800 references has been compiled.

Waves and Oscillations

By R. A. Waldron. 135 pp. Demy 8vo. D. Van Nostrand. 1964. Price 14s.

The author investigates specific wave properties and attributes, and shows how phenomena such as resonance and interference are manifested in waves of many kinds. Analogies are drawn between one kind of wave and another. While the subject is essentially mathematical detail has been reduced to a minimum and wherever possible discussions are carried on by a combination of language and diagrams. Where it is necessary to use sophisticated mathematical concepts, detail and rigor have been abandoned in favour of qualitative descriptions.

Beginner's Guide to Electronics

By T. L. Squires. 194 pp. Crown 8vo. George Newnes Ltd. 1964. Price 15s.

Industrial Switchgear Installation and Maintenance

By E. A. Reeves. 112 pp. Crown 8vo. George Newnes Ltd. 1964. Price 10s. 6d.

Cable Jointing

By R. Thompstone. 112 pp. Crown 8vo. George Newnes Ltd. 1964. Price 10s. 6d.

Electric Motors Installation and Maintenance

By R. Rawlinson. 103 pp. Crown 8vo. George Newnes Ltd. 1964. Price 10s. 6d.

Industrial Transformers Installation and Maintenance

By M. A. Spurway. 110 pp. Crown 8vo. George Newnes Ltd. 1964. Price 10s. 6d.

Electrical Installation Work

By R. A. Mee. 291 pp. Med. 8vo. Macdonald & Co. Ltd. 1964. Price 40s.

Experimental Radio Engineering

By E. T. A. Rapson. 213 pp. Demy 8vo. Sir Isaac Pitman & Sons. 1964. Price 14s.

Japanese Miniature Electronic Components and Assemblies Data Annual 1964-65

Edited by G. W. A. Dummer and J. Mackenzie Robertson. 483 pp. Demy 4to. Pergamon Press. 1964. Price £7

ELECTRONIC EQUIPMENT

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

(Voir page 571 pour la traduction en français; Deutsche Übersetzung Seite 578)

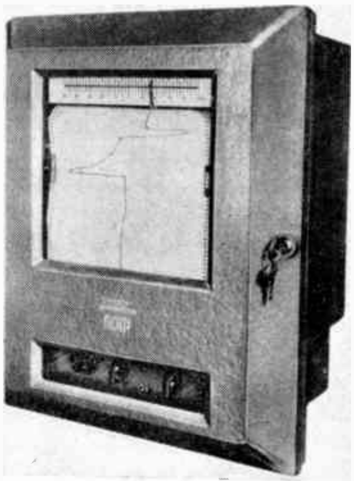
GALVANOMETER RECORDER

Distributed by: Claude Lyons Ltd, 76 Old Hall Street, Liverpool 3

(*Illustrated below*)

The type EG galvanometer recorder, manufactured by Association des Ouvriers en Instruments de Precision (A.O.I.P.) of Paris, is an 8in chart width instrument of robust construction and high reliability.

The recorder employs a sensitive shock-mounted mirror galvanometer, the servo-system being a patented spot-following arrangement employing two photo-resistive cells which, at balance bridge the spot. This system has the advantage that it incorporates inherent automatic searching for the spot on switching on.



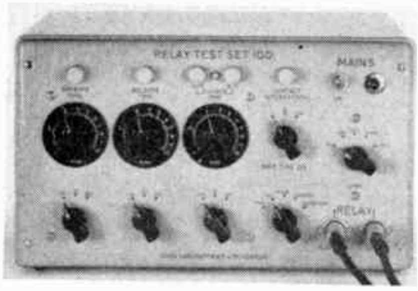
Electrical specifications include accuracy of ±1·5 per cent stability of ±0·5 per cent, 4sec time-constant and sensitivities up to 6μV/mm or 0·03μA/mm. Fourteen paper speeds between 24cm/min and 1·5cm/h are provided.

EE 72 751 *for further details*

RELAY TEST SET

Data Laboratories Ltd, 3-7 Hatfields, London, S.E.1

(*Illustrated above right*)



This unit was developed to carry out a complete test for relay operate and release time and bounce period, and for relay shock and vibrations as described in Defence Spec. DEF5165. In addition, it also allows an operational check on manufacturers' specification figures relating to the above relay features. It offers a convenient method of checking operational characteristics quickly and without ambiguity.

Accordingly when testing relay operation the relay is pulsed at a rate controlled by a variable rate multivibrator which can be set to suit the particular relay under test. It is possible to measure relay operate, release and bounce times simultaneously and these times can be separately set. If any of these characteristics exceed the set figure this is indicated on the corresponding lamp.

The operate time is taken from the point when the coil is energized to the point when the normally open contacts have settled to the closed position i.e. after the bounce period is over. This time can be set as required and if this is exceeded a lamp will indicate failure on that contact. A failure is detected if the contact opens for longer than 1μsec after the specified operate time is over.

Similarly when the relay is de-energized the release time is measured from the time the relay coil is de-energized to the time when the normally closed contact settles to the closed condition. Again this time can be pre-set and if it is exceeded a lamp will indicate failure on that contact.

When contact bounce is measured the time is taken from the point when the open contact closes to the point when it settles into the closed condition. This is taken on both the normally open and normally closed contacts.

When the unit is used for shock and vibration testing a check is made on the contacts under test that they do not open during the test period. Any discontinuity is detected and a lamp will indicate failure.

EE 72 752 *for further details*

REGULATED POWER SUPPLY

Feedback Ltd, Park Road, Crowborough, Sussex

(*Illustrated above right*)

This is a new general purpose unit for bench use which will operate from either 200 to 240V or 100 to 120V 50 to 60c/s supplies.

A special feature is the flexibility of connexion such that the following four modes may be obtained at will:—

(a) Two separate 'floating' 300V supplies delivering 150mA each.

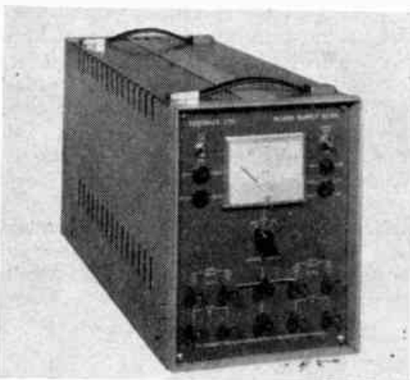
(b) A single h.t. pack delivering + 300V and − 300V at 150mA.

(c) A single h.t. pack delivering + 600V or − 600V at 150mA.

(d) A single h.t. pack delivering + and − 300V at 300mA.

For operation under constant minimum load conditions, a simple internal connexion increases the maximum loading on each supply to 200mA.

There are three 6·3V outputs, two of which are 'floating'. These are capable of feeding up to 16A of heater current. The large monitor meter reads the current from each output. Long term stability against supply variations is greater than 300:1 for total excursions of 15 per cent of the nominal value of the supply. Output impedance is less



than 1Ω and the total ripple content is less than 3mV r.m.s.

The outputs are available either on front panel mounted terminals or from two 6-way Painton sockets at the rear. These are compatible with the Feedback range of analogue and simulator units.

EE 72 753 *for further details*

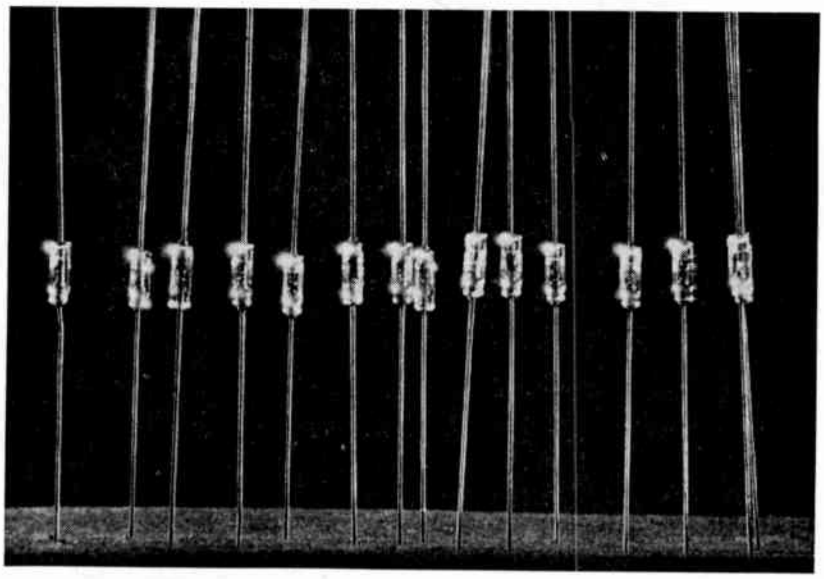
POLYESTER CAPACITORS

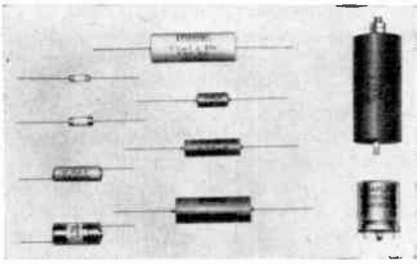
Capacitor Division, Plessey-UK Ltd, Kembrey Street, Swindon, Wiltshire

(*Illustrated on page 563*)

The polyester material known as 'Melinex' in the U.K. and 'Mylar' in the U.S.A. makes an excellent dielectric. It possesses a high dielectric strength which is well maintained at the elevated temperatures often found in modern electronic equipment. The use of metallized 'Mylar' enables capacitors of very high volume efficiency to be manufactured which are smaller than their metallized paper counterparts for the same working voltage.

Three principal ranges of polyester capacitors are now available from





Plessey-UK Limited. These are unprotected 'Melinex', moulded 'Melinex' and moulded metallized 'Mylar' types. In addition, polyester capacitors for very high voltages—up to 15kV—and polycarbonate units providing low power factors at temperatures up to 150°C can be manufactured to special order.

The unprotected 'Melinex' range covers standard values between 10pF and 2·0μF at 250V d.c. working, 1·0μF at 400V and 0·1μF at 1kV. These capacitors are processed at high temperature, and the ends are resin sealed. Humidity classification is H3 and temperature range −40°C to +125°C.

Moulded 'Melinex' capacitors are available in seven sizes which cover approximately similar values and working voltages to those given above; the humidity classification is however improved to H2 standards while the temperature range is −40°C to +100°C.

With the same humidity classification and temperature range, moulded metallized 'Mylar' types are manufactured in values from 4μF at 200V d.c. working to 0·1μF at 2kV. Size of the latter component is only 2 3/16in × ¾in. Metallized 'Mylar' capacitors can also be supplied in round aluminium cans in values up to 100μF at 200V d.c. working.

EE 72 754 *for further details*

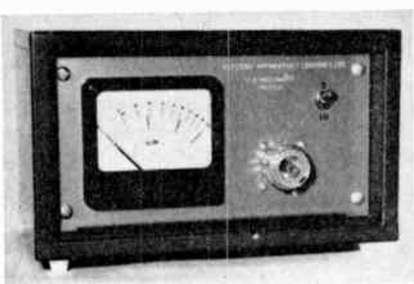
R.F. POWER METER

Electro Apparatus (London) Ltd, Stansted, Essex

(Illustrated below)

This r.f. power meter is designed to meet the needs of those concerned with mobile communications.

It requires no power supply and is equally suitable for laboratory and field use. No setting up is necessary before taking a reading and immediate indication of power is given. Special care is taken in the manufacture of the 50Ω load as well as in the matching of the diode to the coaxial line. In this way it is possible to have a v.s.w.r. of better than 1·1 up to the frequency of



1 000Mc/s. Each instrument is individually calibrated. The overall accuracy of 10 per cent is maintained over the range from 10Mc/s to 1 000Mc/s. 10dB attenuators are available to extend the range to 250mW and 2·5W. Other attenuators for higher powers will be available in the near future.

EE 72 755 *for further details*

TWO-POLE MOTORS

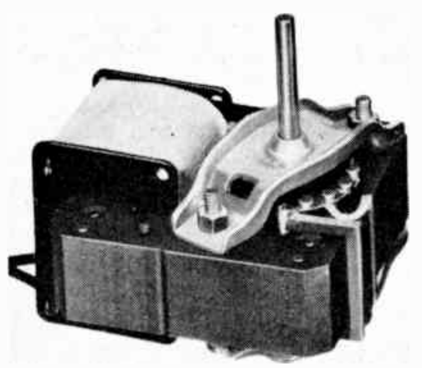
Garrard Engineering Ltd, Swindon, Wiltshire

(Illustrated below)

A new range of two-pole motors, known as the series S20, is now available from Garrard Engineering Ltd.

Design emphasis has been placed on robustness and reliability and new production techniques have been devised to achieve a competitively priced product which is suitable for a wide variety of applications.

Series S20 motors are available in five ratings, covering the range 18 to 40W, which provide starting torques



from 80 to 250g. cm. The motor speeds are dependent on supply frequency, being 2 500 and 3 000rev/min respectively at 50 and 60c/s. Versions are available for clockwise or counter clockwise rotation.

Series S20 motors are each constructed with a single coil wound on a moulded nylon former giving a high degree of insulation and protection.

The insulation is tested at 1·5kV. The motor end plates are formed and pressed aluminium into which self aligning, oil retaining phosphor bronze bearings are assembled.

Quiet and vibration-free running is achieved by a dynamically balanced rotor assembly. The rotor is made from high grade steel, hardened, tempered, ground and super finished. Laminations are of high grade, low loss silicon steel to give maximum efficiency of operation. Protruding shaft length is 1in.

The performance of Garrard Series S20 motors complies with BSS 170.

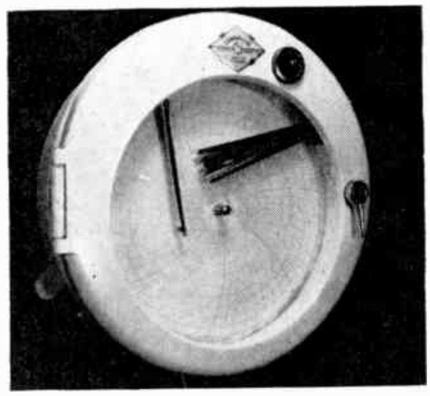
EE 72 756 *for further details*

PROGRAMME CONTROLLER

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Illustrated above right)

The Numalec 'Minor' is virtually a



control panel condensed into one compact, self-contained instrument.

It provides accurate, completely automatic control of temperature, pressure and humidity over long or short periods, and, as each instrument is 'tailor-made' from standard components to meet specific control requirements, it is one of the least expensive yet most versatile programme controllers now available.

The automatic control cycle is started by pressing a button, and at the end of the process the instrument resets itself for an exact repeat programme. New programmes can easily be set by means of the setting arms, adjusted in relation to the scale on the chart face, and by an electric timer, accessible by lifting a corner of the chart. A pre-cut cam can be incorporated to give greater flexibility of control, including rates of rise and fall.

The 10in circular chart provides a continuous record of one or two variables, and as a pen registers a pre-set condition an electric or pneumatic signal is generated, opening or closing solenoid, motorized or pneumatic valves. The variable timed periods—one or two depending on the type of timer used—are similarly initiated, in addition to warning lights and alarms, to provide foolproof, precisely timed programme control. The accuracy of control is directly related to the accuracy of the recorded value, which is better than ±1 per cent.

EE 72 757 *for further details*

STABILIZED POWER UNITS

Newton Brothers (Derby) Ltd, Alfreton Road, Derby

(Illustrated below)

The type KB fully transistorized power supply units have been specially developed to provide a highly stabilized

d.c. supply. They are intended for use in research and development laboratories, test departments, general service workshops and educational establishments. Careful design and construction together with a number of new circuit features have resulted in units having excellent performance specification, coupled with small dimensions and reasonable cost.

Six types are available, covering the range from 15V 200mA to 40V at 1A, with low output impedance. All models are fitted with current limit control and short-circuit protection. Output voltage is continuously adjustable by means of a high resolution potentiometer, with the alternative on some models of a 3 or 10 turn Helipot.

High voltage stability up to ±0·01 per cent is obtained with low noise and ripple.

EE 72 758 *for further details*

VIBRATION METER

Dawe Instruments Ltd, Western Avenue, London, W.1

(Illustrated above)

This vibration meter, the type 1433A, is believed to be the first instrument designed for the measurement of jerk. Calibrated in British or metric units, it gives a direct reading of displacement, velocity, acceleration and jerk. It is portable, self-contained and battery-operated. The instrument operates at frequencies from 1·6c/s to 10kc/s over nine scale ranges.

The system comprises one of two alternative vibration pick-ups, a cathode-follower probe, and the main electronic unit incorporating a high-gain amplifier, integrating and differentiating networks, adjustable attenuator and a direct-reading meter. High stability and calibration accuracy (within ±1 per cent at mid-frequencies) are achieved through negative feedback and stabilized power supply.

Pick-ups are of the accelerometer type. The standard light-weight pick-up weighs 0·7oz, operates over the full frequency range and has a sensitivity of 16mV peak/*g* peak. The high-sensitivity pick-up weighs 8oz and operates from 1·6c/s with a sensitivity of 450mV peak/*g* peak. Resonant frequencies are 115kc/s and 3·8kc/s respectively.

The voltage from the accelerometer is fed into the high-impedance (500mΩ in parallel with 10pF) cathode-follower probe, which is connected to the meter unit by 6ft of three-core screened cable. For jerk measurements, the signal from the probe is first passed through a differentiating circuit, for velocity through an integrating circuit, and for displacement through a double integrating circuit. The required characteristic is read directly by selecting the appropriate switch position.

The operating ranges of the new vibration meter are from 3×10^{-6} to 3×10^{2}in displacement, from 3×10^{-3} to 3×10^{3}in sec, velocity from 3 to 3×10^{4}in/sec^{2} acceleration, and from 30 to 3×10^{6}in/sec^{3} jerk. (The same instrument with metric scales is known as the type 1433B; the ranges are equivalent.)

An output jack allows the amplified signal to be displayed on a c.r.o. or fed into a frequency analyser.

EE 72 759 *for further details*

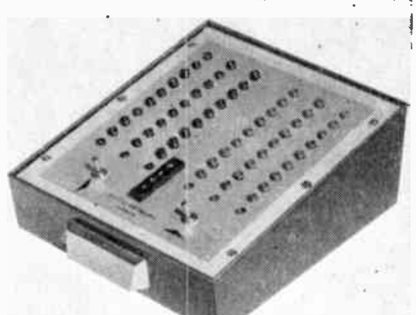
KEYBOARD GRAPH PLOTTER

Scientific Furnishings Ltd, Poynton, Cheshire

(Illustrated below)

Scientific Furnishings Ltd, Electronics Division, has announced the new low-cost model K1 digital keyboard which is used in conjunction with the model XY-1P X-Y Autoplotter (or similar X-Y recorders) to speed up precision graph plotting of tabulated digital data compiled from instrument readings, computers, calculations and experimental and statistical data. The data are simply converted to points on a curve by entering the X and Y co-ordinate on a twin three-digit decade keyboard and pressing the 'plot' bar. The keyboard supplies d.c. voltages to the X and Y axes of the recorder, the magnitudes of which are proportional to the co-ordinates set up on the X and Y key decades with an accuracy of ±0·25 per cent. Suppressed zero and scale expansion permits optimum scaling of the graph and both positive and negative values can be plotted.

In the past the wide use of keyboard conversion of tabulated data to curves has been limited by high cost. This new equipment, costing only about a third of previous units, brings keyboard plotting within reach of the average laboratory.



EE 72 760 *for further details*

PRECISION A.C. DIVIDER

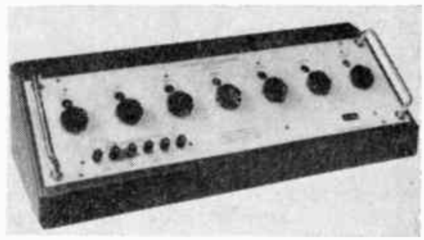
Digital Measurements Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated below)

The precision a.c. divider, DM2101, developed by Digital Measurements Ltd, can be used for the precise measurement and production of a.c. voltage ratios or for the comparison of standards of resistance, inductance, capacitance or ratio. Voltage division with an accuracy and reproducibility of 1 part in 10^{6} can be attained, which is claimed to be an order of magnitude greater than anything previously possible.

The instrument consists of a series of auto-transformers, in which each transformer is divided into 10 equal divisions and any selected division may be further sub-divided by the succeeding transformer. Settings are indicated on instrument dials, or remotely on automatic typewriters or tape punches.

Key to the success of this equipment is an alloy, of exceptionally high re-



liable magnetic permeability, from which the transformer cores are wound. Named 'Supermumetal 100' and produced in the United Kingdom by Telcon Metals Ltd, of Crawley, Sussex, this alloy has the nominal composition 77Ni/14Fe/5Cu/4Mo. It is rigorously tested and controlled for both chemical and physical properties and has an initial permeability greater than 100 000, the maximum exceeding 500 000. 'Supermumetal 100' is vacuum-melted from highly-pure raw materials and rolled to thin foil before being wound into toroidal cores.

EE 72 761 *for further details*

A.C.-D.C. CONVERTOR

The Wayne Kerr Laboratories Ltd, New Malden, Surrey

(Illustrated on page 565)

A new low-priced convertor, designed primarily for use with the Digitec range of d.c. voltmeters, enables a.c. potentials to be measured accurately on any instrument operating from an input of 1V d.c. The Wayne Kerr-Digitec a.c.-d.c. convertor, model 1900, operates over the frequency range 50c/s to 25kc/s with a linearity of ±0·05 per cent full scale. Four input ranges are provided, 0 to 1, 10, 100 and 500V r.m.s. and the input impedance is 1MΩ (100kΩ on 1V range).

