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Commentary

THE Radio and Electronic Component Manufacturers' Federation held their annual exhibition last month and they have also recently issued, to their members, their annual report for the year ending March 1957. Both provide considerable food for thought.

For several years it has been apparent that the Great Hall at Grosvenor House provides inadequate accommodation for an industry, and consequently an exhibition, which has grown so greatly in the last decade. To alleviate this overcrowding it was necessary either to move to a new building altogether or to find additional space in the Park Lane area.

The R.E.C.M.F. exhibition has, so to speak, grown up in Grosvenor House, and one does not lightly disrupt so successful a tradition. This year, therefore, and possibly only as a temporary expedient, the exhibition was split between two venues. The success of this reorganization, from an exhibitors point of view, is not yet fully known, but, from the visitor's viewpoint, it did not seem altogether satisfactory. For this type of arrangement to be really acceptable it is necessary for the two halls to be close together and for the two sections to be approximately equal in magnitude. In the exhibition under review neither of these conditions were met and the net result was a little disappointing. The Federation can, however, be relied upon to find a satisfactory solution to their present exhibition difficulties and also to play their part in any major reorganization of the electronic industry's exhibition programme that may take place in the future.

However, whatever the difficulties of accommodation, there can be no denying the interest and the high quality of the components on show. The diversity of components becomes, year by year, greater and greater, while the components themselves become smaller and smaller. Miniaturization, components for use with transistors and printed circuits were in fact the dominant theme of this year's exhibition, and it is clear that both transistors and printed circuits are accepted now as standard components and are in no way regarded experimental or developmental techniques. Another point that could not fail to impress was the high standard of engineering finish shown by the majority of the components and equipment exhibited.

The Annual Report of the R.E.C.M.F. contains some extremely interesting facts and figures concerning the industry, and of these probably the most significant is that last year, for the first time, the output of the capital, or professional equipment group exceeded in value the output of the domestic or entertainment equipment group. It must be admitted, of course, that last year was not a good one upon which to base judgment since certain classes of trade, particularly the domestic market, were adversely affected by the 'credit squeeze' with its attendant hire purchase restrictions. Nevertheless, it is fairly certain that this trend will continue for many years to come and that the capital and professional equipment side of the industry will ultimately become by far the most important side of the industry as the use of electronics in its various forms becomes increasingly used as an aid to production and control in other industries.

From the national viewpoint this is a welcome trend since there is a good overseas market for capital equipment, whereas the market for domestic radio and television receivers is comparatively small. In fact, only about 1 per cent of the television receivers produced in the past year was exported and more than half of these went no farther than the Channel Islands or Eire! The overall picture of the export market is, however, very encouraging. In the year under review the total value of the exports of the radio and electronic industry was £40·7m which is an increase of almost 20 per cent over the previous record made in 1955. This overall increase of nearly 20 per cent compares with an increase of 13 per cent for the whole of the electrical industry and an increase of 9 per cent in the whole of this country's exports. Another point worthy of note is that within the electrical industry radio and electronic products now form the largest exporting group. The value of £40m comparing with some £35m for electric cables and wire and about £32m for generating machinery.

The story of the growth of the electronic industry in the last decade and of its achievements, particularly in the export market, make encouraging reading for all connected with it. There is, however, another and rather blacker side of the story and that is, in common with all other manufacturing industries, the increase in production failed lamentably to keep pace with the increase in wages and this, assuredly, is a trend that must not only be stopped but reversed, if the country is to prosper in the highly competitive world markets of today.

* * *

In last month's issue it was stated that this issue would contain a preview of the forthcoming Instruments, Electronics and Automation Exhibition and a review of the Physical Society's Exhibition. Rather more space has been devoted to the I.E.A. Exhibition than was anticipated and the review of the Physical Society's Exhibition has, therefore, been omitted, but the majority of the new equipment exhibited will have been described in the preview of the R.E.C.M.F. and I.E.A. Exhibitions.

A Crystal-Controlled Signal Generator

With an output of 5W at 937·5 Mc/s.

By M. T. Stockford*, B.Sc.

A crystal-controlled signal generator delivering 5W output at a frequency of 937·5Mc/s, has been developed for the purpose of driving a VX8175 klystron. A detailed description of the equipment is accompanied by circuit diagrams and operating instructions.