The convertor employs a high-gain feedback system to linearize and stabilize the performance of diodes, and the solid-state circuit is designed to maintain the high accuracy over a wide tem-

Leadership in Semiconductors

TEXAS INSTRUMENTS

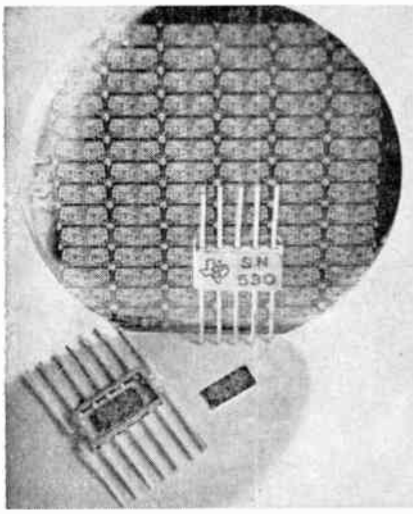
Looking logically at integrated circuits

New high-speed digital semiconductor networks

Series 53 SOLID CIRCUIT* semiconductor networks provide the designer with maximum flexibility through the use of AND/OR INVERT logic with the minimum number of different units.

Features include operation to 5 Mc/s, propogation delay times as low as 5 nanoseconds per AND gate, excellent loading capabilities and the ability to cascade non-inverting logic gates. Series 53 can fulfill most digital requirements in present day computer system applications.

Here is the Series 53 range:
SN530 Single phase, J-K flip-flop
SN531 5-input NAND gate
SN532 5-input AND gate
SN533 Dual 3-input AND gate
SN534 2 and 3-input AND gate
SN535 Clock driver/buffer



Series 53 network, Master Slice wafer*

Master Slice* wafer allows flexibility in mass production

Series 53 as well as TI's established Series 51 and Series 52 networks are made using the Master Slice* fabrication technique. All circuits in a series begin with a slice of silicon containing 50 or more sets of equivalent components.

For Series 53, each component set contains the equivalent of 28 NPN transistors, 10 PNP transistors, 5 capacitors and 26 resistors. Late in the production process, interconnections are made on the Master Slice* to form standard networks, or special networks built to customer specifications. This technique produces maximum network flexibility without sacrificing the economy and consistent reliability of the TI mass production system.

* *Texas Instruments Trademark*

NOW is the time to design semiconductor networks into new equipment

Semiconductor networks may be the answer to your design requirements for reliable, compact and economical circuitry... sooner than you think.

Why now?

Semiconductor networks are penetrating the electronics industry, in Europe as well as in the U.S., even faster than the transistor. Why? Partly because the industry now has the overall experience needed to avoid the sharp transition which we remember when vacuum tube equipment was "transistorized".

Reliability

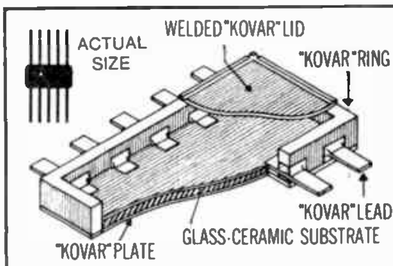
But primarily because experience in transistor reliability evaluation has quickly revealed the remarkable reliability improvements which semiconductor networks will bring. Early tests showed how an integrated circuit containing the equivalent of 20 parts could have the same failure rate as a single conventional transistor. Manufacturers now believe that in a few years a failure rate of around .0001 percent per 1000 hours will be achievable.

Industrial, consumer applications

In addition to the growing military market, industrial and consumer product manufacturers in the U. S. and Europe are moving rapidly to utilize reliable, compact integrated circuits.
For example, Zenith Radio Corporation in the U. S. recently announced a semiconductor hearing aid, developed with TI.

Production, prices

The industry stands just at the beginning of true mass production. Production at TI in 1963 increased eight-fold and more networks were shipped by TI in the fourth quarter of 1963 than the entire industry shipped in the second. And TI is confident that network prices will follow the pattern of steady downward adjustments established by transistors.
In short, the time is *now* for forward looking logic designers to plan for new standards of system reliability and economy with integrated circuits.



Flat package concept becomes standard

In 1959, while most manufacturers were packing more conventional components into transistor cans, TI took a bold short cut and announced the first fully integrated networks in flat packages. It is estimated that 60 to 70% of all integrated circuits sold today are in the flat package configuration.

The flat package provides the most efficient form factor for systems packaging. Shorter bonding leads permit a more rugged construction. Interconnections are easier. And the package is readily welded to printed circuit boards. The package shown above provides a 10 to 1 reduction in size over TO-5 cans when mounted. Total weight is less than 0.1 gram. Package dimensions are 6.4 x 3.2 x 0.9 cm.

Total capability... materials to systems assembly

As semiconductor network applications have developed, there has been a need for related test and assembly equipment. Such equipment is now available from TI for use with networks from TI and other manufacturers. A unique parallel gap welder has been developed to weld networks to printed circuit board. For large users, an in-line tester is available. And special carriers and test jigs assist handling and testing.

Data is now available on this equipment and on the complete line of TI Series 51, 52 and 53 networks. Write for it today

TEXAS INSTRUMENTS
LIMITED
MANTON LANE · BEDFORD · ENGLAND
TELEPHONE: BEDFORD 67466 · CABLES: TEXINLIM BEDFORD · TELEX: 82178

perature range. From 100c/s to 10kc/s the accuracy is ±0·3 per cent full scale (0·4 per cent from 50c/s to 25kc/s). The d.c. output (1V d.c. for 1, 10 or 100V r.m.s., 0·5V d.c. for 500V r.m.s.) is available at low impedance (10kΩ) and the reading time is 1sec.

The unit is available in a portable case to match the Digitec voltmeters, with flanges or for rack-mounting either singly or in association with a voltmeter. The portable unit is 5½in high, 8in wide, 7in deep, and weighs 7lb.

EE 72 762 *for further details*

MULTIPLE SILVERED MICA CAPACITOR BLOCKS

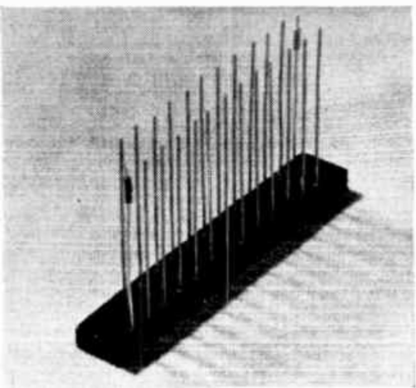
Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London, E.C.1

(Illustrated below)

Johnson, Matthey & Co. Ltd has developed special multiple capacitor blocks for use in the manufacture of delay lines. The capacitors used in the blocks are of identical construction to those in the 'Silver Star' range of precision silvered mica capacitors.

An example of the blocks is shown in the illustration. It is an encapsulated epoxy moulding that measures only 3in × 0·45in × 0·2in and yet contains 23 silvered mica capacitors with capacitance values of 22pF. The capacitors are earthed on one side to a common lead that is brought out at each end of the block. 'Silver Star' fired construction for the capacitors ensures excellent stability and reliability in service.

In this particular size of block the capacitors can have a maximum capacitance of 200pF and are rated at 150V peak. Other block sizes with different numbers of capacitors of different capacitance values can be produced.



These multiple capacitor blocks have been developed to meet demands for components that will speed the assembly of delay lines, and will result in overall savings in space. The leads from the blocks are more easily handled and soldered than leads from individual capacitors, and they are spaced at precise intervals along the block.

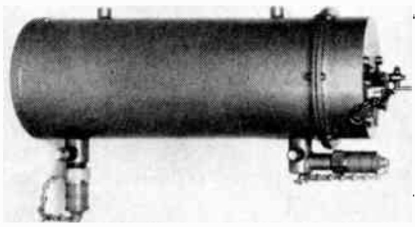
EE 72 763 *for further details*

WAVEGUIDE CALORIMETER

The Marconi Co. Ltd, Chelmsford, Essex

(Illustrated below)

The Specialized Components Division of The Marconi Company announce the development of a new high-power size 12 waveguide calorimeter. It can measure mean powers up to 2kW and employs a new type of ceramic load material that will withstand high temperatures for prolonged periods without damage. With these new calorimeters the method of 'power comparison' is used to ascertain the output power of a transmitter. This technique offers greater accuracy than previous methods which necessitated calculations based on a number of measurements.



The calorimeter covers the frequency range 3·95 to 5·99Gc/s and has a voltage standing wave ratio of less than 1·05 over the whole range. Incorporated in the cooling liquid inlet and outlet ports are diode temperature sensing elements which can be used to indicate the temperature difference between inlet and outlet. In the body of the calorimeter there is a calibrating heater element that is operated from the normal mains supply.

With the r.f. power switched on and a constant flow of coolant through the calorimeter, a meter reading is obtained proportional to the difference in temperature between the inlet and outlet ports. With the r.f. power switched off a calibrating heater is switched on and the mains power feeding it adjusted to produce the same meter reading. This power is equal to the applied r.f. power.

It is possible to operate the calorimeter as a dummy load for short periods without the need for running water supplies. A small header tank holding 1½ pints (0·83 litres) is fitted in place of the water supply pipe to absorb the expansion of the water when hot.

This new design of calorimeter, which is a further development of the high-power water cooled load, overcomes many of the disadvantages of fragility and varying v.s.w.r. associated with earlier types of similar power measuring equipment that employed glass or silica tubes.

EE 72 764 *for further details*

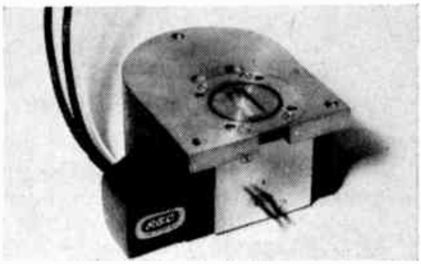
RUGGED MAGNETRON

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Illustrated below)

The latest addition to The M-O Valve Co. Ltd range of rugged reliable magnetrons is an X-band valve with a peak output power of 7kW and a warm-up time of less than 10sec.

This magnetron, type E3094, is capable of withstanding a swept vibration of from 20c/s to 10kc/s with *g* levels commencing at 3*g* and rising to 50*g*. Under these conditions, the output frequency will not change by more than ±3Mc/s. It is also capable of withstanding 40*g* shocks of 10msec duration. Heater supply and high voltage can be switched on simultaneously depending on the type of modulator used. Stability under high rates of rise of voltage at short pulse conditions is better than 0·1 per cent during the first 10sec of operation.



The valve, which features built-in thermocouples for anode temperature measurement, can operate in an ambient temperature range of −65°C to +90°C and at pressures down to 200mm of mercury.

The magnetron measures approximately 4¼ × 4¼ × 2in and weighs 4 lb.

EE 72 765 *for further details*

INDUSTRIAL TACHOMETER

Flight Refuelling Ltd, Wimborne, Dorset

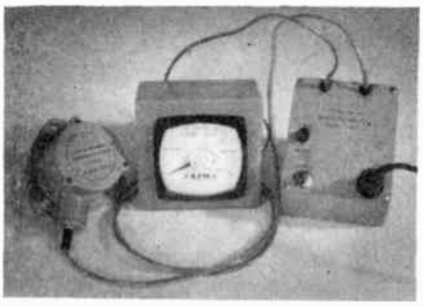
(Illustrated on page 566)

Flight Refuelling Ltd has produced an industrial tachometer which provides an accurate method of measuring speed.

The output is available either as a pulse train, or, with the FR conversion circuit, as a suitable d.c. signal for application to a meter.

The pulse train is derived from the interference of a pick-up head with the sixty magnetic circuits of a rotating tachometer wheel.

The pick-up head and wheel are contained in a robust cast housing for heavy industrial applications, and the assembly is fully sealed to suit most environments. Maximum torque required



to drive the tachometer is less than 1oz. in.

To minimize the risk of component failures, the fully transistorized conversion circuit is contained in a small sealed die cast box, and is protected electrically by careful choice of circuit parameters. D.C. output is proportional to speed to within ±0·2 per cent throughout the complete range of operation.

The tachometer will measure speeds of 3 000rev/min f.s.d. to 15 000rev/min f.s.d. in the ranges recommended by B.S. 3403.

EE 72 766 *for further details*

COUNTING SYSTEM

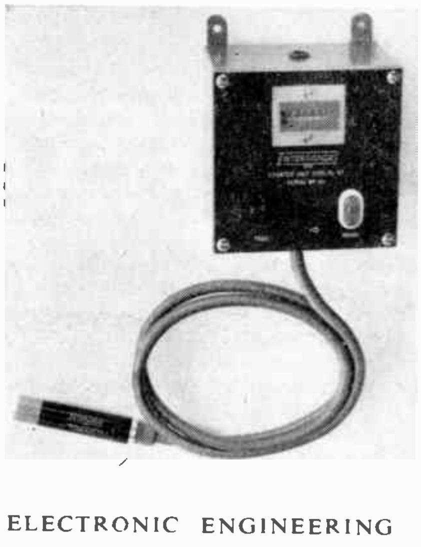
Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustrated below)

To provide a robust and low-priced industrial batch counting system Intersonde Ltd has introduced the type FL87 counter unit for use in conjunction with their series QD inductive proximity switches.

The counter unit takes the form of an easily installed wall mounting steel box measuring 6in × 6in × 4in and containing a manually resettable six-digit electromagnetic counter together with a low voltage d.c. power supply to energize the transistorized proximity switch. The counter unit and the proximity switch are connected by a conventional three core cable which may be up to a maximum of 500ft in length.

The presence of a metal object within 3/8in of the sensing face of the proximity switch causes 24V d.c. to be applied to the electromagnetic counter directly. The maximum counting rate of the complete system is 1 500 per minute although the proximity switch alone will operate at a counting rate of up to 10 000 per minute.



EE 72 767 *for further details*

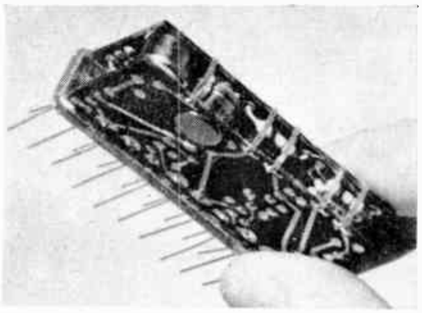
DIGITAL CIRCUIT BLOCKS

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(Illustrated below)

Two new series of digital circuit blocks which are claimed to have an extremely high degree of reliability have been announced by M.E.L. In addition to being encapsulated in a potting compound, the blocks are hermetically sealed in metal cans and the leads are brought out through glass-to-metal seals.

Components are sandwiched between small printed circuit boards. The resulting component density is high and comprehensive functions are performed by individual units. The number of units required for a system is consequently kept to a minimum and design is simplified. Interconnexion and space requirements are also reduced. Maximum dimensions are 53 × 25 × 12·7mm and nineteen connecting wires are standard on all types.



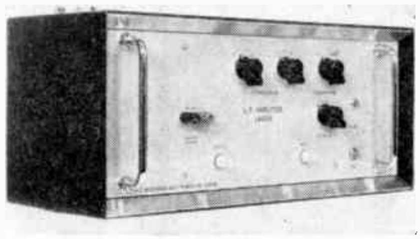
The two series of units are compatible, both employing npn transistors and operating from 12V. The series 10 units use germanium transistors and the maximum permissible operating temperature is 55°C ambient. The basic flip-flop will switch in 3·5μsec.

The series 20 units use silicon epitaxial devices and have a maximum temperature rating of 85°C ambient. The silicon units also have a faster operating speed, the flip-flop being capable of switching in 50nsec. Their higher power rating enables them to be used for driving magnetic stores, the maximum temperature then being 65°C.

Each series is based on 'nand' logic and includes flip-flops, gates, pulse shapers and amplifiers. M.E.L. state that particular care has been taken in the circuit design to prevent spurious operation from switching transients and other kinds of electrical noise that is environments.

The illustration shows one of the new digital circuit blocks before encapsulation and sealing in its metal case.

EE 72 768 *for further details*



L.F. VOLTAGE AMPLIFIER

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustrated above)

This amplifier has been developed for use in conjunction with the Brookdeal phase-sensitive detector PD 629, in measurements involving a modulation frequency of 10c/s to 10kc/s (sine or square wave).

Its very low hum and noise levels, however, coupled with its flat frequency response and absence of phase shift, make it suitable as a general-purpose amplifier for the audio-frequency range.

Between 10c/s and 10kc/s there is negligible phase shift for narrow-band detection purposes (cos α is more than 0·98 at the extremes), and the gain is constant within ±0·25dB. The effects of variation in valve characteristics are eliminated by the application of 20dB of a.c. and d.c. feedback to each stage.

When the feedback is fully in use, the maximum gain of the amplifier is 75dB: in this condition the equivalent noise resistance is about 50kΩ. A 'pre-amplifier' switch gives 20dB more available gain and an equivalent noise resistance of less than 2kΩ. The hum level is very low, being equivalent to about 1μV at the input.

A distributed attenuator enables the gain to be reduced by 55dB in steps of 5dB. A low pass *RC* filter is incorporated with time-constants of 0·05, 0·5 and 5msec.

A stabilized voltage supply (150V at 1mA) is available at terminals on the panel for the energizing of photoconductive detectors etc.

EE 72 769 *for further details*

MULTI-RANGE METERS

Smiths Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated on page 567)

Two versions of this high sensitivity meter are available:-

Type 4S with a sensitivity of 100kΩ/V on d.c. ranges and 20kΩ/V on a.c., has the following ranges; 0-100mV to 0-5 000V d.c.; 0-10μA to 0-1A d.c.; 0-10V to 0-1 000V a.c.; 0-200Ω to 0-500MΩ; 2 000pF to 5μF; -8dB to 50dB and is suitable for use up to 20kc/s.

Type 3S with a sensitivity of 25kΩ/V on d.c. and 2kΩ/V on a.c. is also suitable up to 20kc/s and has ranges from 0-100mV to 0-5 000V d.c. and a.c., 0-5mA and 0-5A d.c. and a.c., 0-200Ω to 0-50MΩ, 100 to 20 000pF and -14dB to 46dB.


Ronald Searle

Wandleside make CABLES

For reliability and efficiency, railway signalling and control systems depend on Wandleside Cables.

Wandleside are specialists in the manufacture of electric wires and cables for all high temperature applications.

In the Television studio you will find Wandleside Cables playing their part in the programme.

Wandleside Cables were used for flight test instrumentation on the first Vickers Armstrong V.C.10.

for every job!

Whatever the application, whatever the industry Wandleside Cables have the right cable for the job. Wherever reliability, efficiency are required. . .whether it's transport, electronics, entertainment or flying. . .Wandleside carry the load. There is a comprehensive range of Wandleside cables to cover the specific requirements of almost every kind of industry. Why not discuss your problem with us.

See us on

STAND NO 273

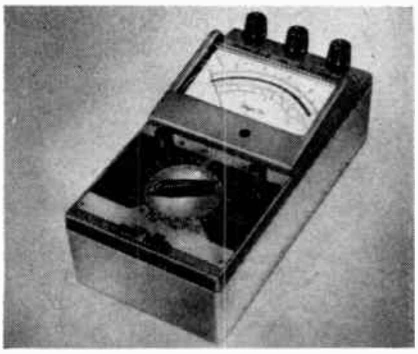
at the Farnborough Air Show

Wandleside Warren Wire Company Limited now hold M.O.A. Design Approval for P.T.F.E. Aircraft Cables.

WANDLESIDE WARREN WIRE COMPANY

In Association with **WANDLESIDE CABLE WORKS LIMITED**

Dunmurry, N. Ireland. Telephone: Dunmurry 2536. Telex: 74339. Telegrams: Derriaghy, Dunmurry, N. Ireland
London Office: 106 Garratt Lane, S.W.18. Telephone: VANdyke 7544. Telex: 261355. Telegrams: Wandleside, London, S.W.18
(Falks Group)



These robust meters have shock-resistant taut ligament suspended movements with glass pointers moving over an inclined scale for ease of readability. A single selector knob allows for ease of operation, and such is the design of the instrument that it remains light and handy to use. The weight of the unit is only 3lb.

The meters are fitted with a patented electrical protection cut-out which acts within 0·005 to 0·01sec, while for protection against high current overloads, a fuse is also provided. Current transformer isolating capacitor and diode rectiers are all in-built, and the meters will measure superimposed a.c. or d.c. signals.

To increase the ranges even further, a wide variety of current transformers, shunts and voltage multipliers are available.

EE 72 770 *for further details*

ELECTRONIC THERMOMETER

Kane-May Ltd, 243 Upper Street, London, N.1

(Illustrated below)

Instantaneous, highly accurate temperature measurements can be carried out with the new 'Dependatherm' electronic thermometer, a pocket size instrument in which advanced circuit techniques are employed to ensure excellent long term stability. Measurements are displayed on a 2½in dial immediately upon application of a small pencil shaped probe to the temperature source, which may be gaseous, liquid or solid.

The rapid response of the new instrument to temperature fluctuations

facilitates measurements hitherto regarded as difficult and permits the display of temperature gradients, localizing of 'hot spots', surface temperature distribution etc, without test rigs of any kind while temperature measurements required in the processing industries are rendered much more convenient. The instrument is equally suitable for production and laboratory applications.

Dependatherm electronic thermometers embody automatic stabilization circuits which obviate any form of pretest balancing (the instrument is immediately ready for use) and which ensure that the accuracy of indication is unaffected, over wide limits, by the gradual decline of the battery voltage.

Single range and multi-range versions are available covering one or more temperature spans within the range of 0° to 200°C or 32° to 350°F. Standard probes of various designs and special probes to meet particular needs in the engineering, chemical and process industries are available. Interconnexion leads between probe and instrument may be up to 50ft long. The instrument measures 4½in × 2½in × 1½in and weighs 7½oz.

EE 72 771 *for further details*

RADIOTELEPHONE

Racal Electronics Ltd, Bracknell, Berkshire

(Illustrated above)

The TRA.355 radiotelephone is a modern compact unit for single sideband communications providing full compatibility with amplitude modulated systems and with the ability to operate using c.w. telegraphy. The nominal power output is 125W p.e.p. under speech conditions. At no frequency is the output less than 100W. The equipment is suitable for operation from a.c. mains or a 12V battery power supply.

The frequency range of the standard version is 3 to 12Mc/s. Extensions of this frequency range up to 15Mc/s and down to 2Mc/s can be carried out at a small additional charge. Four pre-set crystal-controlled channels are available, two lying below 6Mc/s and two above. The unit is designed to operate into a low impedance aerial such as a dipole but, by use of an additional external unit, matching into an open wire aerial can be achieved.

The TRA.355 is entirely self-contained and particular attention has been paid in the design both to keep operation as simple as possible and to ease maintenance problems by employing a fully accessible layout. Additional facilities are available for fitting extended or remote control, voice operated send-receive switching and incorporation into telephone systems. A simple built-in metering system for test and setting-up purposes can be incorporated.

Alternative power supply units for a.c. mains operation or 12V battery supply are mounted on separate sub-chassis which can be fitted quickly in the unit according to requirements.

EE 72 772 *for further details*

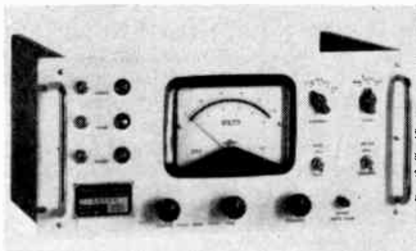
PULSE HEIGHT ANALYSER

Research Electronics Ltd, Bradford Road, Cleckheaton, Yorkshire

(Illustrated below)

An advanced new automatic scanning single channel pulse height analyser is announced by Research Electronics Ltd. Designated the model 9050A the instrument is one of the new 'High Accuracy' range which has been developed to provide every facility and refinement for high grade gamma spectroscopy and pulse height studies.

It is recommended that the analyser be used in conjunction with the new High Accuracy linear pulse amplifier, final details of which will be announced



shortly, and these instruments together will be of particular interest to universities, colleges of technology and other laboratories engaged in teaching and advanced research using isotope techniques.

A special circuit is incorporated giving a 20 to 1 improvement over conventional circuits, and the novel design of discriminator completely eliminates drift from such causes as changing grid base, as it does not rely in any respect upon the cut-off characteristics of multi-electrode valves. The input may be d.c. restored or not at will, to cater for changes in mean value of pulse waveforms with changing p.r.f., according to the application. Scanning (which may be performed manually, automatically, or externally) may be carried out using a fixed channel voltage in the conventional manner, or using a fixed percentage of threshold (sliding channel), giving a constant relative channel width which often gives a better resolution of the lower energy peaks in certain types of work. Threshold level is variable from 2·5V to 100V. Stability is better than 0·01V. Channel width is variable from 0·1V to 50V and 1 per cent of threshold to 2·5 per cent.

EE 72 773 *for further details*

Short News Items

The Fourth International Symposium on Batteries organized by the Inter-Departmental Committee on Batteries is to be held at the Hotel Metropole, Brighton, on 29 to 30 September and 1 October.

Thirty-seven papers, all of which will be published in advance, will be presented covering various aspects of research, development, production and application of battery systems.

Further details are obtainable from F.J.L. Copping—Secretary IDCB Battery Symposium, c/o Ministry of Aviation, Room 413, St. Giles' Court, St. Giles' High Street, London, W.C.2.

The Swiss Industrial Fair announce that the next International Exhibition of Industrial Electronics (INEL) is to be held at Basle on 7 to 11 September 1965.

The Postmaster-General has given his approval to the proposal by the BBC to instal a temporary installation at Sutton Coldfield for the transmission of the BBC 2 Television Service.

The service will be brought into operation by the end of this year, some eight or nine months earlier than the permanent service, and will work on Channel 40, vision 623·35Mc/s, sound 629·25Mc/s with horizontal polarization.

The permanent high power u.h.f. station at Sutton Coldfield will involve a considerable amount of work including the rebuilding of the 750ft mast. In the meantime a temporary aerial on a 150ft mast is to be employed and temporary radio links from London will be provided until the permanent Post Office links are available.

The United Kingdom Atomic Energy Authority has developed a unit (or building block) system of electronics which offers maximum flexibility and simplifies the incorporation of improvements. It is known as the HARWELL 2000 Series Unit System and is manufactured commercially under Authority licence.

Among some 50 types of general-purpose units so far developed are amplifiers, discriminators, single channel analysers, coincidence units, scalers, ratemeters, clockpulse generators, timing units and e.h.t., h.t. and l.t. power units. The majority of these are transistorized and the valve type equipments will be superseded as transistorized units, at present under development, come into production. This range of units provides assemblies which cover most nuclear counting applications in laboratories, both for routine measurement and research work.

The units are available from the following three member firms of the British Scientific Instrument Manufacturers Association:

Dynatron Electronics Ltd, St. Peter's Road, Furze Platt, Maidenhead, Berks.

E.M.I. Electronics Ltd, (Instrument Division), Hayes, Middlesex.

Fleming Instruments Ltd, Caxton Way. Stevenage, Herts.

"Intrinsic Electric Strength and Electromechanical Breakdown of Polythene" by R. A. Fava. B.Sc., (Report No. 5044) has been issued to members of the Electrical Research Association in February 1964 and is now generally released.

Copies of the report priced at 12s 6d, plus postage 6d., are obtainable from the Electrical Research Association, Cleeve Road, Leatherhead, Surrey.

The M.E.L. Equipment Co. Ltd and Telecommunications Radioelectriques et Telephoniques have recently concluded agreements covering technical and commercial co-operation. They have been drawn up with the object of offering a combined source of supply of equipment for the Concord supersonic airliner programme. Each company is capable of producing in its factories equipment developed by either company.

Members of the Agricultural Instruments Group of the Scientific Instrument Manufacturers' Association recently paid a visit to the Glasshouse Crops Research Institute at Rustington, Sussex, at the invitation of the Director of the Institute.

The purpose of this visit was to see the work undertaken by the Institute and the instrumentation problems associated with it. It is hoped to promote collaboration between the instrument and agricultural industries so that each will gain knowledge of the other's requirements and potentialities.

This visit was one of a series being undertaken by the SIMA Agricultural Instruments Group to promote collaboration between the instrument and the various sectors of the agricultural industry.

"Reference Data for Radio Engineers" produced by the International Telephone and Telegraph Corporation is now available in the United Kingdom from Standard Telephones and Cables Limited. This replaces the earlier and smaller volume of the same name produced by STC.

The bulk of the book is a compilation of equations, tables and graphs frequently needed in ratio and electronic engineering. The material has been selected and prepared by a large group of practical engineers, each contributing on his own subject. The result is a concise reference library in one compact volume of 1121 pages with an index.

The book is available from STC Publicity Manager, Therese House, 29-30 Glasshouse Yard, Aldersgate Street, London, E.C.1, or through booksellers, at 42s.

The BBC's television relay station, operated on Channel 5 (vision 66·75Mc/s, sound 63·25Mc/s with vertical polarization) has now been brought into service in the Canterbury area.

The new station will provide improved reception of BBC-1 television for about 30 000 people in Canterbury, where reception of the Crystal Palace station on Channel 1 is at times unsatisfactory, particularly when interference from continental television stations is prevalent.

G.E.C. (Electronics) Ltd is to extend its activities in the field of remote control and instrumentation. This follows the completion of a licensing agreement with Quindar Electronics Inc., of Bloomfield, New Jersey, U.S.A.

Under the agreement, G.E.C. Electronics will manufacture and market Quindar's entire range of remote indication and control equipment in the U.K., Commonwealth and Scandinavia.

This equipment will thus supplement G.E.C. Electronics' existing range of Teledata frequency division multiplex equipment which is already widely used for remote control and supervisory applications.

The Administracion Nacional de Telecommunicaciones of the Republic of El Salvador has awarded a contract to G.E.C. (Telecommunications) Ltd for the supply of 3-circuit and 12-circuit open-wire carrier telephone systems and telegraph equipment.

The contract, which forms the first stage of a new nation-wide telecommunications network in El Salvador, consists of 12 terminals of 3-circuti and 24 terminals of 12-circuit fully transistorized open-wire equipment. The telegraph equipment is for use over the open-wire systems.

The systems, which will be delivered early in 1965, will be installed and commissioned by G.E.C. (Telecommunications) Ltd and manufactured at the Telephone Works in Coventry.

Hawker Siddeley Dynamic's Industrial Electronics Group has received an order from British Railways for electronic equipment for the remote control of railway signals.

The equipment avoids the long runs of cables incorporated in the traditional signalling systems.

Only four small connexion wires at the side of the track are needed by the Hawker Siddeley Dynamics equipment, which is completely self-contained, and uses up-to-date transistor techniques. Up to 250 control or indication signals can be sent in both directions at the same time over a range of ten miles without repeaters.

Teldix of Heidelberg has produced a pocket sized instrument known as 'Teldicord' designed to give an acoustic indication of the heart-beats of an injured and unconscious person, when the heart pulse cannot be felt.

The instrument, which was designed in conjunction with the Physiological Institute of Heidelberg, contains seven transistors in a printed circuit and weighs only 13 oz.

In action, two electrodes are attached to the wrists or ankles of the patient when the heart action currents are made audible, enabling the heart condition to be monitored.

The Silicon Gate-Controlled Switch is the title of a one-day symposium which will be held at the Enfield College of Technology on Wednesday, 2 December, 1964. Four one-hour lectures will be given by staff from J. Lucas, Standard Telephones, Texas Instruments and Transitron. It is believed that this will be the first symposium in this country dealing specifically with the silicon gate-controlled switch. Contributions to the discussion period are invited and there will be good facilities for practical demonstrations.

Full details will be available in October from the Head of the Department of Electrical Engineering, Enfield College of Technology, Queensway, Enfield.

The hydro-electric network in Wales, consisting of three dams, two hydro-power stations and a switching station at Cwm Rheidol is to be supervised entirely by remote control equipment designed, manufactured and installed by the supervisory systems division of G.E.C. (Electronics) Ltd, Coventry.

The control suite at Cwm Rheidol power station forms the nerve centre of the system. It provides for the control and indication of generators, electrical switchgear and water gates; also for the indication of alarms and power, current and voltage levels. Thus the condition of the system can be seen at a glance.

A direct wire control system is used between Cwm Rheidol power station and its nearby dam. In the case of Rhyd Lydan power station, Dinas power station, Dinas dam and Nant-y-moch dam, the distances involved are too great for economic direct wire operation, and in these cases the G.E.C. Electronics' E.M.5 remote control and indication equipment is used.

This is an electromechanical selective control using pulsed code signalling over a 4-wire line. It can provide control and indication facilities for up to 110 two-position devices, and 'indication only' facilities for a further 240 devices when fully equipped.

Automatic telephone facilities are by means of a P.A.X. system incorporated in the control equipment.

Rank Nucleonics and Controls has received an order from Stewart and Lloyds Ltd for a Rank Flatronic electronic belt weighing and control equipment for installation in the Corby Steelworks, the largest of its kind in Europe.

The order covers the complete control system and belt weighing equipment which will monitor and automatically ratio and blend raw materials in the steelworks' 3, 4 and 5 sinter plants. It will be incorporated in the existing conveyor belt system.

Both continuous and static weighing is involved. All instrumentation is sited in two control rooms, where operators will be able to set up the ratio of coke and return fines to the flow rate of iron ore. Vibrating equipment will also be controlled by the equipment to maintain the bed depth of raw mix on each sinter strand.

Outgoing finished sinter will be weighed and the rates, together with the incoming flow rates of iron ore, will be indicated in the control rooms, where operators will be able to keep plant running continuously and keep check on the movement of raw materials.

Nine new orders for Elliott 503 Data Processing Systems, worth approximately £1M, have been received by Elliott-Automation in the past four months. Production of the 503 computer is now running at the rate of two machines a month. New customers for the 503 computer include the Army Operational Research Establishment, the Armament Aircraft Experimental Establishment; the Technion Institute in Haifa, NATO, Italy, The Martin Luther University at Witternberg-Halle and the Finnish Cable Co., Helsinki.

The 503 installation ordered by the Martin Luther University, which is the fifth educational establishment to order a 503 to assist its research programme, will cost over £400 000 and is the largest single 503 system announced to date.

The Israel Institute of Technology—the Technion in Haifa—has ordered an Elliott 503 computer for use in connexion with a number of major research projects. The system ordered includes an additional 16 384 words of core store, over and above the basic 8 192 words of the main processor, and also three magnetic tape handlers and a graph plotter. The computer will be equipped to handle research problems involving very large matrices of information and will be able to produce the results of calculations in graph form. The computer installation at the Technion will be the first one overseas to employ both an Elliott 803 and an Elliott 503 computer.

The 1965 International Solid-State Circuits Conference, the 12th annual meeting, will be held on 17 to 19 February, 1965 in Philadelphia at the University of Pennsylvania and at the Sheraton Hotel.

The conference is sponsored jointly by the University of Pennsylvania and the Institute of Electrical and Electronic Engineers.

Papers are invited on Solid State subjects and authors are requested to submit both a 35-word abstract and a 300-500-word summary of the paper on or before 26 October, 1964 to the secretary of the programme committee Mr. E. J. Lechner, RCA Laboratories, Princeton, New Jersey, U.S.A.

The I.E.E. has announced a new monthly publication 'Current Papers', the first issue of which is appearing this month.

The publication will consist each month of a list of about 700 titles, with authors and references, covering the fields of electrical and electronic engineering that will have been selected from several hundred of the current and more important British and foreign periodicals. Foreign titles will be translated into English and an indication given of the language in which the original is written.

'Current Papers', which has been designed as a service to all electrical and electronics engineers, will supplement the service at present provided by 'Science Abstracts' by offering a speedy reference to current publications.

Voucher copies of the first issue are available on request from the Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2.

Microwave links, worth about £175 000, are to be provided by Standard Telephones and Cables Ltd between London's new 600ft GPO tower and Lille in northern France. This equipment will boost the number of international telephone circuits and also provide a permanent 625-line Eurovision link to replace the BBC's temporary system between Folkestone and London. Equipment to meet the present order will be made at STC's East Kilbride and North Woolwich factories. The company's microwave engineering laboratories are at St. Mary Cray, Kent.

A three-year semiconductor research project to seek experimental confirmation of recent theory on electrical conduction in certain materials on the borderline between insulators and semiconductors is to be carried out by the Electrical Research Association with the support of a DSIR 'earmarked' grant of up to £17 400. The balance of the £20 400 total cost estimated for the project will be provided from the Research Association's funds.

A British Standard for The International System (SI) Units (B.S. 3763) has recently been published, in view of the widespread consideration now being given to the possibility of adopting the metric system in this country.

The SI units were adopted by the Conférence Générale des Poids et Mesures (C.G.P.M.) in 1954 and subsequently endorsed by the International Organization for Standardization. In both these bodies the United Kingdom voted formally in favour of their adoption.

Copies of B.S.3763 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price 6s. each.

Russia is to buy £70 000 worth of vibration test equipment from Pye-Ling Ltd of Royston.

The equipment to be supplied includes control and monitoring units, vibrators, amplifiers and remote control systems. The contract has been negotiated in Moscow by Mr. F. Semark, Director and General Manager of Pye-Ling Ltd.

This latest order brings the company's sales to Russia, during the past four years, to over £800 000.

The Japanese Electrical & Radio Importers' Association of 103 Kingsway, London, W.C.2, has been formed and will be known under the initials of J.E.R.A.

The purpose of this association is to promote the orderly marketing of electrical and electronic products and to guarantee the reliability, quality and servicing of the products handled by the members of the association, and by advertising these aims to both consumer and trade.

An HPE Contourmatic milling machine fitted with Ferranti Mk. IV continuous path control equipment and installed at the Colnbrook toolmaking firm of Smiths Jig and Tool Ltd is currently producing turbine blade aerofoil forms of high quality. Although this machine has been installed for less than a year, Smiths have already built up sufficient expertise in programming and operating numerically controlled machines to compete successfully with firms using conventional methods of producing one-off tooling jobs in turbine and compressor blade production.

The Annual Report of the Scientific Instrument Manufacturers' Association for the year 1963-1964 has recently been published. The Report has been considerably amplified in order that the activities and progress of the Association may be recorded in greater details, and an indication be given of the prospects and problems of the instrument industry as they appear at the present time.

During 1963 production of the instrument industry rose by 10 per cent to a total in excess of £150M. Of the various sectors of the industry most showed increases in output, the largest being in deliveries of industrial process measuring and control instruments. These increased from £45·8M in 1962 to £53·7M in 1963.

An electronic swim timing assembly, 'ESTA', has been installed at the new swimming hall at the Crystal Palace National Recreation Centre, which was officially opened by H.R.H. The Duke of Edinburgh in July.

Designed to L.C.C. specifications, the assembly enables times to be displayed to one-thousandth of a second for each of eight swimming lanes with an additional 'order of finish' display, the times being simultaneously transcribed by print-out equipment incorporated in the console.

Venner Electronics Limited supplied the complete timing equipment including the read-out and print-out facilities.

An electronic belt weighing and control equipment made by Rank Nucleonics and Controls of Welwyn Garden City, Hertfordshire, is helping the drive to modernize Britain's coal industry.

At the National Coal Board's Bestwood Colliery, Nottingham, Rank Flotronic equipment has just been installed to overcome a major problem involved in the washing of coal and the disposal of stone refuse.

This electronic weighing device has been fitted to the conveyor which carries the stone refuse from the washbox, and should the weight of stone per hour rise above a predetermined level, the Flotronic automatically limits the supply of coal and stone being fed from the mine.

PUBLICATIONS RECEIVED

MULLARD SEMICONDUCTOR DESIGNERS GUIDE, 1964 edition, is now available from the company. The guide provides a quick reference to Mullard semiconductor devices for industrial and communication equipment. Three Quick Find Charts for transistors have been introduced which list the devices under the main headings of collector voltage, total dissipation and cut-off frequency. Requests for copies should be made on company headed notepaper to the following address: Central Enquiry Handling, Mullard Limited, Mullard House, Torrington Place, London, W.C.1.

AIRMAX FANS is the title of a new abridged catalogue covering the complete range of Airmax Fans. Copies of this catalogue may be obtained on request to A. K. Fans Limited, 20 Upper Park Road, London, N.W.3.

PHOTODIODES AND PHOTOTRANSISTORS is the title of a booklet recently published by Mullard Ltd which describes the characteristics of photoconductive diodes, photovoltaic diodes, and phototransistors. Circuits are given, which illustrate typical applications of these devices. They include an infra-red relay alarm using an OCP71 phototransistor and a read-out circuit using the BPY10 photodiode. A section on the photo-electric effect and basic illumination theory is also included. Requests for copies should be made to: Mullard Ltd, Mullard House, Torrington Place, London, W.C.1.

ALCAN ALUMINIUM IN THE ELECTRICAL INDUSTRY is the title of a publication which is intended as a convenient summary for the electrical engineer. It gives brief details of the Alcan aluminium alloys used in electrical engineering and the various forms in which these are normally available. It also lists the company's other relevant publications and generally describes the service that Alcan Industries offers in the electrical field.

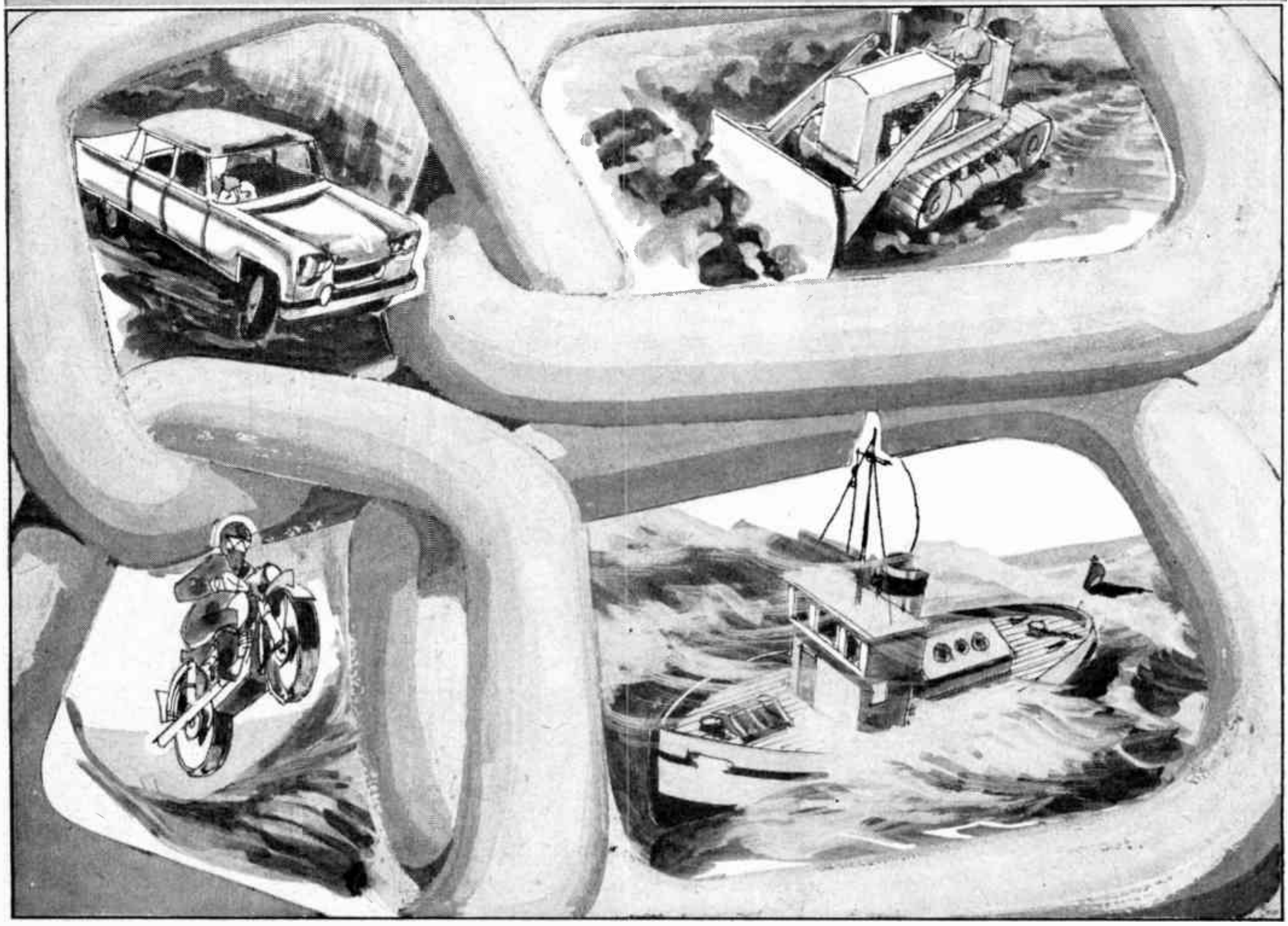
WESTINGHOUSE BRAKE AND SIGNAL COMPANY LTD have recently issued their latest publication relating to rectifiers. Three of these, publications EP.25-3, EP.25-6 and EP.25-8, describe silicon diodes from the large range now available; the last is a completely new range of low power diodes, type SxGR2. Copies of these publications are available on request to the Publicity Department, Rectifier Division, Westinghouse Brake and Signal Ltd., 82, York Way, King's Cross, London, N.1.