A VERY stable 937·5Mc/s source is required as part of the test equipment for the VX8175 klystron[1], which has been developed in this laboratory. For test purposes the short term frequency stability required is better than $1{\cdot}2 \times 10^{-6}$/sec, while long term stability is not critical. To achieve this stability, a 12·5Mc/s crystal oscillator is used, shielded from draughts and operating at room temperature. Two quintupler stages and a coaxial line tripler stage with associated buffer amplifiers provide the final output of 5W at 937·5Mc/s.

The whole signal generator has been designed conservatively for long service to eliminate the need to replace aging valves. For example, at 312·5Mc/s about 17W are available to drive the final tripler stage which requires only about 6 to 7W drive, thus allowing the power to fall by about 4dB before valve replacements are necessary.

The elimination of unwanted sidebands is achieved to a large extent in the high-Q coaxial-line tripler stage and they are at a sufficiently low level in the output for driving the VX8175 in our application.

Circuit Design

The Crystal Oscillator

In order that certain circuits might be used in a microwave frequency standard, a somewhat similar equipment being developed concurrently, the oscillator frequency was specified as 12·5Mc/s. An STC Type 4013 crystal in an evacuated glass bulb on a B7G base was chosen. The approximate temperature coefficient of this crystal is 8 parts in 10^7/°C and hence to achieve the required frequency stability, the temperature must not vary more than 1·5°C/sec. Consequently, only shielding from draughts is required.

The oscillator circuit of Fig. 1 was recommended by Standard Telephones and Cables Limited for their crystal.

The microwave frequency standard, however, requires greater stability than that achieved with the above circuit and it must incorporate an oven-controlled crystal of an inherently more stable type. One suggestion for this equipment is to use a 500kc/s crystal as the reference standard followed by four quintuplers to 312·5Mc/s. At this stage, or at 937·5Mc/s, signals at 1Mc/s, 25Mc/s etc., derived also from the standard crystal, are to be injected and mixed in an X-band crystal mixer unit. The resultant harmonics generated are to be used for the calibration of X-band wavemeters and for the provision of standard frequencies in the band. Consequently, the work carried out here on frequency multipliers is of considerable interest and assistance in the frequency standard project, and the experience gained should enable many difficulties to be avoided.

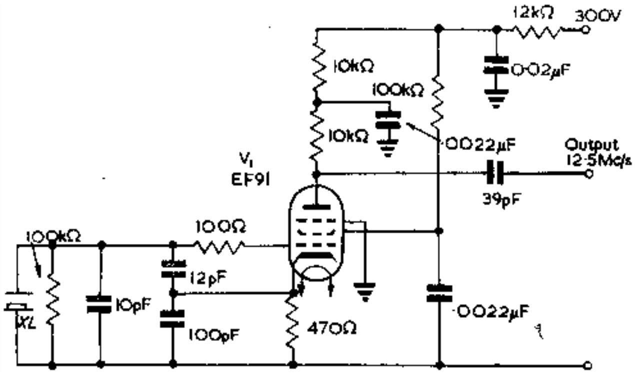


Fig. 1. Crystal oscillator circuit

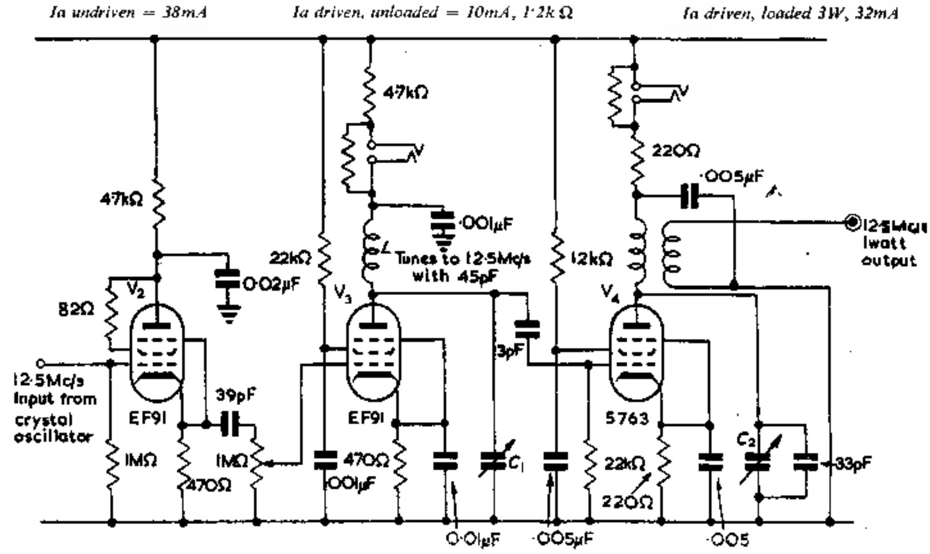
Frequency Multipliers and Power Amplifiers