APPLICATION NOTES FOR COLD CATHODE INDICATOR TUBES, Publication MS/126, is a new 24-page booklet from the STC Valve Division. Alphabetical and numerical indicator tubes are described and these include end-viewing and side-viewing devices. Most of the booklet is devoted to useful information for the control of indicator tubes and many circuit diagrams are given. Copies of the booklet may be obtained from STC Valve Division, Footscray, Kent.

CSF—Compagnie Generale de Telegraphie Sans Fil have recently issued a number of pamphlets, in English, on their equipment. They are as follows: RF520, Radiotelegraph Receiver Rack; TM112, Mobile Television Repeater; SM100A and SM305, Mass Spectrometer; Differential D.C. Amplifier; Optically Pumped Self-Oscillator Magnetometer; Gammatrol; Humidity-Meter, Density-Meter. Copies of these pamphlets can be obtained on request to Mr. J. G. David, CSF, 79 Bvd Haussmann, Paris 8, France.

EXIDE BATTERIES FOR LIGHT ELECTRICAL AND ELECTRONIC APPLICATIONS is the title of a new booklet published by Chloride Batteries Ltd. The publication reflects the development of transistorized circuiting and other electronic devices, often operating under abnormal conditions, which has called for the design of special batteries. Technical data of 71 dry batteries and 14 portable lead-acid types are clearly set out in see-at-a-glance tables to enable designers to select quickly the right battery for their purpose. Copies of the 8-page booklet can be obtained from Chloride Batteries Ltd, 50 Grosvenor Gardens, London, S.W.1.

COSSOR COMMANDO
COMMUNICATIONS LINK



CC.300 MOBILE 15 WATT V.H.F. F.M. RADIOTELEPHONE

Special Features:

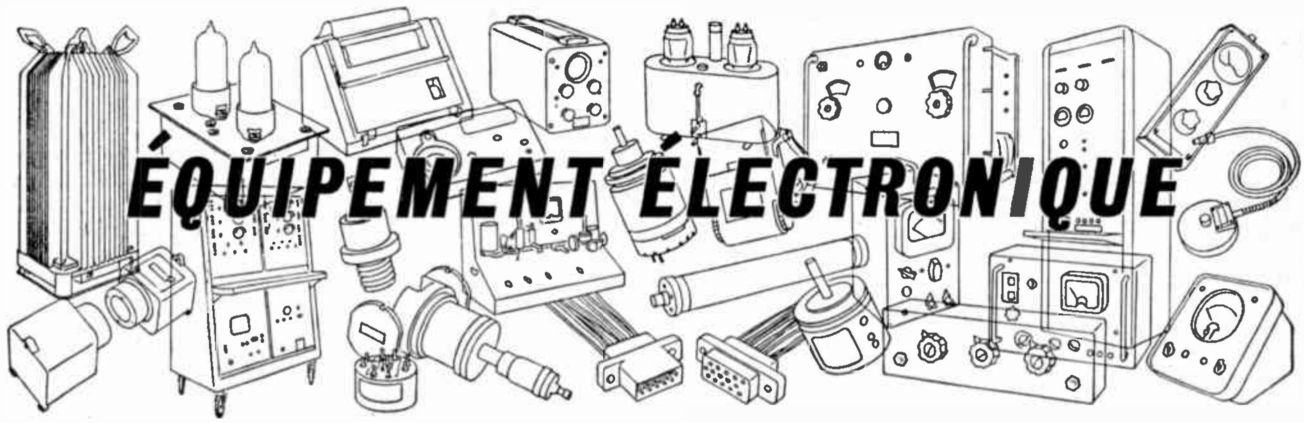
* **Choice of 6 channels**
* **15 watts Power Output**
* **Frequency Range 35 – 175 Mc/s**
* **Marine version available**

The Type CC.300 radiotelephone is a V.H.F. F.M. mobile equipment specifically designed to meet the severe environmental conditions encountered in an equipment designed for use on motorcycles. The CC.300 is completely waterproof and is recommended for applications such as locomotives, fork lift trucks, cranes, fire tenders, agricultural and earth moving appliances.
Ancillary items include an Extended Control Unit which may be separately mounted in a position convenient to the operator.

COMMANDO RANGE
COMMUNICATION– THE KEY TO EFFICIENCY

The Key to the success of Cossor Communications System Planning is undoubtedly the Commando Range of Radiotelephone equipment. This complete range of highly developed equipments employing the latest circuit techniques utilises a new modular system of construction unique in the field of radiotelephony.
All small signal applications are confined to a number of printed circuit boards which may be removed for servicing or replacement in a matter of moments.
In addition great attention has been paid to the mechanical design of all housings and ancillary units so that each equipment in the range is ideally suited for the intended operational environment.

COSSOR COMMUNICATIONS CO. LTD. **THE PINNACLES, HARLOW, ESSEX TEL: HARLOW 26862**
(a Subsidiary of A. C. COSSOR LTD. and RAYTHEON CO., U.S.A.)



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 562 à 567*

ENREGISTREUR GALVANOMÉTRIQUE

Distributeurs: Claude Lyons Ltd, 76 Old Hall Street, Liverpool 3

(*Illustration à la page 562*)

L'enregistreur galvanométrique type EG fabriqué par l'Association des Ouvriers en instruments de Précision (A.O.I.P.) de Paris, est un instrument à bande de 20 cm de largeur, de construction extrêmement robuste et d'une grande fiabilité.

Il comporte un galvonomètre sensible à miroir antivibratoire, le système d'asservissement consistant en un dispositif suiveur de spot breveté comprenant deux cellules photorésistives qui franchissent le spot lorsqu'elles sont en équilibre. Ce système a l'avantage d'effectuer la détection automatique du spot lorsqu'il est branché.

Les spécifications électriques prévoient une précision de ±1.5 %, une stabilité de ±0.5 %, une constante de temps de 4 sec et des sensibilités allant jusqu'à 6 μV/mm ou 0,03 μA/mm- Quatorze vitesse d'avance du papier, de 24 cm/min à 1,5 cm/h sont également prévues.

EE 72 751 *pour plus amples renseignements*

CONTRÔLEUR DE RELAIS

Data Laboratories Ltd, 3-7 Hatfields, London, S.E.1

(*Illustration à la page 562*)

Cet appareil a été conçu pour effectuer le contrôle complet de la durée de fonctionnement et de réenclenchement de relais ainsi que le contrôle de la durée acoustique. Il peut en outre vérifier les chocs et vibrations de relais, tels que décrits dans la Spécification militaire DEF.5165. Grâce à cet appareil, on peut aussi contrôler en cours de fonctionnement les données de spécifications de fabricants se rapportant aux susdites caractéristiques. Ces dernières peuvent être vérifiées rapidement et sans erreur.

Pour le contrôle de fonctionnement de relais, ces derniers sont modulés par impulsions à taux contrôle par un multivibrateur à taux variable pouvant être réglé suivant les caractéristiques du relais soumis à l'essai. On peut mesurer simultanément les durées de fonctionnement, de réenclenchement et de rebondissement et ces durées peuvent être réglés séparément. Si l'une de ces caractéristiques dépasse le chiffre fixé, l'écart est indiqué sur la lampe correspondante.

Le temps de fonctionnement est calculé à partir du moment où la bobine est amorcée jusqu'au moment où les contacts normalement ouverts se sont fixés à la position fermée, c'est à dire après la fin de la période de rebondissement. Cette durée peut être réglée suivant les nécessités et si elle est dépassée une lampe indiquera la panne sur ce contact. Une panne est détectée si le contact s'ouvre pour plus de 1 μsec après la fin de la durée spécifiée de fonctionnement.

Lorsque le relais est désamorcé, la mesure de la durée de réenclenchement du moment où la bobine de relais est désamorcée jusqu'à l'instant où le contact normalement fermé se fixe à la position fermée s'effectue dans les mêmes conditions. Cette durée peut être préréglée et si elle est dépassée, une lampe indique la panne sur ce contact.

Pour la mesure du rebondissement de contact, la durée est calculée de l'instant où le contact ouvert se ferme jusqu'à l'instant où il se fixe dans la position fermée. Cette mesure s'effectue sur les contacts normalement ouverts aussi bien que sur les contacts normalement fermés.

Lorsque l'appareil est utilisé pour le contrôle de chocs et de vibrations, on s'assure que les contacts soumis à l'essai ne s'ouvrent pas durant la période de contrôle. Toute discontinuité est décelée et une lampe indiquera la panne.

EE 72 752 *pour plus amples renseignements*

BLOC D'ALIMENTATION RÉGULÉE

Feedback Ltd, Park Road, Crowborough, Sussex

(*Illustration à la page 562*)

Il s'agit d'un nouvel appareil universel pour banc d'essai fonctionnant sur alimentation de 200-240 V ou de 100-120 V, 50-60 Hz.

Il se caractérise par sa souplesse de connexion, dont le degré est tel que les quatre modes suivants peuvent être obtenus à volonté:

(a) Deux alimentations flottantes de 300 V fournissant 150 mA chacune

(b) Un seul bloc à haute tension fournissant +300 V et −300 V à 150 mA

(c) Un seul bloc à haute tension fournissant +600 V ou −600 V à 150 mA

(d) Un seul bloc à haute tension fournissant +300 V et −300 V à 300 mA.

Pour le fonctionnement dans des conditions de charge minima constante, une simple connexion intérieure augmente la charge maxima sur chaque alimentation à 200 mA.

Il y a trois sorties de 6,3 V, dont deux sont flottantes. Elles peuvent fournir jusqu'à 16 A de courant de chauffage. Un grand contrôleur lit le courant à chacune des sorties. La stabilité à long terme par rapport aux variations est supérieure à 300:1 pour des déviations de 15% de la valeur nominale de l'alimentation. L'impédance de sortie est inférieure à 1 Ω et le facteur d'ondulation total est inférieur à 3 mV efficaces.

Les sorties sont prévues soit sur bornes montées sur le panneau frontal soit sur deux douilles Painton à deux directions montées à l'arrière. Elles sont compatibles avec la gamme Feedback d'éléments analogiques et simulateurs.

EE 72 753 *pour plus amples renseignements*

CONDENSATEURS AU POLYESTER

Capacitor Division, Plessey-UK Ltd, Kembrey Street, Swindon, Wiltshire

(*Illustration à la page 563*)

Le polyester, appelé "Melinex" au Royaume-Uni et "Mylar" aux Etats-Unis constitue un exellent matériau diélectrique. Il est d'une force diélectrique élevée, qui se maintient bien aux températures élevées qui caractérisent fréquement le matériel électronique moderne. L'emploi du "Mylar" métallisé permet de fabriquer des condensateurs d'une efficacité de volume très élevée. Ces derniers sont plus petits que les condensateurs de papier métallisés tout en ayant la même tension de travail.

Trois gammes principales de condensateurs au polyester sont maintenant fournies par la société Plessey -UK Ltd. Il s'agit des modèles non protégés au "Melinex," au "Melinex" moulé et au "Mylar" métallisé moulé. En outre, des condensateurs au polyester pour très hautes tensions—jusqu'à 15 kV—et des condensateurs au polycarbonate à facteurs de puissance réduits à des températures allant jusqu'à 150° C peuvent être fabriqués sur commande spéciale.

La gamme de condensateurs "Melinex" non protégée couvre des valeurs normales de 10 pF à 2 μF à une tension continue de travail de 250 V et de 1 μF à 400 V et 0.1 μF à 1 kV. Ces condensateurs sont traités à une température élevée et les extrémités sont scellés à la résine. Leur classification d'humidité est de H3 et la gamme de température s'étend de $-40°$ C à $+125°$ C.

Les condensateurs moulés "Melinex" sont prévus en sept formats couvrant à peu près les mêmes valeurs et les mêmes tensions de travail indiquées précedemment; la classification d'humidité est, cependant, portée au niveau H2, la gamme de température allant de $-40°$ C à $+100°$ C.

Des modèles moulés et métallisés au "Mylar," ayant la même classification d'humidité et la même gamme de température sont fabriqués à des valeurs de 4 μF à 200 V c.c. de travail à 0,1 μF à 2 kV. Ce dernier composant ne mesure que 55,6 mm sur 19,1 mm. Des condensateurs au "Mylar" métallisé peuvent être fournis également en boîtiers ronds d'aluminium à des valeurs atteignant 100 μF à 200 V c.c. de travail.

EE 72 754 *pour plus amples renseignements*

VOLTMÈTRE DE SORTIE HF

Electro Apparatus (London) Ltd. Stansted, Essex

(*Illustration à la page 563*)

Ce voltmètre HF a été conçu pour répondre aux besoins des techniciens des communications mobiles.

Il n'exige aucune alimentation et peut être employé en laboratoire et en campagne. Nul réglage n'est nécessaire avant d'effectuer une lecture et une indication immédiate de la puissance peut être obtenue. Un soin tout particulier a été apporté à la fabrication de la charge de 50 Ω ainsi qu'à l'adaptation de la diode à la ligne coaxiale. On peut ainsi obtenir un rapport d'amplitude de tension supérieur à 1,1 jusqu'à une fréquence de 1000 MHz. Chaque instrument est étalonné individuellement. La précision totale de 10% est maintenue sur toute la gamme de 10 MHz à 1000 MHz. Des atténuateurs de 10dB sont prévus pour pouvoir porter la gamme à 250 mW et 2.5 W. D'autres atténuateurs pour des puissances plus élevées seront livrables prochainement.

EE 72 755 *pour plus amples renseignements*

MOTEURS BIPOLAIRES

Garrard Engineering Ltd. Swindon, Wiltshire

(*Illustration à la page 563*)

Une nouvelle gamme de moteurs bipolaires. appelée série S20 a été réalisée par la société Garrard Engineering Ltd.

Ces moteurs se distinguent par leur robustesse et leur fiabilité. De nouvelles méthodes de production on été conçues pour réaliser un article d'un prix compétitif convenant à une grande variété d'applications.

Les moteurs de la série S20 sont prévus pour cinq puissances nominales allant de 18 à 40 W et assurant des couples de mise en marche de 80 à 250 g. cm. Les vitesses du moteur dépendent de la fréquence d'alimentation, étant de 2500 et de 3000 tours/minute respectivement à 50 et 60 Hz. Il existe des versions à rotation dans le sens des aiguilles d'une montre et dans le sens contraire aux aiguilles d'une montre.

Les moteurs de la série S20 sont tous construits avec une seule bobine enroulée sur une carcasse en nylon moulé assurant un degré élevé d'isolement et de protection.

L'isolement est controlé à 1,5 kV. Les plaques d'extrémité du moteur sont en aluminium formé et pressé dans lequel des paliers en bronze phosphoreux retenant l'huile et à alignement automatique sont assemblés.

Un fonctionnement silencieux et sans vibrations est réalisé par un assemblage de rotor à équilibrage dynamique. Le rotor est en acier de haute qualité, durci, trempé, meulé et poli. Les tôles sont en acier au silicium à faible perte et de haute qualité, garantissant une efficacité maxima de fonctionnement. La longueur de l'arbre en saillie est de 2,5 cm.

Les performances des moteurs Garrard de la série S20 sont conformes à la spécification de la norme britannique 170.

EE 72 756 *pour plus amples renseignements*

CONTRÔLEUR DE PROGRAMME

Cambridge Instrument Co. Ltd.
13 Grosvenor Place, London, S.W.1

(*Illustration à la page 563*)

Le Numalec "Minor" consiste pratiquement en un panneau de contrôle formant un instrument autonome et fort compact.

Il assure le contrôle précis et entièrement automatique de la température, de la pression et de l'humidité pendant des périodes de longue ou de courte durée et, vu que chaque instrument est "fait sur mesure" avec des composants normaux pour répondre à des conditions spécifiques de contrôle, il représente l'un des contrôleur de programme les moins coûteux et les plus souples d'emploi qu'on puisse obtenir actuellement.

Le cycle de contrôle automatique est déclenché en pressant un bouton et, au terme du processus, l'instrument se réenclenche pour un nouveau programme exactement identique. De nouveaux programmes peuvent être facilement mis en oeuvre au moyen des bras de calage, réglés en fonction de l'échelle sur la surface de la bande, ainsi que par une minuterie électrique, à laquelle on a accès en soulevant un coin du diagramme. Une came préfabriquée peut être incorporée à l'apareil pour lui donner une plus grande souplesse de contrôle ainsi que des taux de montée et de chute.

Le diagramme circulaire de 25 cm fournit un enregistrement continu d'un ou de deux variables, et à mesure que la plume enregistre un phénomène préréglé, un signal électrique ou pneumatique est produit, ouvrant ou fermant des valves pneumatiques, motorisées ou à solénoïdes. Les périodes minutées—dont une ou deux dépendent du type de minuterie utilisé—sont déclenchées de la même manière, ainsi que des voyants lumineux et des dispositifs d'alarme, de sorte que le contrôle minuté avec précision s'effectue sans possibilité d'erreur aucune. La précision du contrôle est en fonction directe de la précision de la valeur enregistrée, qui est supérieure à ± 1 %.

EE 72 757 *pour plus amples renseignements*

ALIMENTATIONS STABILISÉES

Newton Brothers (Derby) Ltd. Alfreton Road, Derby

(*Illustration à la page 563*)

Les blocs d'alimentation entièrement transistorisés, type KB, ont été spécialement réalisés pour la fourniture d'un courant continu à haute stabilité. Ils sont prévus à l'usage des laboratoires de recherche et de mise au point, des services d'essai, des ateliers d'entretien général et des établissement d'enseignement. Le soin apporté à la conception et à la construction de ces blocs, ainsi qu'un certain nombre de nouvelles caractéristiques de circuits, en ont fait des appareils d'une excellente performance de spécification, d'un encombrement réduit et d'un prix raisonnable.

Il existe six modèles, couvrant la gamme de 15 V, 200 mA à 40 V, 1 A, avec une faible impédance de sortie. Tous les modèles sont munis d'une commande de limitation du courant et d'un dispositif de protection contre les courts-circuits. La tension de sortie est à réglage continu au moyen d'un potentiomètre à haute résolution. En variante, certains modèles sont munis d'un Helipot à 3 ou 10 tours.

La stabilité de haute tension atteignant $\pm 0{,}01$ % est obtenue avec un facteur de bruit et d'ondulation réduit.

EE 72 758 *pour plus amples renseignements*

ENREGISTREUR DE VIBRATIONS

Dawe Instruments Ltd, Western Avenue, London, W.1

(Illustration à la page 564)

L'enregistreur de vibrations, type 1433A, serait le premier instrument conçu pour la mesure des secousses. Pouvant être étalonné en unités anglaises ou métriques, cet appareil fournit la lecture directe du déplacement, de la vitesse, de l'accélération et des à coups. Il est portatif, autonome et fonctionne sur batterie. Il peut être utilisé à des fréquences de 1,6 Hz à 10 kHz sur 9 gammes d'échelle.

Il comprend un ou deux capteurs de vibrations, une sonde à charge cathodique et le principal élément électronique comporte un amplificateur à gain élevé, des circuits intégrateurs ou différentiateurs, un atténuateur réglable et un instrument de mesure à lecture directe. La haute stabilité et la précision de l'étalonnage (±1 % aux fréquences moyennes, sont obtenues grâce à l'alimentation stabilisée à contre réaction.

Les capteurs sont du type à accéléromètres. Le capteur léger standard ne pèse que quelques grammes, fonctionne sur toute la gamme de fréquence et a une sensibilité de 16 mV crête/g crête. Le capteur à haute sensibilité pèse 242 g. et fonctionne sur 1.6 Hz avec une sensibilité de 450 mV crête/g crête. Les fréquences de résonance sont de 115 yHz et 3,8 kHz respectivement.

La tension de l'accéleromètre est injectée à la sonde à charge cathodique à haute impédance (500 mΩ en parallèle avec 10 pF). Cette dernière est reliée à l'instrument de mesure par un cable blindé à trois âmes de 182 cm de longueur. Pour la mesure des à coups, on fait d'abord passer le signal de la sonde à travers un circuit différentiateur; Pour la mesure de la vitesse, on fait passer ce signal par un circuit intégrateur et pour la mesure du déplacement à travers un double circuit intégrateur. La caractéristique voulue peut être lue directement en choisissant la position de commutation appropriée.

Les gammes de fonctionnement du nouvel enregistreur de vibrations vont de 3×10^{-6} à 3×10^{2} pouces pour le déplacement, de 3×10^{-3} à 3×10 pouces sec pour la vitesse, de 3 à $3\ 10^{4}$ pouces/sec² pour l'accélération et de 30 à 3×10^{6} pouces/sec³ pour les secousses. (le même instrument avec échelle métrique porte la référence 1433B; les gammes sont les mêmes).

Un jack de sortie permet d'afficher le signal amplifié sur un tube cathodique ou de l'injecter à un analyseur de fréquence.

EE 72 759 *pour plus amples renseignements*

APPAREIL DE REPORT À GRAPHIQUE ET À CLAVIER

Scientific Furnishings Ltd, Poynton, Cheshire

(Illustration à la page 564)

La division d'électronique de la société Scientific Furnishings Ltd. vient d'annoncer la réalisation du nouveau clavier numérique, modèle K1, à prix réduit, utilisé en liaison avec l'appareil de report automatique, modèle XY-1P X-Y (ou avec d'autres enregistreurs analogues) afin d'accélérer le report graphique de précision de données numériques sur table, compilées à partir de lectures d'instruments, de calculs et de données statistiques et expérimentales. Les données sont simplement converties en points sur une courbe en portant les valeurs de co-ordonnées X et Y sur un clavier à double décade à trois chiffres et en pressant la barre de report. Le clavier fournit des tensions continues aux axes X et Y de l'enregistreur, dont les grandeurs sont proportionnelles aux co-ordonnées fixées sur les décades à clef X et Y avec une précision de ±0,25%. La suppression du zéro et l'élargissement de l'échelle permettent l'etalonnage optimum du graphique et on peut reporter des valeurs positives et négatives.

Jusqu'ici l'emploi étendu de la conversion par clavier de données sur table en courbes a été limitée par son coût élevé. Le nouvel appareil, ne coûtant qu'un tiers du prix des appareils précédents, met le report sur clavier à la portée du laboratoire moyen.

EE 72 760 *pour plus amples renseignements*

DIVISEUR DE PRÉCISION DU COURANT ALTERNATIF

Digital Measurements Ltd, 25 Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustration à la page 564)

Le diviseur à courant continu de précision DM1201, réalisé par la société Digital Measurements Ltd est destiné à la mesure de précision et à la production de rapports de tension alternative ou à la vérification d'étalons de résistance, d'inductance, de capacité ou de rapport. Il permet d'obtenir la division de tension avec une précision et une reproductibilité de 1 partie dans 10^{6}, ce qui constitue un ordre de grandeur supérieur à ceux qui ont pu être atteints jusqu'ici.

L'appareil se compose d'une série d'auto-transformateurs, dont chacun d'eux est divisé en 10 divisions égales, toute division choisie pouvant être subdivisée par le transformateur suivant. Les repères de calage sont indiqués sur des cadrans, ou à distance sur machines à écrire automatiques ou perforateurs à bande.

Le succès de cet appareil est dû à l'alliage, d'une perméabilité magnétique particulièrement élevée, dont les noyaux de transformateur sont constitués. Cet alliage, dont la composition nominale est 77Ni/14Fe/5Cu/4Mo, porte le nom de "Supermumetal 100" et il est produit au Royaume-Uni par Telcon Metals Ltd., de Crawley, Sussex. Ses propriétés chimiques et physiques sont rigoureusement contrôlées, sa perméabilité initiale étant supérieure à 100 000 et dépassant 500 000 au maximum. Le "Supermumetal 100" est fondu au vide à partir de matières premières d'une grande pureté et il est laminé en feuilles minces avant d'être enroulé en noyaux toroïdaux.

EE 72 761 *pour plus amples renseignements*

CONVERTISSEUR C.A./C.C.

The Wayne Kerr Laboratories Ltd, New Malden, Surrey

(Illustration à la page 565)

Le nouveau convertisseur Wayne Kerr-Digitec, modèle 1900, est un appareil d'un prix modique, essentiellement prévu pour l'emploi avec la gamme de voltmètres à c.c. Digitec et permettant de mesurer avec précision les courants alternatifs sur tout instrument fonctionnant à partir d'une entrée de 1V c.c. Il est utilisé dans la gamme de fréquence de 50 Hz à 25 kHz avec une linéarité de ±0,05% sur la totalité de l'échelle. Quatre gammes d'entrée sont prévues, soit 0 à 1, 10, 100 et 500 V efficaces et l'impédance d'entrée est de 1 MΩ (100 kΩ dans la gamme de 1 V).

Le convertisseur comporte un système à réaction à gain élevé pour linéariser et stabiliser la performance de diodes. Le circuit constitué de corps solides est conçu pour maintenir la précision élevée dans une gamme étendue de températures. De 100 Hz à 10 kHz, la précision est de ±0,3% sur la totalité de l'échelle (0,4% de 50 Hz à 25 kHz). La sortie de courant continu (1 V c.c. pour 1, 10 ou 100 V efficaces, 0,5 V cc. pour 500 V efficaces peut être obtenue à une faible impédance (10 kΩ) et la durée de lecture est de 1 sec.

Il est livrable dans un coffret protatif s'harmonisant avec les voltmètres Digitec ou, en variante, avec flasques pour montage sur bâti soit indépendamment soit en liaison avec un voltmètre. L'appareil portatif mesure 14 cm de hauteur sur 20,32 cm de largeur sur 17,78 cm de profondeur et son poids est de 3,17 kg.

EE 72 762 *pour plus amples renseignements*

BLOCS MULTIPLES DE CONDENSATEURS AU MICA D'ARGENT

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London, E.C.1

(Illustration à la page 565)

La société Johnson, Matthey & Co. Ltd a réalisé des blocs multiples spéciaux de condensateurs à l'usage de la fabrication de lignes de transmission à retard. Les condensateurs employés dans ces blocs sont de construction identique à ceux de la gamme "Silver Star" de condensateurs au mica argenté de précision.

Un de ces blocs figure dans notre illustration. C'est un moulage d'époxyde encapsulé ne mesurant que 7,62 cm × 1,20 cm × 1,11 cm mais contenant 23 condensateurs au mica argenté d'une valeur de capacitance de 22 pF. Les condensateurs sont reliés à la masse sur un côté par un conducteur ordinaire émergeant à chaque extrémité du bloc. La construction "Silver Star" des condensateurs leur assure une excellente stabilité et une grande fiabilité à l'usage.

Dans ce format de bloc, les condensateurs peuvent avoir une capacitance maxima de 200 pF et leur valeur nominale est de 150 V de crête. D'autres dimensions de blocs avec un nombre différent de condensateurs d'une capacité différente peuvent être produits.

Ces blocs à plusieurs condensateurs ont été mis au point pour répondre à la demande de composants pouvant accélérer l'assemblage de lignes à retard et ils réduisent considérablement les exigences d'espace. Les conducteurs des condensateurs individuels peuvent être manipulés et soudés plus facilement que les conducteurs de condensateurs individuels, et ils sont placés à intervalles précis le long du bloc.

EE 72 763 *pour plus amples renseignements*

CALORIMÈTRE DE GUIDE D'ONDES

The Marconi Co. Ltd, Chelmsford, Essex

(Illustration à la page 565)

La Division des Composants Spéciaux de la société Marconi annonce la mise au point d'un nouveau calorimètre de guide d'ondes de grande puissance, format 12. Cet appareil peut mesurer des puissances moyennes de 2 kW et il comporte un nouveau type de matériau de charge en céramique pouvant résister à des températures élevées pendant des périodes de temps prolongées sans inconvénient aucun. Avec ces nouveaux calorimètres on emploie la méthode de comparaison de puissance pour vérifier la puissance la sortie d'un transmetteur. Cette technique est d'une plus grande précision que les méthodes précédentes qui nécessitaient des calculs basés sur un certain nombre de mesures.

Le calorimètre couvre la gamme de fréquence de 3,95 à 5,99 GHz et son taux d'ondes stationnaires est inférieur à 1,05 sur toute la gamme. Les ouvertures d'entrée et de sortie de liquide de refroidissement comprennent des éléments sensible à la température des diodes qui peuvent être utilisés pour indiquer la différence de température entre l'entrée et la sortie. Le corps du calorimètre comprend un élément chauffant étalonné qu'on met en oeuvre à partir de l'alimentation secteur normale.

Lorsque la puissance HF est branchée et qu'un flux constant de liquide de refroidissement passe à travers le calorimètre, on obtient une lecture proportionnelle à la différence de température entre les ouvertures d'entrée et de sortie. Lorsque, par contre, le courant HF est débranché, cela a pour effet de brancher une tension de chauffage, de sorte que l'alimentation de courant secteur est réglée de manière à produire la même lecture. Cette puissance est égale à la puissance HF appliquée.

On peut utiliser le calorimètre comme charge fictive pendant de courtes périodes sans qu'il soit nécessaire d'avoir recours à des alimentations en eau courante. Un petit réservoir d'une contenance de 0,83 litres remplace le tuyau d'alimentation en eau et sert à absorber la dilatation de l'eau lorsque cette dernière est chaude.

Ce nouveau type de calorimètre, qui constitue un nouveau développement de la charge à eau de refroidissement de grande puissance obvie aux nombreux désavantages de la fragilité et de taux d'ondes stationnaires variables qui caractérisaient des modèles précédents de matériel de mesure de puissance à tubes de verre ou de silice.

EE 72 764 *pour plus amples renseignements*

MAGNÉTRON ROBUSTE

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Illustration à la page 565)

Le dernier-né de la gamme des magnétrons robustes et sûrs produits par The M-O Valve Co. Ltd est un tube à bande X avec une puissance de sortie de pointe de 7 kW et un temps de chauffage de moins de 10 sec.

Le magnétron type E3094 peut résister à une vibration de balayage de 20 Hz à 10 kHz avec des niveaux de g commençant à 3 g et montant à 50 g. Dans ces conditions, la fréquence de sortie ne change pas de plus de ± 3 MHz. Il peut également résister à des chocs de 40 g d'une durée de 10 msec. Le chauffage et la haute tension peuvent être branchés simultanément selon le type de modulateur utilisé. La stabilité dans des conditions de haute montée de tension à courtes impulsions est supérieure à 0,1% durant les dix premières secondes de fonctionnement.

Il comprend des thermocouples incorporés pour la mesure de la température anodique, peut être utilisé dans une gamme de température ambiante de $-65°C$ à $+90°C$ et à une pression minima de 200 mm de mercure.

Il mesure environ 12 cm × 10,79 cm × 5,08 cm et son poids est de 0,90 kg.

EE 72 765 *pour plus amples renseignements*

TACHYMÈTRE INDUSTRIEL

Flight Refuelling Ltd, Wimborne, Dorset

(Illustration à la page 566)

La société Flight Refuelling Ltd vient de produire un tachymètre industriel qui représente un moyen précis pour mesurer la vitesse.

La sortie est constituée soit par un train d'impulsions soit, avec le circuit de conversion de réponse de fréquence, par un signal de courant continu pouvant être appliqué à un instrument de mesure.

Le train d'impulsions émane de l'interférence d'une tête de pick-up avec les soixante circuits magnétiques d'une roue tournante de tachymètre.

La tête et la roue du pick-up sont enfermées dans un robuste coffret propre aux applications industrielles lourdes, et l'ensemble est entièrement scellé et protégé contre toutes les conditions d'environnement. Le couple maximum requis pour entraîner le tachymètre est inférieur à 283,5 g/2,5 cm.

Pour réduire tout risque de panne des composants, le circuit de conversion entièrement transistorisé est enfermé dans un petit coffret scellé et moulé en coquille. Il est protégé électriquement par un choix minutieux des paramètres de circuit. La sortie de courant continu est proportionnelle à la vitesse à $\pm 0,2\%$ près pendant toute la gamme de fonctionnement.

La nouveau tachymètre peut mesurer des vitesses de 300 tours/minute de déviation totale à 15 000 tours/minute de déviation totale dans les gammes recommandées par la norme britannique BS 3403.

EE 72 766 *pour plus amples renseignements*

SYSTÈME DE COMPTAGE

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustration à la page 566)

La société Intersonde Ltd a créé un système de comptage par lots qui est à la fois robuste et d'un prix modique. Il peut être utilisé en liaison avec la série Intersonde de commutateurs de proximité QD.

Le compteur FL87 se présente sous la forme d'un boîtier d'acier à montage mural mesurant 15,24 × 15,24 × 10,16 cm et contenant un compteur électromagnétique à six chiffres pouvant être réenclenché manuellement, ainsi qu'un bloc d'alimentation en tension continue basse servant à exciter le commutateur de proximité transistorisé. Le compteur et le commutateur de proximité sont reliés par un câble conventionnel à trois âmes dont la longueur maxima peut aller jusqu'à 152 m.

La présence d'un objet métallique à une distance de 0,95 cm de la face sensible du commutateur de proximité provoque l'application directe d'un courant continu de 24 V au compteur électromagnétique. La vitesse de comptage maxima du système complet est de 1500 coups par minute bien que le commutateur de proximité à lui seul fonctionne à une vitesse de comptage de 10 000 coups par minute.

EE 72 767 *pour plus amples renseignements*

BLOCS DE CIRCUITS NUMÉRIQUES

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(Illustration à la page 566)

Deux nouvelles séries de blocs de

circuits numériques d'une très grande fiabilité viennent d'être réalisées par la société M.E.L. Ces blocs sont encapsulés dans une composition et hermétiquement scellés dans des coffrets métalliques. Le conducteurs sortent par des scellements verre-métal.

Les composants sont comprimés entre des petites plaquettes à circuit imprimé. La densité des composants qui en résulte est élevée et des fonctions générales sont exécutées par les éléments individuels. Le nombre des éléments requis pour un système est par conséquent réduit au minimum et la construction simplifiée. Les inter-connexions et les exigences d'espace sont également réduites. Les dimensions minima sont de 53 × 25 × 12.7 mm et les 19 fils métalliques de connexion sont de type standard sur tous les modèles.

Les deux séries d'éléments sont compatibles, toutes deux étant à transistors npn et fonctionnant sur 12 V. Les éléments de la série 10 sont à transistors au germanium et la température ambiante de fonctionnement maximum est de 55° C. Le flip-flop effectue la commutation en 3,5 µsec.

Les éléments de la série 20 utilisent des dispositifs épitaxiaux au silicium et leur gamme de température ambiante maxima est de 85° C. Les éléments au silicium ont une vitesse de fonctionnement supérieure, le flip flop pouvant commuter en 50 nsec. Le régime nominal plus élevé permet de les utiliser pour entraîner des réservoirs magnétiques, la température maxima étant alors de 65° C.

Chaque série est basée sur la logique "nand" et comprend des univibrateurs, des circuits de porte, des correcteurs d'impulsions et des amplificateurs. La société M.E.L. affirme qu'un soin tout particulier à été apporté à la mise au point du circuit afin d'empêcher le fonctionnement erroné par suite de transitoires de commutation et d'autres genres de bruits électriques.

Notre photographie montre l'un de ces nouveaux blocs de circuits numériques avant l'encapsulage et le scellement dans son boîtier métallique.

EE 72 768 *pour plus amples renseignements*

AMPLIFICATEUR DE TENSION BF

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(*Illustration à la page 566*)

Cet amplificateur a été conçu pour être utilisé en liaison avec le détecteur sensible aux phases Brookdeal PD 629, pour la mesure comportant une fréquence de modulation de 10 Hz à 10 kHz (ondes carrées ou sinusoïdales).

En raison, cependant, de son niveau de bruit et de bourdonnement réduit, ainsi que de sa réponse de fréquence linéaire et de son absence de déphasage, il se prête également à l'emploi comme amplificateur universel pour la gamme des fréquences acoustiques.

Entre 10 Hz et 10 kHz, le déphasage est négligeable pour la détection à bande étroite (cos α est supérieur à 0,98 aux extrêmes), et le gain est constant à ±0,25 dB près. Les effets de variations dans les caractéristiques de tubes sont éliminés par l'application de 20 dB de réaction alternative ou continue à chaque étage.

Lorsque la réaction est pleinement utilisée, le gain maximum de l'amplificateur est de 75 dB. A ce niveau, la résistance de bruit équivalente est d'environ 50 kΩ. Un commutateur préamplificateur donne 20 dB de plus de gain disponible et une résistance équivalente de bruit inférieure à 2 kΩ. Le niveau de bourdonnement est très bas, correspondant à peu près à 1 µV à l'entrée.

Un atténuateur réparti permet de réduire le gain de 55 dB en plots de 5 dB. Un filtre passe-bas RC est incorporé et ses constantes de temps sont de 0,05, 0.5 et 5 msec.

Une alimentation de tension stabilisée (150 V à 1 mA) est disponible aux bornes du panneau pour l'amorçage de détecteurs photoconductifs.

EE 72 769 *pour plus amples renseignements*

INSTRUMENTS DE MESURE MULTIGAMMES

Smiths Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(*Illustration à la page 567*)

Il exisîe deux versions de cet instrument de mesure d'une haute sensibilité, à savoir:

le type 4S, dont la sensibilité est de 100 kΩ/V sur les gammes de tension continue et 20 kΩ/V sur les gammes de tension alternative. Ses gammes sont comme suit: 0-100 mV à 0-5000 V c.c., 0-10 µA à 0-1 A c.c., 0-10 V à 0-1000 V c.a., 0-200 Ω à 0-500 MΩ, 2000 pF à 5 µF, −8 dB à 50 dB, et il peut être utilise jusqu'à 20 kHz,

le type 3S, dont la sensibilité est de 25 kHz/V sur tension continue et de 2 kΩ/V sur tension alternative. Il peut également être utilisé jusqu'à 20 kHz et ses gammes sont les suivantes: 0-100 mV à 0-5000 V c.c. et c.a., 0-5 mA et 0-5 A c.c. et c.a., 0-200 Ω à 0-50 MΩ, 100 à 20 000 pF et −14 dB à 46 dB.

Ces robustes instruments de mesure comportent des mouvements antichocs suspendus à ligament tendu et des aiguilles en verre se déplaçant sur une échelle inclinée afin de faciliter la lecture. Un seul sélecteur facilite l'emploi et la conception de l'instrument est telle qu'il demeure léger et facilement maniable. Il ne pèse que 1,5 kg.

Ces appareils sont munis d'un dispositif de coupure breveté assurant la protection électrique. Ce dispositif agit entre 0,005 et 0,01 sec. Un fusible est prévu, en outre, pour la protection contre les surcharges de courant élevées. Les redresseurs isolants à condensateurs et diodes pour la transformation du courant sont tous incorporés et les appareils peuvent mesurer des signaux à tension alternative ou continue superposés.

On peut obtenir, en outre, une variété étendue de transformateurs de courant, de shunts et de multiplicateurs de tension qui permettent d'accroître les gammes.

EE 72 770 *pour plus amples renseignements*

THERMOMÈTRE ÉLECTRONIQUE

Kane-May Ltd, 243 Upper Street, London, N.1

(*Illustration à la page 567*)

On peut effectuer des mesures de température instantanées et d'une grande précision grâce au nouveau thermomètre électronique Dependatherm, qui constitue un instrument de poche pour le lequel on a fait usage de méthodes de circuit avancées afin d'assurer une excellente stabilité à long terme. Les indications sont affichées sur un cadran de 6,35 cm immédiatement après application d'une petite sonde en forme de crayon à la source de température, qui peut être gazeuse, liquide ou solide.

La réponse rapide du nouvel instrument aux fluctuations de température facilite les mesures considérées jusqu'ici comme difficiles et permet l'affichage de gradients de température, la localisation de points chauds, la répartition de la température de surface, etc., sans devoir faire usage d'appareillage de contrôle d'aucune sorte. D'autre part, les mesures de température exigées dans les industries de traitement sont rendues beaucoup plus aisées. L'appareil se prête également aux applications de production et de laboratoire.

Les thermomètres électroniques Dependatherm comprennent des circuits automatiques de stabilisation qui obvient à toute forme d'équilibrage de précontrôle (l'instrument étant prêt à l'emploi immédiat) et qui assurent le maintien de la précision de l'indication dans des limites étendues, sans qu'elle soit affectée par le déclin graduel de la batterie.

Des versions à gamme unique et à plusieurs gammes sont prévues; elles couvrent une ou plusieurs gammes de temperature entre 0° et 200°C ou entre 32° et 350°F. Des sondes normales de différents modèles, ainsi que des sondes spéciales répondant à des usages particuliers des industries chimiques, du traitement et de la construction, peuvent être fournies. Des câbles d'interconnexion entre la sonde et l'instrument peuvent aller jusqu'à 15 m de longueur. L'appareil mesure 11,43 cm × 6,35 cm × 3.81 cm et son poids est de 233 gr.

EE 72 771 *pour plus amples renseignements*

RADIOTÉLÉPHONE

Racal Electronics Ltd, Bracknell, Berkshire

(*Illustration à la page 567*)

Le radiotéléphone TRA.355 est un

élément moderne et compact pour les communications à bande latérale unique. Il est parfaitement compatible avec les systèmes modulés en amplitude et avec la possibilité de pouvoir l'utiliser pour la télégraphie par ondes entretenues. La sortie de puissance nominale est de 125 W de tension de crête au niveau des courants vocaux. La sortie n'est jamais inférieure à 100 W, quelle que soit la fréquence. Il peut être employé sur courant alternatif de secteur ou sur batterie de 12 V.

La gamme de fréquence du modèle standard est de 3 à 12 MHz. Elle peut être étendue, à un maximum de 15 MHz et un minimum de 2 MHz moyennant un léger supplément de prix. Quatre canaux préréglés pilotés au cristal sont également prévus, dont deux sont au-dessous de 6 MHz et deux au-dessus. L'appareil a été étudié pour l'emploi avec une antenne à faible impédance tel qu'un dipôle; cependant, à l'aide d'un élément extérieur supplémentaire, on peut réaliser l'équilibrage avec une antenne à fil ouvert.