Since this equipment is for laboratory use only, no attempt has been made to economize in the number of valves in the circuit and the whole emphasis has been on reliability (Figs. 2, 3 and 4). Two quintuplers are used to multiply to 312·5Mc/s followed by a coaxial-line tripler to 937·5Mc/s. For generating fifth harmonics, the anode current must have only a very small angle of flow which necessitates very high grid bias and drive voltages. Consequently, at each frequency level in the multiplier chain an amplifier is inserted.

Fig. 2. Buffer and output stages, 12·5Mc/s

To prevent the 5763 from t.a.t.g. oscillation, input grid capacitance = 3pF, copper screen across valve base to reduce Cga

Ia undriven = 38mA *Ia driven, unloaded = 10mA, 1·2kΩ* *Ia driven, loaded 3W, 32mA*



* *Mullard Ltd.*

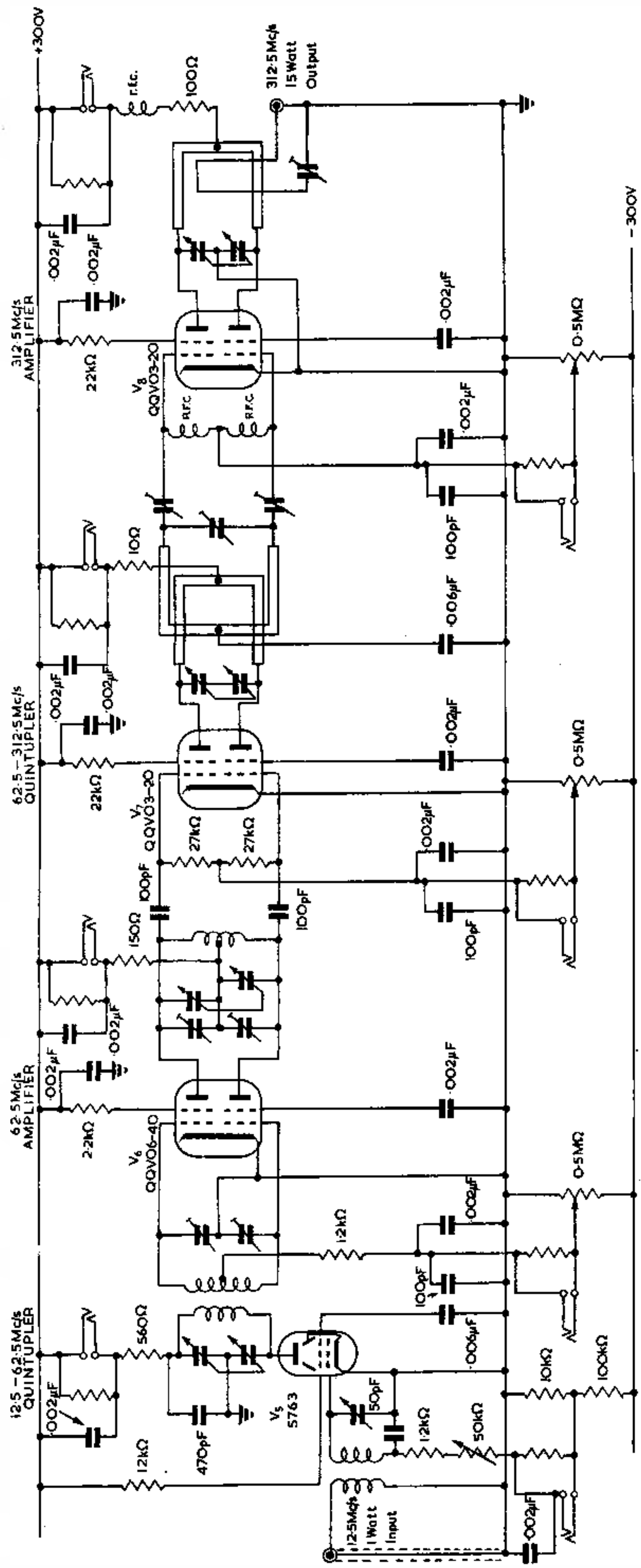


Fig. 3. Multipliers 12·5 to 312·5Mc/s

At 12·5 and 62·5Mc/s conventional lumped circuits are quite suitable and the only special precautions taken have been to prevent the QQV 06-40 amplifier from oscillating at a very high frequency with a tank circuit composed of the anode lines, the tuning capacitor and back along the chassis. This danger can be eliminated by not earthing the rotor of the split stator anode tuning capacitor or by taking it to earth via a resistor of a few hundred ohms in value.