Le TRA.355 est un appareil entièrement autonome et un soin tout particulier a été apporté à rendre le fonctionement aussi simple que possible et à simplifier les problèmes d'entretien grâce à un montage pleinement accessible. On peut, en outre, munir l'appareil de dispositifs de télécommande, de commutation de transmission-émission à fonctionnement vocal et d'incorporation à des systèmes téléphoniques. Un système d'enregistrement de contrôle et de montage peut être incorporé sur demande.

Des blocs d'alimentation pour fonctionnement sur tension secteur alternative ou sur batterie de 12 V sont montés sur des sous-châssis séparés et peuvent être fixés rapidement dans l'appareil selon les besoins.

EE 72 772 *pour plus amples renseignements*

ANALYSEUR DE HAUTEUR D'IMPULSION

Research Electronics Ltd, Bradford Road, Cleckheaton, Yorkshire

(Illustration à la page 567)

Un nouvel analyseur de hauteur d'impulsion à voie unique et à balayage automatique a été réalisé par la société Research Electronics Ltd. L'analysateur modèle 9050 A est, en effet, un instrument de grande précision conçu pour faciliter la spectroscopie gamma et l'étude des hauteurs d'impulsions.

Il est recommandé d'utiliser l'analyseur en liaison avec les nouvel amplificateur d'impulsions linéaires de grande précision, dont les caractéristiques finales seront indiquées prochainement. L'ensemble de ces instruments sera d'un intérêt particulier pour les universités, collèges de technologie et laboratoires d'enseignement et de recherches poussées au moyen des méthodes isotopiques.

Un circuit spécial a été incorporé à l'appareil. Il constitue une amélioration de 20 à 1 par rapport aux circuits classiques. La conception inédite du discriminateur élimine entièrement la dérive due a un changement de la base de la grille, le discriminateur ne dépendant en aucune manière de la caractéristique de coupure des tubes à plusieurs électrodes. La puissance d'entrée peut être fournie en tension continue, rétablie ou non à volonté, afin de l'adapter aux changements de valeur moyenne des formes d'onde d'impulsion à fréquence d'impulsions. Selon l'application prévue, le balayage (manuel, automatique ou extérieur) peut être effectué à l'aide d'une tension de voie fixe de la manière classique, ou à l'aide d'un seuil à pourcentage fixe (voie glissante), donnant une largeur relative de voie constante, produisant fréquemment une meilleure résolution des pointes d'énergie inférieure dans certains genres de travaux. Le niveau de seuil varie de 2,5 V à 100 V. La stabilitée est supérieure à 0,01 V. La largeur de voie varie de 0,1 V à 50 V et de 1% de seuil à 2,5.

EE 72 773 *pour plus amples renseignements*

Résumés des Principaux Articles

Circuits de démodulation pour récepteurs de télévision en couleur PAL par W. Bruch

Résumé de l'article aux pages 512 à 518

Cet article décrit certains circuits caractéristiques pour la démodulation de signaux en couluer PAL. Un décodeur à commutation permet de traiter le signal suivant le principe PAL et d'évaluer ce signal par la méthode PAL_{DL}. Le débranchage des circuits PAL permet le fonctionnement aux signaux NTSC. L'article décrit, en outre, un certain nombre de circuits de base ainsi que leur réalisation pratique. Il fournit, enfin, quelques suggestions pour réduire le coût du modulateur de lignes à retard.

Un transistor de commutation économique au silicium et son emploi dans les compteurs annulaires par J. Palmer

Résumé de l'article aux pages 519 à 523

Un compteur en anneau à transistor au silicium a été décrit par Ringrose (1) et Bradley (2). La conception du compteur est analysée en détail et des suggestions sont faites qui permettent de fabriquer une version économique, le rapport de prix étant d'environ une décade du nouveau modèle pour une seule paire de déclenchement de l'ancienne. Les limites de vitesse, etc sont également discutées. Le compteur utilise un transistor au silicium et une diode à bas prix. Un exemple de réalisation est donné. Le compteur peut être utilisé comme compteur simple ou comme un type de commutateur, effectuant une opération ou une série d'opérations le long d'un anneau, à raison d'un degré à la fois.

Un amplificateur linéaire d'auto-accord à 5 kW par J. Wood

Résumé de l'article
aux pages 524 à 529

Cet article décrit un amplificateur linéaire HF d'auto-accord avec une puissance d'enveloppe de pointe de 5 kW. Il a été prévu pour être actionné par un émetteur d'entraînement à puissance réduite fournissant une puissance de sortie de 500W de puissance d'enveloppe de pointe dans la gamme de fréquences de 2,5 à 25MHz. L'amplificateur comprend un système d'asservissement commandé par des discriminateurs permettant le réglage automatique de l'amplificateur à la fréquence voulue dans quelques secondes; le plus long changement d'onde est de 10 secondes avec des changements de voie moyens de 5 secondes.

Un modulateur de phase simple par C. T. Kohn

Résumé de l'article
aux pages 530 à 533

Cet article analyse un modulateur de phase simple consistant en trois réactances et un tube. Des conditions de fonctionnement maxima sont obtenues et les effets d'une caractéristique de tube non linéaire sont discutés. Dans une réalisation pratique, une distorsion de fréquence de 10% a pu être obtenue avec une déviation de 48°.

L'élément d'emmagasinage CIRRUS à noyau de courant coïncident par I. R. Butcher

Résumé de l'article
aux pages 534 à 539

Un réservoir à noyau de courant coïncident, d'une capacité de 16.384.19 mots binaires et un cycle moyen de 6µsec, est décrit dans cet article. Le but de la conception était de réaliser un réservoir sûr d'un prix modique plutôt qu'un élément à action rapide. Les circuits de sélection X et Y sont basés sur la commutation à répartition de charge qui assure un usage efficace mais modéré des transistors d'entraînement. L'utilisation dans une gamme de température de 10 à 50°C avec des marges étendues de fonctionnement est rendue possible grace à la compensation des transistors d'entraînement.

Un stabilisateur de tension pour circuits à cathode froide par A. J. Oxley

Résumé de l'article
aux pages 548 à 549

L'auteur décrit un tube de déclenchement à cathode froide utilisé dans les circuits stabilisateurs pour des sorties de 150 à 1 000V. Un de ces modèles, prévu pour l'alimentation des Dékatrons et des tubes de déclenchement, a produit jusqu'a 20mA à 470V, avec une impédance de sortie de 250Ω et un rapport de stabilisation de 35: 1.

Sous-étalons de tension et de courant par M. Pacak

Résumé de l'article
aux pages 550 à 553

Cet article décrit un stabilisateur électronique de tension ou de courant équipé d'un amplificateur de grande efficacité et de plusieurs autres dispositifs assurant le réglage précis et reproductible du volume de la puissance de sortie. Cette dernière est déterminée presque exclusivement par un diviseur de référence et par une tension de référence. On peut effectuer le réglage continu ou par plots sur toute la gamme d'environ 1 000V ou 1A à zéro ou au-dessous (polarité inversée). La stabilité et une précision de calage de 10^{-4} ou davantage permettent d'utiliser l'instrument comme échelle de courant ou de tension pour le report de données sur diagramme, l'étalonnage d'instruments, etc.

Un régulateur transistorisé de haute tension à commande alternative par I. Izumi et M. Kokubu

Résumé de l'article
aux pages 554 à 556

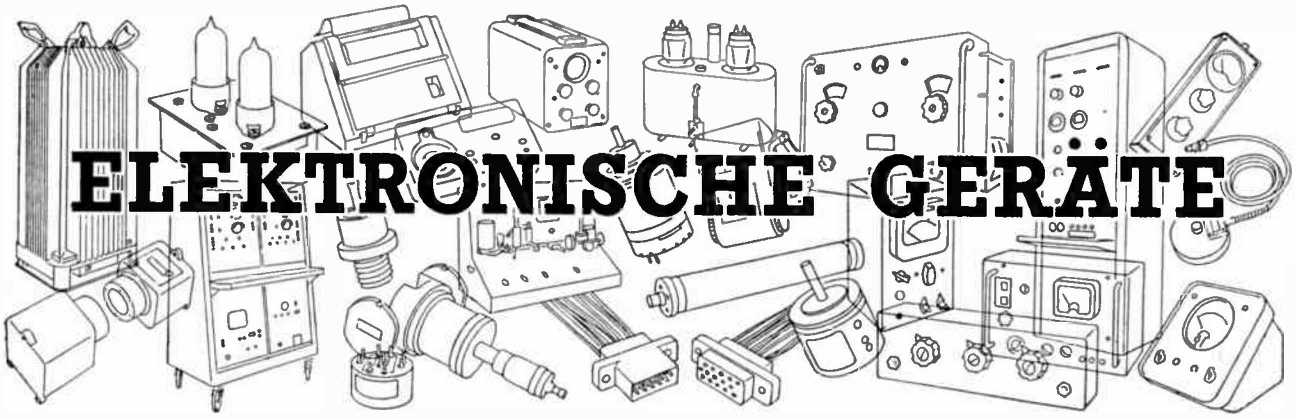
La méthode simple et inédite dont il est question ici permet de régler le niveau de basse tension alternative et de le convertir en une sortie de haute tension à l'aide d'une impédance en série avec un primaire de transformation. Le dispositif est analogue à un régulateur du type convertisseur de courant continue en courant alternatif, ou à un régulateur du type à amplificateur magnétique, mais il n'utilise aucun convertisseur de courant continu en courant alternatif et sa durée de rétablissement n'est pas limitée par les oscillations ou la fréquence de ligne.

On décrit, en outre, un circuit pratique pour 200-800V, 0-5mA, soit un rapport de stabilisation de l'ordre de 100, du courant continu à plusieurs centaines de cps.

Une ligne à retard binaire à tube à cathode froide par D. Q. Mayne

Résumé de l'article
aux page 557

L'auteur décrit une ligne à retard binaire simple à tube à cathode froide. Cette ligne utilise deux alimentations commutées qui permettent de transférer un signal binaire d'une ligne A à une ligne B et, après un retard d'une unité, de le renvoyer à la ligne A.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Ubersetzung der Seiten 562 bis 567

Galvanometer-Schreiber

Vertrieb Claude Lyons Ltd, 76 Old Hall Street, Liverpool 3

(Abbildung Seite 562)

Der von Association des Ouvriers en Instruments de Precision (A.O.P.I.), Paris, hergestellte Galvanometer-Schreiber EG hat eine Streifenbreite von 203 mm und ist ein äusserst robustes und zuverlässiges Instrument.

Der Schreiber hat ein empfindliches, gefedertes Spiegelgalvanometer; das Servosystem hat eine patentierte Punktfolgeanordnung mit zwei Fotowiderständen, die bei Nullabgleich auf den Punkt einschleifen. Dieses System hat den Vorteil, beim Einschalten des Gerätes automatisch nach dem Punkt zu suchen.

Unter den elektrischen Kenndaten sind zu nennen: Genauigkeit $\pm 1{,}5$ %. Konstanz $\pm 0{,}5$%. Zeitkonstante 4 Sekunden und Empfindlichkeiten bis zu 6 μV/mm oder 0,03 μA. Zwischen 24 cm/min und 1,5 cm/h sind 14 Vorschübe vorhanden.

EE 72 751 *für weitere Einzelheiten*

Relais Tester

Data Laboratories Ltd, 3-7 Hatfields, London, S.E.1

(Abbildung Seite 562)

Dieses Gerät wurde für die im MIL-Pflichtenblatt DEF 5165 vorgeschriebene komplette Prüfung der Arbeits- und Abfallzeit, Prellperiode sowie Stoss- und Schwingungsfestigkeit von Relais entwickelt. Ausserdem ermöglicht das Gerät die schnelle und eindeutige Überprüfung der in den Pflichtenblättern der Hersteller für die obigen Kenndaten genannten Werte unter Betriebsbedingungen.

Bei Prüfung des Arbeitens eines Relais wird dasselbe mit Impulsen beaufschlagt, die von einem regelbaren Multivibrator mit einer Folgefrequenz abgegeben werden, die für das zu prüfende Relais geeignet ist. Arbeits-, Abfall- und Prellzeit lassen sich gleichzeitig messen, und für jede der Zeiten kann ein Höchstwert eingestellt werden. Bei Überschreiten dieses Wertes für irgendeine dieser Kenndaten leuchtet eine entsprechende Signallampe auf.

Die gemessene Arbeitszeit beginnt, wenn die Spule erregt wird, und endet, wenn die normalerweise offenen Kontakte fest geschlossen sind, d.h. die Prellperiode ist einbegriffen. Die Zeit kann je nach Wunsch eingestellt werden, und bei Überschreiten dieses Wertes zeigt die Signallampe einen Fehler dieses Kontaktes an. Ein Fehler wird angezeigt, wenn der Kontakt eine Mikrosekunde nach Ablauf der eingestellten Arbeitszeit offen bleibt.

Beim Abschalten des Relais wird die Abfallzeit in gleicher Weise vom Zeitpunkt des Aberregens der Relaisspule bis zum sicheren Schliessen des Ruhekontaktes gemessen. Auch diese Zeit ist je nach Wunsch einstellbar; bei Überschreiten des Wertes leuchtet wiederum eine entsprechende Lampe auf.

Kontaktprellen wird von dem Zeitpunkt, in dem sich der offene Kontakt schliesst, bis zu demjenigen, in dem er geschlossen bleibt, gemessen. Diese Messung wird sowohl an Arbeitskontakten wie auch Ruhekontakten vorgenommen.

Bei Stoss- und Schwingungstests mit dem Gerät wird darauf geprüft, dass sich geschlossene Kontakte nicht öffnen. Auch Stromkreisunterbrechungen werden entdeckt und mittels Signallampe angezeigt.

EE 72 752 *für weitere Einzelheiten*

Geregelte Stromversorgung

Feedback Ltd, Park Road, Crowborough, Sussex

(Abbildung Seite 562)

Dieses neue Mehrzweckgerät in Tischgehäuseausführung kann mit Netzspannungen von 200...2400 V oder 100...125 V bei 50...60 Hz betrieben werden.

Ein besonderes Merkmal ist die Vielseitigkeit der Schaltungsmöglichkeiten, die die folgenden vier Betriebsarten ermöglicht:

(a) zwei getrennte massefreie Ausgangsspannungen von 300 V bei 150 mA_{max}
(b) eine Hochspannungsversorgung, die bei +300 V und −300 V 150 mA abgibt
(c) eine Hochspannungsversorgung, die bei +600 V und −600 V 150 mA abgibt
(d) eine Hochspannungsversorgung, die bei + und − 300 V 300 mA abgibt.

Eine einfache interne Verbindung erhöht bei Betrieb unter konstanten Mindestlastbedingungen die Höchstbelastung an jedem der Ausgänge auf 200 mA.

Drei 6,3-V-Ausgänge sind vorhanden, von denen zwei massefrei sind; sie können bis zu 16 A Heizstrom abgeben. Das grosse Kontrollgerät zeigt die Stromentnahme für jeden Ausgang an. Für Gesamtnetzschwankungen von 15 % des Nennwertes ist die Langzeitkonstanz grösser als 300:1. Die Ausgangsimpedanz liegt unter 1 Ω und die Gesamtrestwelligkeit unter 3 mV_{eff}.

Die Spannungen können entweder Klemmen auf der Frontplatte oder über eine 6 polige Painton-Steckverbindung auf der Rückseite des Gerätes entnommen werden. Diese Anordnungen stehen mit den Erfordernissen der Analog- und Simulatorgeräte des Feedback-Fertigungsprogrammes in Einklang.

EE 72 753 *für weitere Einzelheiten*

Polyester-Kondensatoren

Geschäftsbereich Kondensatoren, Plessey-UK Ltd, Kembrey Street, Swindon, Wiltshire

(Abbildung Seite 563)

In Grossbritannien als "Melinex" und in den USA als "Mylar" bekannte Polyester-Kunststoffe ergeben ein ausgezeichnetes Dielektrikum. Sie haben—auch bei den in modernen elektronischen Geräten auftretenden erhöhten Temperaturen—eine hohe Durchschlagsfestigkeit. Bei Verwendung von metallisiertem "Mylar" kann man Kondensatoren mit sehr hoher Raumwirtschaftlichkeit herstellen, die für dieselbe Arbeitsspannung kleiner sind als die entsprechenden MP-Kondensatoren.

Polyester-Kondensatoren werden nunmehr von Plessey-UK in drei Hauptausführungen geliefert, und zwar als ungeschützte "Melinex"-Typen, umpresste "Melinex"-Typen und umpresste metallisierte "Mylar"-Typen. Ausserdem können im Sonderauftrag Polyester-Kondensatoren für sehr hohe Spannungen—bis zu 15 kV—und Polykarbonat-Einheiten mit niedrigen Verlustfaktoren bei Temperaturen bis zu 150° C hergestellt werden.

Die ungeschützten "Melinex"-Kondensatoren sind in Standardwerten von 10 pF bis zu 2,0 μF für 250 V Arbeitsgleichspannung, 1,0 μF für 400 V= und 0,1 μF für 1 kV= lieferbar. Die Fertigung dieser Kondensatoren erfolgt bei hoher Temperatur, und die Enden sind mit Kunstharz verschlossen. Sie entsprechen der Feuchtigkeitsklasse H3 und haben einen Temperaturbereich von −40° C...+125° C.

Umpresste "Melinex"-Kondensatoren werden in sieben Grössen hergestellt. Die lieferbaren Werte und Arbeitsspannungen entsprechen ungefähr den oben angegebenen, jedoch ist der Feuchtigkeitswiderstand verbessert und entspricht Klasse H2. Der Temperaturbereich ist −40° C...+100° C.

Umpresste metallisierte "Mylar"-Typen werden für die gleiche Feuchtigkeitsklasse und denselben Temperaturbereich in Werten von 4 μF bei 200 V= Arbeitsspannung bis zu 0,1 μF bei 2 kV hergestellt. Die Abmessung des letzteren ist nur 55,6 mm × 19,1 mm. Metallisierte "Mylar"-Kondensatoren sind auch in zylindrischen Aluminiumgehäusen bis zu 100 μF bei 200 V= Arbeitsspannung lieferbar.

EE 72 754 *für weitere Einzelheiten*

HF-Leistungsmesser

Electro Apparatus (London) Ltd, Stansted, Essex

(Abbildung Seite 563)

Dieser HF-Leistungsmesser wurde zur Deckung des Bedarfs in der beweglichen Nachrichtenverkehrstechnik entwickelt.

Er erfordert keinen Netzanschluss und ist in gleicher Weise für Einsatz im Laboratorium wie Aussendienst geeignet. Das Gerät braucht vor dem Messen nicht eingeregelt zu werden und gibt eine sofortige Leistungsanzeige. Der Fertigung des 50-Ω-Abschlusswiderstandes und der Anpassung der Diode an die koaxiale Leitung wird besondere Aufmerksamkeit gewidmet. Auf diese Weise wird ein Stehwellenverhältnis von besser als 1,1 bis zu Frequenzen von 1 GHz erzielt. Jedes Gerät wird einzeln geeicht, und über den Bereich von 10 MHz...1 GHz wird eine Gesamtmessunsicherheit von 10 % aufrechterhalten. Zur Erweiterung des Messumfanges auf 250 mW und 2,5 W sind 10-dB-Abschwächer lieferbar. In Kürze wird die Fertigung weiterer Abschwächer für höhere Leistungen aufgenommen.

EE 72 755 *für weitere Einzelheiten*

Zweipolige Motoren

Garrard Engineering Ltd, Swindon, Wiltshire

(Abbildung Seite 563)

Garrard Engineering kündigt neue zweipolige, als Serie S20 bezeichnete Motoren an.

Robuste Konstruktion und Zuverlässigkeit wurden in der Entwicklung betont, und neue Fertigungsverfahren ermöglichen es, ein wettbewerbsfähiges Erzeugnis für eine grosse Auswahl von Anwendungsmöglichkeiten herzustellen.

Motoren der Serie S20 werden für fünf Leistungen im Bereich 18 bis 40 W hergestellt, die Anlaufmomente von 80 bis zu 250 g. cm. abgeben. Die Geschwindigkeit der Motoren hängt von der Netzfrequenz ab und ist 2500 bzw. 3000 UPM für 50 bzw. 60 Hz. Rechts- und linkslaufende Ausführungen sind lieferbar.

Jeder Motor der Serie S20 ist mit einer auf einen Nylon-Formkörper gewickelten Einzelspule gebaut, die hohe Isolation und Schutz gibt.

Die Isolierung wird mit 1,5 kV geprüft. Die Lagerdeckel des Motors bestehen aus verformtem und gepresstem Aluminium, in das Phosphorbronze-Pendellager mit Öldichtung eingebaut sind.

Geräusch- und schwingungsfreies Laufen wird durch einen dynamisch ausgewuchteten Rotor erzielt. Der aus Qualitätsstahl hergestellte Rotor ist gehärtet, angelassen, geschliffen und sehr fein poliert. Zur Erzielung eines maximalen Betriebswirkungsgrades werden die Lamellen aus verlustarmem Siliziumstahl hoher Qualität gefertigt. Die überstehende Wellenlänge ist 25,4 mm.

Die Leistung der Garrard-Motoren Serie S20 entspricht den Anforderungen der britischen Norm BSS 170.

EE 72 756 *für weitere Einzelheiten*

Programmregler

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Abbildung Seite 563)

Der "Numalec Minor" ist im Prinzip ein in ein kompaktes, geschlossenes Instrument verdichtetes Steuerfeld.

Er regelt Temperatur, Druck und Feuchtigkeit genau und vollautomatisch für lange oder kurze Perioden. Da jedes der Instrumente aus Standard-Bausteinen für die Anforderungen bestimmter Regelaufgaben zusammengebaut wird, ist dieser Programmregler einer der preisgünstigsten und vielseitigsten, die zur Zeit angeboten werden.

Der automatische Regelgang wird durch Drücken einer Taste eingeleitet, und nach Ablauf des Vorganges wird das Instrument selbsttätig auf Wiederholung genau desselben Programmes zurückgestellt. Neue Programme kann man ohne Schwierigkeiten durch Verschieben von Armen in Bezug auf die Kreisblattskala und mittels einer Schaltuhr, die durch Anheben des Diagrammblattes an einer bestimmten Stelle zugänglich wird, eingeben. Für grössere Anpassungsfähigkeit der Regelung einschliesslich Anstiegs- und Abfallraten kann ein vorbearbeiteter Nocken eingebaut werden.

Das Diagrammblatt hat einen Durchmesser von 254 mm und registriert fortlaufend ein oder zwei Variable; wenn der Schreibstift einen vorgegebenen Zustand erreicht, wird ein elektrisches oder pneumatisches Signal erzeugt, das Magnet- oder Stellventile oder pneumatische Regler betätigt. Je nach Bauart der Schaltuhr lassen sich ein oder zwei veränderliche Zeitspannen benutzten, die in gleicher Weise anlaufen und zusätzlich zu Warnlampen und Alarmeinrichtungen eine narrensichere, zeitgerechte Programmregelung ermöglichen. Die Regelgenauigkeit steht in direkter Beziehung zur Genauigkeit der registrierten Werte, die besser als ±1 Prozent ist.

EE 72 757 *für weitere Einzelheiten*

Regelnetzgeräte

Newton Brothers (Derby) Ltd, Alfreton Road, Derby

(Abbildung Seite 563)

Die volltransistorisierten Netzgeräte Typ KB wurden speziell als hochkonstante Gleichstromversorgungen entwickelt, die für Einsatz in Forschungs- und Entwicklungslaboratorien, Prüffeldern, Service-Werkstätten und Ausbildungsstätten gedacht sind. Sorgfältige Konstruktion sowie eine Anzahl neuer Schaltungen ergeben Geräte mit ausgezeichneten Leistungseigenschaften, kleinen Abmessungen und mässigen Anschaffungskosten.

Die Geräte haben niedrige Ausgangsimpedanz und kommen in sechs Ausführungen für Leistungen von 15 V bei 200 mA bis zu 40 V bei 1 A. Alle Modelle sind mit Strombegrenzern und Kurzschlussschutz ausgerüstet. Die Ausgangsspannung wird mittels eines Potentiometers mit hohem Auflösungsvermögen, das in einigen Modellen durch ein 3- oder 10gängiges Wendelpotentiometer ersetzt werden kann, kontinuierlich geregelt.

Bei niedrigem Rauschen und geringer Restwelligkeit werden hohe Spannungsstabilitäten bis zu ±0.01 Prozent erreicht.

EE 72 758 *für weitere Einzelheiten*

Schwingungsmesser

Dawe Instruments Ltd, Western Avenue, London, W.1

(Abbildung Seite 564)

Der Schwingungsmesser 1443A soll das erste für das Messen von Ruck ausgelegte Gerät sein. Das in britischen oder metrischen Einheiten geeichte Instrument zeigt Weg, Geschwindigkeit, Beschleunigung und Ruck an; es ist tragbar, in sich geschlossen und batteriebetrieben. Das Gerät hat einen Frequenzumfang von 1,6 Hz...10 kHz bei neun Skalenbereichen.

Das System besteht aus einem von zwei wahlweise lieferbaren Schwingungswandlern, einem Kathodenfolger-Messkopf und dem elektronischen Hauptgerät mit einem Hochleistungsverstärker, integrierenden und differenzierenden Netzwerken, regelbarem Abschwächer und direktanzeigendem Messgerät. Hohe Stabilität und Eichgenauigkeit (innerhalb ±1 Prozent bei Mittelfrequenzen) werden durch Gegenkopplung und Konstant-Stromversorgung erzielt.

Als Wandler werden Beschleunigungsmesser benutzt. Der leichte Standard-Wandler wiegt ca. 20 g, arbeitet über den ganzen Frequenzbereich und hat eine Empfindlichkeit von 16 mV_S/g_S. Der hochempfindliche Wandler wiegt ca. 23 g und arbeitet von 1,6 Hz ab mit einer Empfindlichkeit von 450 mV_S/g_S. Die Resonanzfrequenzen sind 115 kHz bzw. 3,8 kHz.

Die vom Beschleunigungsmesser abgegebene Spannung wird in den hochohmigen (500 MΩ parallel zu 10 pF) Kathodenfolger-Messkopf gespeist, der über ein 1,80 m langes dreiadriges Abschirmkabel mit dem Messgerät verbunden ist. Das dem Messkopf entnommene Signal wird für Ruck-Messungen erst durch eine differenzierende Schaltung, für Geschwindigkeit durch eine integrierende und für Wegmessungen durch eine doppelintegrierende Schaltung geleitet. Bei Einstellen der entsprechenden Schalterposition wird die gewünschte Charakteristik direkt angezeigt.

Die Arbeitsbereiche des neuen Schwingungsmessers überstreichen 3×10^{-6} bis zu 3×10^{2} Zoll für Weg, 3×10^{-3} bis zu 3×10^{3} Zoll/Sek für Geschwindigkeit, 3 bis zu 3×10^{-4} Zoll/Sek² für Beschleunigung und 30 bis zu 3×10^{6} Zoll/Sek³ für Ruck. Das gleiche Instrument wird mit metrischen Skalen und gleichen Bereichen als Typ 1433B geliefert.

Das verstärkte Signal kann über eine Ausgangsklinke für Darstellung auf dem Schirm eines Oszillografen oder Speisung in einen Frequenzanalysator entnommen werden.

EE 72 759 *für weitere Einzelheiten*

Tastatur-Koordinatenschreiber

Scientific Furnishings Ltd, Poynton, Cheshire

(Abbildung Seite 564)

Der Fachbereich Elektronik der Scientific Furnishings Ltd kündigt eine neue, preisgünstige Digitaltastatur Typ K1 an, deren Einsatz zusammen mit dem selbsttätigen Koordinatenschreiber XY-1P die genaue grafische Darstellung tabellierter Digitaldaten beschleunigt. Die Datentabellen können von Messwertanzeigen, Rechenanalysen, Berechnungen, Versuchen oder statistischen Erfassungen herrühren. Die Daten werden durch Eintasten auf zwei dreistelligen Zehnertastaturen und Drücken einer "Schreib"-Taste in Punkte einer Kurve umgesetzt. Die Tastatur beaufschlagt die X- und Y-Achsen des Schreibers mit Gleichspannungen, deren Grössen mit einer Genauigkeit von ±0,25 Prozent den in die X- und Y-Tastaturen eingegebenen Koordinaten proportional sind. Nullunterdrückung und Skalendehnung erlauben optimale Darstellung; man kann sowohl positive wie auch negative Werte auftragen.

Hohe Kosten haben die Tastaturumwandlung tabellierter Daten in Kurven bisher begrenzt. Die neue Ausrüstung kostet nur ein Drittel des Anschaffungspreises früherer Geräte und macht Tastatur-Koordinatenschreiben somit nun auch den allgemein üblichen Laboratorien zugänglich.

EE 72 760 *für weitere Einzelheiten*

Präzisions-Wechselspannungsteiler

Digital Measurements Ltd, 25 Salisbury Grove, Mychett, Aldershot, Hampshire

(Abbildung Seite 564)

Der von Digital Measurements Ltd entwickelte Präzisions-Wechselspannungsteiler DM 2001 kann zum genauen Messen oder Erzeugen von Wechselspannungsverhältnissen, sowie zum Vergleichen von Widerstands-, Induktivitäts- und Kapazitätsnormalen oder -verhältnissen eingesetzt werden. Spannungen können mit einer Genauigkeit und Reproduzierbarkeit von 1×10^{-6} geteilt werden, und das soll eine Grössenordnung besser sein, als bisher erreichbar.

Das Gerät besteht aus einer Serie von Spartransformatoren, von denen jeder in zehn gleiche Abschnitte unterteilt ist und jeder gewählte Abschnitt sich durch den nachfolgenden Transformator weiter unterteilen lässt. Einstellungen werden auf Messgerätskalen angezeigt oder entfernt über automatische Schreibmaschinen oder Streifenstanzer ausgeschrieben.

Der Erfolg des Gerätes beruht auf der Verwendung einer Legierung mit einer ungewöhnlich hohen, zuverlässigen Permeabilität, aus der der Kern gewickelt ist. Diese mit "Supermumetal 100" bezeichnete Legierung wird in Grossbritannien von Telcon Metals Ltd, Crawley, Sussex hergestellt und hat die Sollzusammensetzung 77Ni/14Fe/5Cu/4Mo. Sie wird sowohl in Bezug auf chemische, als auch physikalische Eigenschaften strengen Tests unterworfen und hat eine Anfangspermeabilität von über 100 000 bei einem Höchstwert von über 500 000. "Supermumetal 100" wird durch Vakuumschmelzen aus reinsten Rohmaterialien hergestellt und zu dünnen Folien gewalzt, bevor es in Ringkerne gewickelt wird.

EE 72 761 *für weitere Einzelheiten*

Gleichrichter

The Wayne Kerr Laboratories Ltd, New Malden, Surrey

(Abbildung Seite 565)

Der neue, preiswerte Gleichrichter wurde in erster Linie für Verwendung mit "Digitec"-Gleichspannungsmessern entwickelt, ermöglicht jedoch auch genaues Messen von Wechselspannungspegeln mit jedem beliebigen Messgerät, das mit Eingangsspannungen von 1 V= arbeitet. Der Wayne Kerr-Digitec-Gleichrichter 1900 arbeitet mit einer Linearität von ±0,05 des Skalenendwertes über den Frequenzbereich 50 Hz...25 kHz. Es sind vier Eingangsspannungsbereiche vorhanden, und zwar 0...1, 10, 100 und 500 V_{eff}, deren Eingangsimpedanz 1 MΩ (100 kΩ für den 1-V-Bereich) ist.

In diesem Gleichrichter werden die Diodeneigenschaften mittels eines Gegenkopplungssystems hoher Verstärkung linearisiert und konstantgehalten, und die Festkörperschaltung ist für die Aufrechterhaltung grosser Genauigkeit über einen breiten Temperaturbereich ausgelegt. Die Unsicherheit ist von 100 Hz... 10 kHz ±0,3 Prozent des Skalenendwertes (von 50 Hz...25 kHz 0,4 Prozent). Die Ausgangsgleichspannung (1 V= für 1, 10 oder 100 V_{eff}, 0,5 V= für 500 V_{eff}) kann bei niedriger Impedanz (10 kΩ) entnommen werden, und die Einstellzeit ist 1 s.

Das Gerät kann in einem den Digitec-Voltmetern angepassten Traggehäuse, mit Flanschen oder für Gestelleinbau (einzeln oder zusammen mit Voltmeter) geliefert werden. Die tragbare Bauart ist 140 mm hoch, 203 mm breit, 178 mm tief und wiegt ca. 3,2 kg.

EE 72 762 *für weitere Einzelheiten*

Mehrfach-Glimmerkondensatoren

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London, E.C.1

(Abbildung Seite 565)

Johnson, Matthey & Co. Ltd. hat für die Herstellung von Verzögerungsleitungen Spezial - Mehrfachkondensatorenblöcke entwickelt. Die in den Blöcken verwendeten Kondensatoren sind mit der Bauform der unter dem Namen "Silver Star" bekannten Präzisionskondensatoren mit versilberten Glimmerplättchen identisch.

Einer dieser Blöcke wird in der Abbildung gezeigt. Er ist mit Epoxydharz umhüllt und enthält bei Abmessungen von nur 76,2 × 11,4 × 5,1 mm 23 versilberte Glimmerkondensatoren mit Kapazitätswerten von 22 pF. Die Kondensatoren liegen an einer Seite über einen gemeinsamen Anschlussdraht, der an beiden Enden des Blocks herausgeführt ist, an Masse. Die eingebrannte Bauart "Silver Star" gewährleistet ausgezeichnete Stabilität und Zuverlässigkeit der Kondensatoren in Betrieb.

In dieser Blockgrösse können die Kondensatoren eine Höchstkapazität von 200 pF haben und sind für 150 V_s bemessen. Andere Blöcke mit anderen Kondensatorenzahlen und Kapazitätswerten können hergestellt werden.

Diese Mehrfachkondensatorenblöcke wurden entwickelt, um den Bedarf an Bauelementen zu decken, die die Zusammenbauzeit für Verzögerungsleitungen verkürzen und zu Raumersparnissen führen. Die Anschlussdrähte lassen sich einfacher handhaben und verlöten, als die von Einzelkondensatoren, und sind in genauen Abständen voneinander über den Block verteilt.

EE 72 763 *für weitere Einzelheiten*

Hohlleiter-Kalorimeter

The Marconi Co. Ltd, Chelmsford, Essex

(Abbildung Seite 565)

Die Abteilung für Spezialbauelemente der Marconi Co Ltd kündigt die Entwicklung eines Hohlleiter-Kalorimeters Grösse 12 an. Es kann mittlere Leistungen bis zu 2 kW messen und benutzt ein neues keramisches Belastungsmaterial, das ohne Schaden über längere Zeitspannen hohe Temperaturen aushalten kann. Mit diesen neuen Kalorimetern wird die Ausgangsleistung von Sendern nach dem Leistungsvergleichsverfahren ermittelt. Gegenüber älteren Messverfahren, die Berechnungen auf Grund einer Anzahl von Messungen erforderten, bietet diese Methode grössere Messgenauigkeit.

Das Kalorimeter überstreicht den Frequenzbereich 3,95...5,99 GHz und hat über den gesamten Bereich ein Stehwellenverhältnis von weniger als 1,05. In die Ein- und Ausflussöffnungen für das Kühlmittel sind Dioden-Temperaturfühler eingebaut, die man zur Anzeige der Temperaturdifferenz zwischen Ein- und Ausfluss verwenden kann. Im Körper des Kalorimeters ist für Eichzwecke ein netzbetriebenes Heizelement vorhanden.

Bei eingeschalteter HF-Leistung und konstanter Kühlmitteldurchströmung des Kalorimeters gibt das Messgerät einen der Temperaturdifferenz zwischen Ein- und Ausflussöffnung proportionalen Ausschlag. Bei abgeschalteter HF-Leistung wird das Eichheizelement eingeschaltet und die Netzspeisung so geregelt, dass sich derselbe Ausschlag ergibt. Diese dem Netz entnommene Leistung ist dann der angelegten HF-Leistung gleich.

Das Kalorimeter kann für kurze Zeit ohne laufendes Kühlwasser als Absorber betrieben werden. Statt der Wasserspeiseleitung wird ein kleiner Ausgleichbehälter von 0,83 l Kapazität angebaut, der die bei der Erwärmung erfolgende Ausdehnung des Wassers auffängt.

Die neue Konstruktion des Kalorimeters ist eine Weiterentwicklung der wassergekühlten Hochleistungsabsorber und überkommt die Nachteile älterer Typen ähnlicher Leistungsmessausrüstungen mit Glas- oder Quarzröhren, wie z.B. Brechbarkeit und veränderliche Stehwellenverhältnisse.

EE 72 764 *für weitere Einzelheiten*

Robustes Magnetron

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Abbildung Seite 565)

Die neuste Ergänzung des Fertigungsprogrammes der M-O Valve Co. Ltd für robuste, zuverlässige Magnetrons ist eine Röhre für das X-Band mit einer Spitzenleistung von 7 kW und einer Anheizzeit von unter 10 Sekunden.

Dieses Magnetron—Typ E 3094—kann Schwingungsdurchläufe von 20 Hz...10 kHz bei Beschleunigungen, die mit 3 g beginnen und bis zu 50 g ansteigen, aushalten. Unter diesen Umständen schwankt die Ausgangsfrequenz nicht mehr als ±3 MHz. Es kann 40-g-Stössen für 10 ms widerstehen. Heizerstrom und Hochspannung können gleichzeitig eingeschaltet werden, wobei es jedoch auf die jeweilige Modulatorschaltung ankommt. Bei hohen Spannungs-Anstiegsraten und Betrieb mit kurzen Impulsen ist die Stabilität während der ersten 10 Betriebssekunden besser als 0,1 Prozent.

Die Röhre, die mit einem eingebauten Thermoelement für das Messen der Anodentemperatur ausgerüstet ist, kann in einem Umgebungstemperaturbereich von −65° C...+90° C und Drücken bis zu 200 mm Quecksilber herunter betrieben werden.

Die Abmessungen des Magnetrons sind ca. 120,6 × 108 × 50,4 mm, das Gewicht rund 1,8 kg.

EE 72 765 *für weitere Einzelheiten*

Industrielles Tachometer

Flight Refuelling Ltd, Wimborne, Dorset

(Abbildung Seite 566)

Ein von Flight Refuelling Ltd hergestelltes industrielles Tachometer bietet ein genaues Messverfahren für die Bestimmung von Drehzahlen.

Der Ausgang steht entweder als Impulsreihe oder—mit einer FR-Umformerschaltung—als für ein Messgerät geeignetes Gleichspannungssignal zur Verfügung.

Die Impulsreihe entsteht aus der Interferenz eines Abtastkopfes mit den sechzig magnetischen Kreisen eines rotierenden Tachometerrades.

Abtastkopf und Rad sind in einem robusten Gussgehäuse für schwere industrielle Verwendungszwecke untergebracht und der Zusammenbau in für die meisten Umgebungszustände geeigneter Form abgedichtet. Das maximale zum Treiben des Tachometers erforderliche Drehmoment ist 72 cm. g.

Zur Herabsetzung des Ausfallrisikos ist die volltransistorisierte Umformerschaltung in einem kleinen, dichten Spritzgusskästchen untergebracht und durch sorgfältige Wahl der Schaltungsparameter elektrisch geschützt. Das abgegebene Gleichstromsignal ist über den gesamten Betriebsbereich der Drehzahl innerhalb ±0,2 Prozent proportional.

Das Tachometer misst Geschwindigkeiten mit Bereichendwerten von 3000 UPM bis zu 15 000 UPM in den in der britischen Norm BS 3403 empfohlenen Bereichen.

EE 72 766 *für weitere Einzelheiten*

Zählsystem

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Abbildung Seite 567)

Zur Schaffung eines robusten und preiswerten industriellen Vorwahl-Zählsystems hat Intersonde Ltd ein Zählgerät FL87 für Einsatz mit induktiven Nahwirkungsschaltern ihrer Serie QD eingeführt.

Das Zählgerät ist als leicht installierbares Stahlgehäuse für Wandmontage mit Abmessungen von ca. 152 × 152 × 102 mm ausgeführt und enthält einen sechsstelligen elektromagnetischen Zähler mit manueller Rückstellung, sowie eine Stromversorgung für die niedrige Gleichspannung, die für die Erregung der Nahwirkungsschalter erforderlich ist. Zählgerät und Nahwirkungsschalter werden durch ein gebräuchliches dreiadriges Kabel verbunden, das bis zu 152 m lang sein kann.

Die Gegenwart eines Metallobjektes innerhalb einer Entfernung von 9,5 mm von der Fühlerfläche des Nahwirkungsschalters veranlasst die direkte Beaufschlagung des elektromagnetischen Zählers mit 24 V=. Das komplette System kann mit einer Zählgeschwindigkeit von bis zu 1500 je Minute arbeiten, der Nahwirkungsschalter allein kann jedoch bis zu 10 000 Zählungen je Minute ausführen.

EE 72 767 *für weitere Einzelheiten*

Schaltungsblöcke für Digitaltechnik

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(Abbildung Seite 566)

M.E.L. kündigte zwei neue Serien von Schaltungsblöcken für die Digitaltechnik an, für die äusserst hohe Zuverlässigkeit in Anspruch genommen wird. Die Blöcke sind nicht nur in Giessmasse verkapselt, sondern auch hermetisch dicht in Metallgehäuse eingebaut; die Anschlüsse werden durch Glas-Metall-Verschmelzungen herausgeführt.

Die Bauelemente liegen zwischen kleinen Leiterplatten. Das ergibt eine hohe Bauelementdichte, und die einzelnen Blöcke führen umfangreiche Funktionen aus. Die Anzahl der für ein System erforderlichen Blöcke ist daher klein und die Systemkonstruktion vereinfacht. Auch die Anzahl der Verbindungen sowie der erforderliche Raum werden reduziert. Die Aussenabmessungen sind 53 × 25 × 12,7 mm, und alle Typen haben 19 Anschlussdrähte.

Die beiden Serien sind miteinander kompatibel, sind beide mit npn-Transistoren bestückt und werden mit 12 V betrieben. Die Serie 10 ist mit Germanium-Transistoren bestückt, und die höchste zulässige Betriebstemperatur ist 55° C (Umgebung). Der Grund-Flipflop schaltet in 3,5 µs.

Die Serie 20 ist mit Silizium-Epitaxial-Elementen bestückt und für Umgebungstemperaturen bis zu 85° C geeignet. Ausserdem arbeiten die Silizium-Einheiten schneller, und der Flipflop kann in 50 ns schalten. Mit ihrer höheren Leistung können sie Magnetspeicher treiben, wobei die höchstzulässige Temperatur dann 65° C ist.

Jede Serie beruht auf NAND-Logik und enthält Flipflops, Torschaltungen, Impulsformer und Verstärker. M.E.L. weist darauf hin, dass beim Entwurf besonders auf die Verhinderung ungewollter Betätigung durch Schaltstösse und andere Arten regelloser elektrischer Schwankungen in der Umgebung Wert gelegt wurde.

Die Abbildung zeigt einen der neuen Schaltungsblöcke für Digitaltechnik vor Verkapselung und Einbau in das dichte Metallgehäuse.

EE 72 768 *für weitere Einzelheiten*

NF-Spannungsverstärker

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Abbildung Seite 566)

Dieser Verstärker wurde für Einsatz mit dem phasenempfindlichen Brookdeal-Detektor PD629 in Messungen, in denen sinus- und rechteckförmige Modulationsfrequenzen zwischen 10 Hz und 10 kHz auftreten, entwickelt.