To achieve a high Q at 312·5Mc/s, lecher line circuits are used in the anode circuit of the second quintupler and in the 312·5Mc/s amplifier.

The valves used have been chosen on grounds of present availability though it appears now that the Mullard QQV 03-10 (or QQV 03-05 when available) might well be used in place of the 5763 quintupler. This would have the added advantage that this stage too could be readily balanced to earth and hence help to reduce even further the 12·5Mc/s sidebands produced in the final output.

For the final tripler stage, a CV397 (DET24) coaxial line circuit is used with grid and anode lines both $\lambda/4$ in length (at 312·5 and 937·5Mc/s respectively).

Power Monitors

To assist in the lining-up procedure, each stage after the 12·5Mc/s output stage has a crystal detector circuit loosely coupled to the anode tank circuit. The rectified output of these detector crystals is proportional to the power at that frequency. The crystal detector used is a BTH germanium crystal Type CG4C and the typical circuit used is shown in Fig. 5. It was found essential to screen the two tuned circuits coupled to the 312·5Mc/s lecher line circuits due to the very considerable radiation fields.

Power Measuring Gear

For this purpose, Phillips festoon lamps have been found suitable, even at 937·5Mc/s. To facilitate matching a fairly high impedance lamp is preferable—12V at 6W or even 24V at 6W. To measure power a photocell connected to a galvanometer is placed in close proximity to the lamp. The galvanometer current due to r.f. power in the lamp is recorded and then 50c/s a.c. power is fed to the lamp until the galvanometer reading is repeated. Assuming that r.f. and 50c/s power produce equivalent illumination, the r.f. power is known.

The Stabilized Power Supply

A supply of +300V d.c. at 500mA and −300V at 5mA is required by the complete signal generator. The h.t. supply has been electronically stabilized using six EL37 valves as the series element and an EF91 as the parallel valve (Fig. 6) and the bias line has been derived from high stability neons. The heater power required by the generator is 6·3V at 10A and care has been exercised to ensure that fall in heater voltage supply due to lead resistance losses is kept to a minimum.

The h.t. voltage may be adjusted by means of R_{14} and the meter M_1 checks both h.t. and bias voltage lines.

Lining-up Procedure

This is carried out in the light of the results shown in the section dealing with operational circuit conditions. In

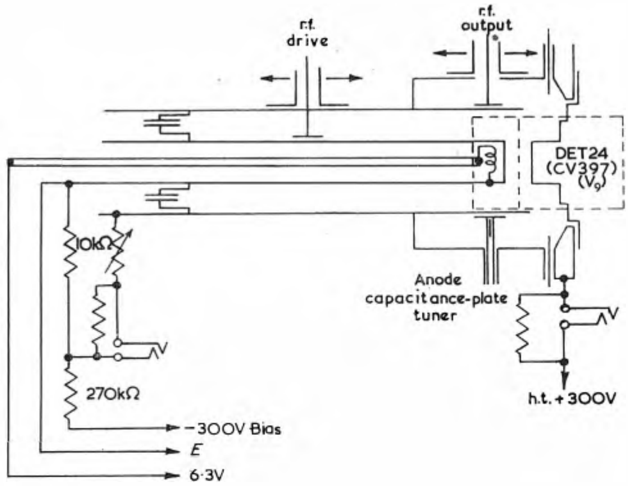


Fig. 4. Multiplier 312·5 to 937·5Mc/s

general, however, the equipment may be systematically lined up by tuning a series of recessed controls, to obtain maximum readings on the power monitors coupled to stages V_{5-8}, or alternatively to give maximum grid current in stages V_{6-9}.

(1) V_3 and V_4 anode circuits are tuned to give maximum dip in V_4 anode current. This may also be checked by observation of V_5 grid current which should be optimized.

(2) V_5 grid circuit is tuned for maximum grid current and V_5 anode circuit for maximum reading of V_5 power monitor.

(3) V_6 anode circuit is tuned for maximum reading of V_6 power monitor.

(4) V_7 anode circuit is tuned for maximum reading of V_7 power monitor.

(5) V_8 anode circuit is tuned for maximum reading of V_8 power monitor.