Durch seine sehr niedrigen Brumm- und Rauschpegel, sowie seinen gradlinigen Frequenzgang und die Abwesenheit von Phasenverschiebung ist er als Mehrzweckverstärker für den Tonfrequenzbereich geeignet.

Zwischen 10 Hz und 10 kHz ist die Phasenverschiebung für Schmalband-Suchbereichzwecke vernachlässigbar klein (cos α ist an den extremen Enden mehr als 0,98) und die Verstärkung innerhalb ±0,25 dB konstant. Die Einflüsse der Röhrenkenndatenstreuung werden durch Anwendung von 20 dB Wechsel- und Gleichstromgegenkopplung in jeder Stufe beseitigt.

Bei voller Ausnutzung der Gegenkopplung ist die maximale Verstärkung des Gerätes 75 dB; in diesem Zustand ist der äquivalente Rauschwiderstand 50 kΩ. Ein Schalter "Vorverstärkung" gibt 20 dB zusätzliche Verstärkung und einen äquivalenten Rauschwiderstand von unter 2 kΩ. Der Brummpegel ist sehr niedrig und entspricht 1 µV am Eingang.

Die Verstärkung kann durch einen verteilten Abschwächer in 5-dB-Stufen um 55 dB reduziert werden. Ein Tiefpassfilter mit Zeitkonstanten von 0,05, 0,5 und 5 ms ist eingebaut.

An Klemmen der Frontplatte steht eine Konstant-Stromversorgung (150 V, 1 mA) zur Erregung von Fotowiderstandsdetektoren zur Verfügung.

EE 72 769 *für weitere Einzelheiten*

Mehrbereich-Messgerät

Smiths Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Abbildung Seite 567)

Dieses hochempfindliche Messgerät ist in zwei Ausführungen lieferbar:

Modell 4S hat bei einer Empfindlichkeit von 100 kΩ/V für Gleichstrom und 20 kΩ/V für Wechselstrom die folgenden Messbereiche: 0 ... 100 mV bis zu 0 ... 5000 V=, 0 ... 10 µA bis zu 0 ... 1 A≅, 0 ... 10 V bis zu 0 ... 1000 V∼, 0 ... 200 Ω bis zu 0 ... 500 MΩ, 2000 pF ... 5 µF, −8 dB ... 50 dB und ist für Frequenzen bis zu 20 kHz geeignet.

Modell 3S ist bei einer Empfindlichkeit von 25 kΩ/V für Gleichstrom und 2 kΩ/V für Wechselstrom auch für Frequenzen bis zu 20 kHz geeignet und hat Bereiche von 0 ... 100 mV bis zu 0 ... 5000 V≅, 0 ... 5 mA bis zu 0 ... 5 A≅, 0 ... 200 Ω bis zu 0 ... 50 MΩ, 100 ... 20 000 pF und −14 dB ... 46 dB.

Diese robusten Messgeräte haben ein stossfestes Messwerk mit Spannbandaufhängung und Glaszeigern, die eine für leichtere Ablesbarkeit geneigte Skala überstreichen. Ein Einzeldrehschalterknopf erlaubt bequeme Bedienung, und das Gerät ist durch seine Konstruktion leicht und handlich. Es wiegt nur ca. 1,4 kg.

Die Messgeräte sind mit einem patentierten elektrischen Sicherungsautomaten ausgerüstet, der innerhalb 0,005 ... 0,01 s abschaltet, während eine Schmelzsicherung gegen hohe Stromüberlastung schützt. Stromwandler, Sperrkondensatoren und Diodengleichrichter sind eingebaut. Die Geräte können überlagerte Wechsel- oder Gleichstromsignale messen.

Zum weiteren Ausbau der Messbereiche sind Stromwandler, Nebenschlusswiderstände und Spannungsvervielfacher lieferbar.

EE 72 770 *für weitere Einzelheiten*

Elektronisches Thermometer

Kane-May Ltd, 243 Upper Street, London, N.1

(Abbildung Seite 567)

Mit dem neuen elektronischen Thermometer "Dependatherm" können momentane, sehr genaue Temperaturmessungen vorgenommen werden. Es ist ein Taschen-instrument in fortschrittlicher Schaltungstechnik, die ausgezeichnete Langzeitstabilität gewährleistet. Messergebnisse werden sofort nach Anlegen eines bleistiftförmigen Messkopfes an die Temperaturquelle, die gasförmig, flüssig oder fest sein kann, auf einer 63,5 mm langen Skala angezeigt.

Das schnelle Ansprechen des Instrumentes auf Temperaturschwankungen erleichtert Messungen, die bisher als schwierig betrachtet wurden, und ermöglicht die Darstellung von Temperaturgefällen, die Eingrenzung von überhitzten Stellen, die Verteilung der Oberflächentemperatur usw. ohne Testeinrichtungen irgendwelcher Art. Die in der Verfahrenstechnik erforderlichen Temperaturmessungen können viel bequemer ausgeführt werden. Das Gerät ist in gleicher Weise für Einsatz in der Fertigung wie im Labor geeignet.

Das elektronische Thermometer Dependatherm enthält eine Stabilisationsautomatik, die Abgleich in jeder Form vor dem Test unnötig macht—das Instrument ist sofort einsatzbereit—und die gewährleistet, dass die Anzeigegenauigkeit in breiten Grenzen von dem allmählichen Abfallen der Batteriespannung unabhängig ist.

Einzel- und Mehrbereichmodelle sind für eine oder mehrere Temperaturspannen im Bereich 0 ... 200°C oder 32 ... 350°F lieferbar. Standard-Messköpfe in verschiedenen Ausführungen sowie Spezial-Messköpfe, die Sonderanforderungen der Metallbearbeitung, Chemie und Verfahrenstechnik genügen, stehen zur Verfügung. Verbindungen zwischen Messkopf und Instrument können bis zu 15 m lang sein. Die Abmessungen des Gerätes sind 114 × 63,5 × 38 mm, das Gewicht ca. 215 g.

EE 72 771 *für weitere Einzelheiten*

Funkfernsprecher

Racal Electronics Ltd, Bracknell, Berkshire

(Abbildung Seite 567)

Der Funkfernsprecher TRA.355 ist ein modernes kompaktes Gerät für Einseitenbandverkehr, das völlig mit AM-Systemen kompatibel ist und im tonlosen Telegrafiebetrieb arbeiten kann. Die Sollspitzenleistung im Sprechbetrieb ist 125 W. Die Ausgangsleistung sinkt bei keiner Frequenz unter 100 W. Das Gerät kann an das Wechselstromnetz angeschlossen oder mit einer 12-V-Batterie betrieben werden.

Die Standard-Ausführung hat einen Frequenzbereich von 3 ... 12 MHz, der gegen einen kleinen Zuschlag bis auf 15 MHz und bis zu 2 MHz herunter erweitert werden kann. Vier quarzgesteuerte Rastfrequenzkanäle sind vorhanden, und zwar zwei über und zwei unter 6 MHz. Das Gerät ist für Betrieb mit einer niederohmigen Antenne, z.B. einem Dipol, ausgelegt, kann jedoch mittels eines Zusatzes an eine Freiantenne angepasst werden.

Modell TRA.355 ist ein völlig in sich geschlossenes Gerät, bei dessen Konstruktion besonders darauf geachtet wurde, die Bedienung einfach zu halten und die Wartung durch freie Zugänglichkeit zu erleichtern. Zusätzliche Einrichtungen für Verlängerung von Bedienung oder Fernsteuerung, sprachgesteuertes Sende-Empfangsschalten und Einbau in ein Fernsprechsystem sind lieferbar. Ein einfaches Messystem für Test und Einstellarbeiten kann eingebaut werden.

Alternative Stromversorgungen für Wechselstromnetzanschluss und 12-V-Batteriebetrieb sind auf getrennten Subchassis zusammengebaut und je nach Wunsch schnell zu installieren.

EE 72 772 *für weitere Einzelheiten*

Impulshöhenanalysator

Research Electronics Ltd, Bradford Road, Cleckheaton, Yorkshire

(Abbildung Seite 567)

Ein neuer, fortschrittlicher Einkanal-Impulshöhenanalysator mit Abtastautomatik wurde von Research Electronics Ltd angekündigt. Das als Modell 9050A bezeichnete Gerät gehört zum Fertigungsprogramm "Hohe Genauigkeit", das entwickelt wurde, um alle Anwendungsmöglichkeiten und Verfeinerungen für hochwertige Gamma-Spektroskopie und Impulshöhenuntersuchungen einzuschliessen.

Es wird empfohlen, den Analysator zusammen mit dem linearen Impulsverstärker der Serie "Hohe Genauigkeit" einzusetzen, für den technische Daten in Kürze bekanntgegeben werden. Die Kombination dieser beiden Geräte dürfte für Universitäten, Ingenieurschulen und andere Laboratorien interessant sein, die sich mit Ausbildung und fortgeschrittenen Forschungsaufgaben, die Isotoptechnik erfordern, befassen.

Eine Spezialschaltung gibt gegenüber herkömmlichen Schaltungen eine Verbesserung von 20:1. Da sich die neuartige Diskriminatorschaltung in keiner Weise auf die Grenzcharakteristik von Mehrelektrodenröhren stützt, wird jegliche Drift, die auf Ursachen wie z.B. Änderungen der Gitterspannung zurückzuführen sein könnte, völlig beseitigt. Je nach Anwendungsaufgabe kann man, um Änderungen des Mittelwertes der Impulswellenform bei Änderungen der Impulsfolgefrequenz gerecht zu werden, die Gleichstromkomponente im Eingang auf Wunsch einführen. Die Abtastung—von Hand, automatisch oder extern steuerbar—kann in der überlichen Weise mit einer feststehenden Kanalspannung, oder bei Benutzung eines Festprozentsatzes der Schwellenspannung (gleitender Kanal) durchgeführt werden, die eine konstante relative Kanalbreite und damit bei bestimmten Aufgaben ein besseres Auflösungsvermögen für die niedrigeren Energiespitzen gibt. Der Schwellenpegel ist zwischen 2,5 V und 100 V regelbar; die Konstanz ist besser als 0,01 V. Die Kanalbreite ist von 0,1 V ... 50 V und 1% bis zu 2,5% der Schwelle regelbar.

EE 72 773 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Demodulationsschaltungen für PAL-Farbfernseher von W. Bruch

Zusammenfassung des Beitrages auf Seite 512-518

In diesem Beitrag werden einige charakteristische Schaltungen für die Demodulation des PAL-Farbsignals beschrieben. Ein umschaltbarer Decoder erlaubt Verarbeitung des Signals nach dem PAL_S-Prinzip und ausserdem Auswertung des Signals nach dem PAL_{DL}-Verfahren; Abtrennen der PAL-Schaltungen ermöglicht Betrieb mit NTSC-Signalen. Eine Anzahl von Grundschaltungen und ihre praktische Durchrechnung werden zusammen mit Vorschlägen zur Verringerung des Aufwandes für den Modulator der Verzögerungsleitung beschrieben.

Ein preiswerter Silizium-Schalttransistor und seine Verwendung in Ringzählern von J. Palmer

Zusammenfassung des Beitrages auf Seite 519-523

Ein Silizium-Transistor-Ringzähler wurde von Ringrose (1) und Bradely (2) beschrieben. Die Konstruktion des Zählers wird eingehend besprochen und Vorschläge für den Bau einer preisgünstigen Ausführung gemacht, nach denen die Kosten einer Dekade in der neuen Ausführung denen eines Einzel-Triggerpaares in der alten entsprechen. Es bestehen Geschwindigkeitsbeschränkungen, die ebenfalls besprochen werden. Der Zähler ist mit einem preiswerten Transistor und einer Diode bestückt. Ein Entwurf wird als Beispiel beschrieben. Der Zähler kann als einfacher Zähler oder als eine Art Kommutator, in dem ein eindeutiger Zustand oder eine Zustandsmenge von Punkt zu Punkt den Ring entlang verschoben wird, eingesetzt werden.

Ein linearer 5-kW-Verstärker mit selbsttätiger Abstimmung von J. Wood

Zusammenfassung des Beitrages auf Seite 524-529

Der Beitrag beschreibt einen linearen HF-Verstärker mit Abstimmautomatik und 5 kW Oberstrichleistung. Der Verstärker ist für Speisung aus einem Steuersender mit einer Oberstrichleistung von 500 W über den Frequenzbereich 2,5 . . . 25 MHz ausgelegt. Im Verstärker ist ein von Diskriminatoren gesteuertes Nachlaufsystem vorgesehen, durch das er selbsttätig in wenigen Sekunden auf die erforderliche Frequenz abgestimmt wird. Die langsamste Verstimmung dauert 10 Sekunden, die durchschnittliche Kanaländerung 5 Sekunden.

Ein einfacher Phasenmodulator von C. T. Kohn

Zusammenfassung des Beitrages auf Seite 530-533

Ein aus drei Reaktanzen und einer Röhre bestehender, einfacher Phasenmodulator wird analysiert. Optimale Arbeitsbedingungen werden abgeleitet und die Einflüsse einer nichtlinearen Röhrencharakteristik besprochen. In einem praktischen Beispiel wurden mit 48 Grad Hub 10% Frequenzverzerrung erreicht.

Der CIRRUS-Koinzidenzspeicher von I. R. Butcher

Zusammenfassung des Beitrages auf Seite 534-539

Ein Koinzidenzspeicher mit einer Kapazität von 16 384 19-Bit-Worten und einem konservativen 6-µs-Zyklus wird beschrieben. Die Konstruktionsaufgabe war die Schaffung eines zuverlässigen Speichers mit mässigem Aufwand, wobei auf grosse Geschwindigkeit weniger Wert gelegt wurde. Die X- und Y-Ansteuerungsschaltungen beruhen auf dem Belastungs-Multiplexschalter und nutzen die Treibertransistoren wirtschaftlich, aber konservativ. Durch Kompensation der Treiber wird Betrieb über einen Temperaturbereich von 10 . . . 50°C mit breitem Arbeitsspielraum möglich.

Ein Spannungskonstanthalter für Schaltungen mit Kaltkathodenröhren von A. J. Oxley

Zusammenfassung des Beitrages auf Seite 548-549

Konstanthalterschaltungen für Ausgangsspannungen von 150 bis zu 1 000 V, die mit einer Kaltkathoden-Triggerröhre bestückt sind, werden beschrieben. Eine als Stromversorgung für Dekatrons und Triggerröhren gedachte Ausführung gab bei 470 V bis zu 20 mA ab, hat eine Ausgangsimpedanz von 250 Ω und ein Regelverhältnis von 35: 1.

Sekundäre Spannungs- und Stromnormale von M. Pacak

Zusammenfassung des Beitrages auf Seite 550-553

Ein beschriebener elektronischer Spannungs- oder Stromkonstanthalter ist mit einem leistungsfähigen Verstärker und mehreren anderen Einrichtungen ausgerüstet, die die genaue und reproduzierbare Einstellung der stabilisierten Ausgangsgrösse, für die fast ausschliesslich ein Bezugsteiler und eine Bezugsspannung massgebend sind, gewährleistet. Eine stufenweise oder kontinuierliche Regelung kann über den gesamten Bereich von ungefähr 1 000 V oder 1 A bis zu Null oder darunter (Umkehrpolarität) erreicht werden. Mit einer Konstanz und Einstellgenauigkeit von 10^{-4} oder besser kann das Gerät als Spannungs- oder Stromuntersetzer für die Aufnahme von Kennlinien, das Eichen von Messgeräten usw. eingesetzt werden.

Ein transistorisierter Hochspannungskonstanthalter mit Wechselspannungsregelung von I. Izumi und M. Kokubu

Zusammenfassung des Beitrages auf Seite 554-556

Nach einem beschriebenen Verfahren wird ein Wechselstrom bei Niederspannungspegel geregelt und mittels einer Impedanz in Serie mit der Primärwicklung eines Umspanners in die gewünschte hohe Ausgangsspannung umgewandelt. Das Verfahren ist Reglerschaltungen nach dem Gleichstrom-Gleichstromwandler-Prinzip oder magnetischen Verstärker-Prinzip ähnlich, jedoch ist seine Erholungszeit nicht durch Schwingungen oder Netzfrequenz beschränkt.

Eine praktische Schaltung für 200 . . . 800 V, 0 . . . 5 mA mit einem Regelverhältnis in der Grössenordnung von 100 zwischen Null und mehreren hundert Hertz wird beschrieben.

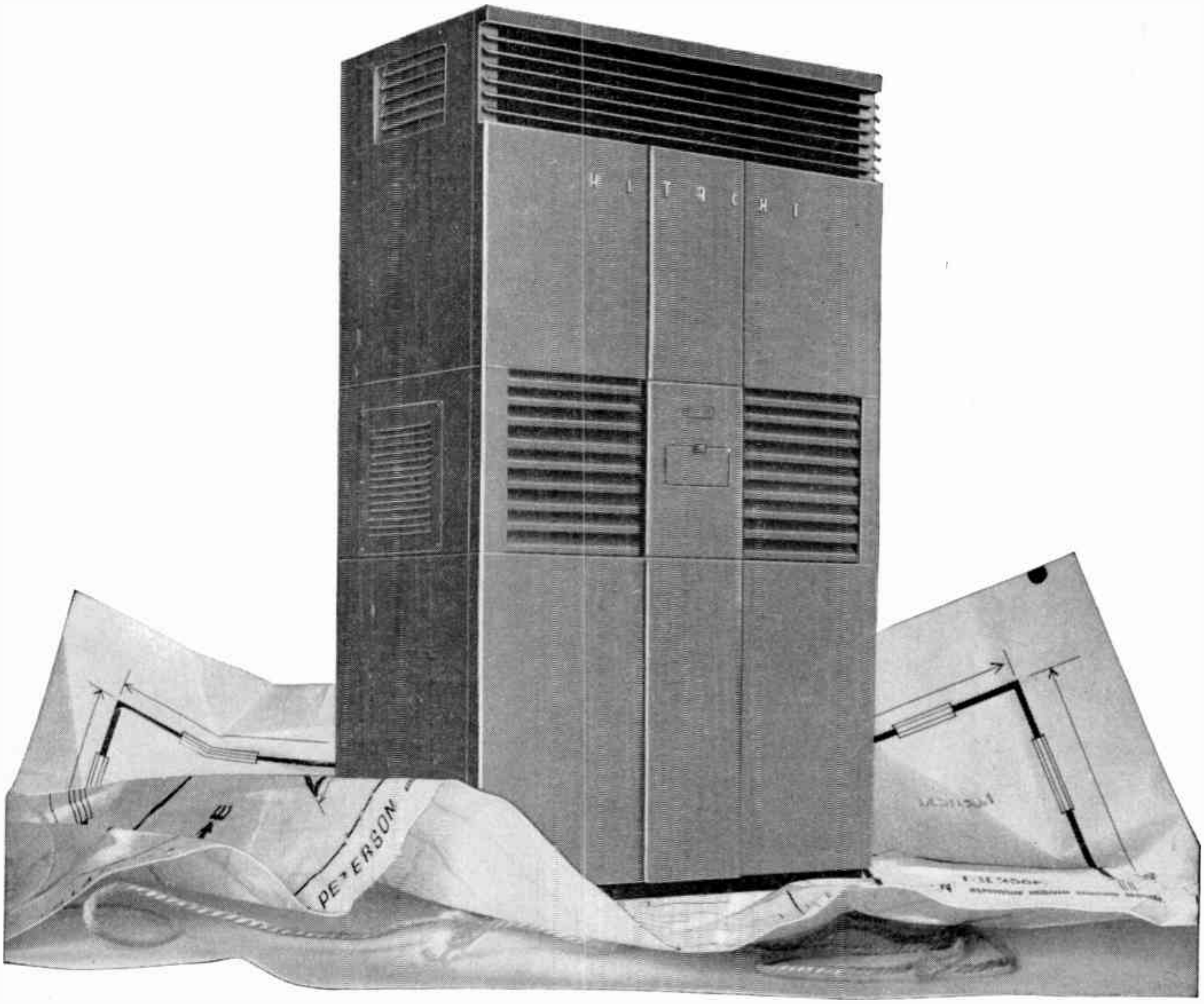
Eine Binärverzögerungsleitung mit Kaltkathodenröhre von D. Q. Mayne

Zusammenfassung des Beitrages auf Seite 557

Eine einfache Binärverzögerungsleitung mit Kaltkathodenröhre, die das Umschalten von zwei Stromversorgungen ausnutzt, wird beschrieben. Die beiden Stromversorgungen ermöglichen die Übertragung eines Binärsignals von einer A-Leitung zu einer B-Leitung und nach Verzögerung von einer Einheit die Rückübertragung zur A-Leitung.

EE 72 130 *for further details*

creating the right atmosphere



It is at the design stage that those responsible should consider in their plans the advantages to be gained by the industrial temperature control and comfort cooling of Hitachi packaged air conditioners. Entirely automatic, compact and vibration-proof, with a low decibel rating, the Hitachi range of equipment is highly adaptable to a wide variety of duct installations. Superior designs, high performance and easy operation are standard throughout as befits the products of Hitachi, one of the world's leaders in technology.

Range of packaged air conditioners from 5 H.P. to 30 H.P. Compact, robust and attractive appearance. Quietest, vibration-proof operation. Supplied complete with plenum chamber. Independently adjustable aerofoil louvres. Fully or semi-hermetic. Automatic control with visual eye indicators and safety controls. 50 cycle electrical operation. Also available, a comprehensive range of air conditioning and water-chilling equipment up to 800 H.P.

UNITEMP by Hitachi

creates the right atmosphere

AIR CONDITIONING AND REFRIGERATION EQUIPMENT

For further details just write your name and address on this advertisement and send it to:

Sole Concessionaires for U.K., Europe and Middle East: **UNIVERSITY AIRCOOL LTD., 1-7 BALHAM HIGH ROAD, LONDON, S.W.12. PHONE: KELvin 3351**

FULLY QUALIFIED SERVICE AND COMMISSIONING AGENTS APPOINTED THROUGHOUT THE UNITED KINGDOM