(6) V_9 grid current is observed and by adjustment of V_8 anode tuning, V_9 grid tuning and position of grid input probe it is adjusted to maximum value.

(7) With output load connected to Telconnector output socket, the anode circuit is tuned for maximum power output while at the same time adjusting load matching to optimum condition.

With the equipment now lined up, the preset bias controls are adjusted at the back of the main V_{5-8} chassis to give

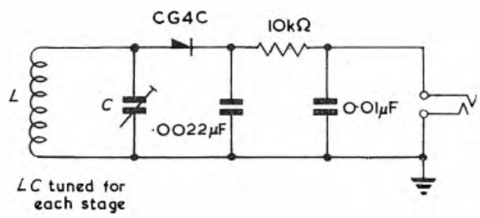


Fig. 5. Power monitor

approximately the conditions shown in the section on operational circuit conditions.

General Layout

The signal generator is rack mounted in five separate units in an enclosed cabinet. Ventilation louvres are provided and an extracting fan is mounted at the top of the unit.

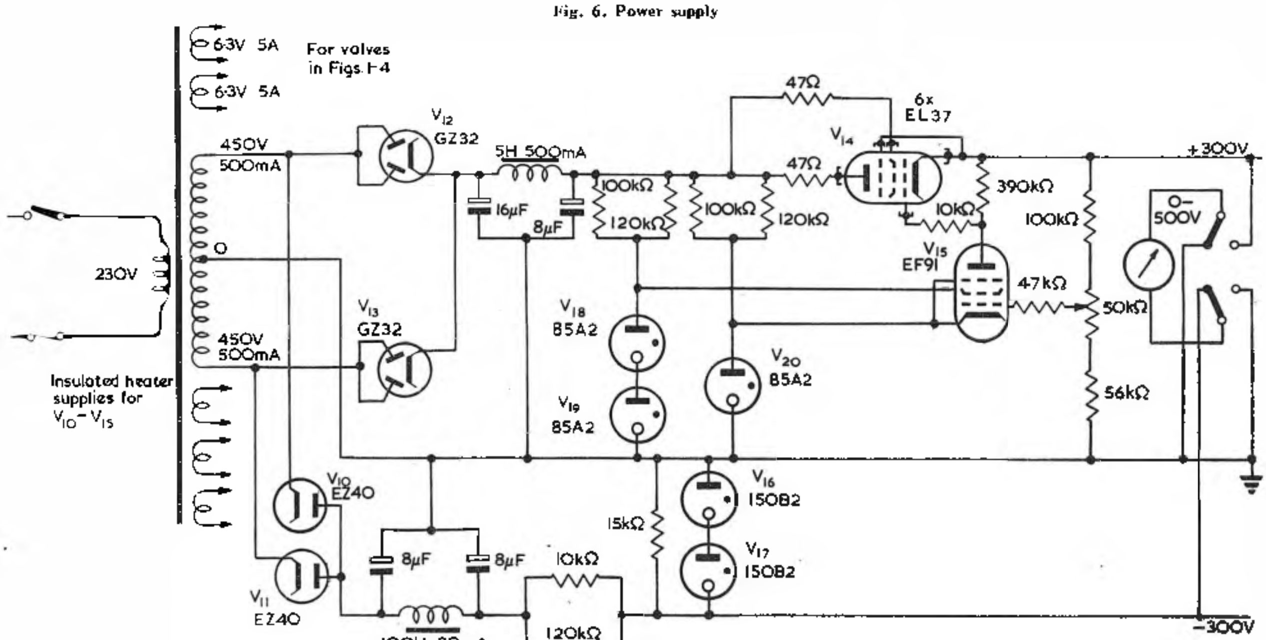
The five units, in order from the bottom of the rack are as follows:

(1) Crystal oscillator with buffer and output stages. These are situated in a separate enclosed case to eliminate draughts. Power supply input and 12·5Mc/s power output connectors are provided.

(2) Power supply unit with three power output connectors.

(3) Multiplier chassis comprising valve circuits V_{5-8}. Power supply input connector, 12·5Mc/s power input connector and 312·5Mc/s power output connector are provided. Screening is provided between the grid and anode circuits of V_{6-8} as shown in Fig. 7 and the four power monitor circuits are coupled to their respective anode circuits. The 312·5Mc/s power output is coupled into a screened

Fig. 6. Power supply



coaxial feeder by means of a tuned loop, coupling into the lecher line tuned anode circuit of V_8. To economize on size, the lecher lines used at 312·5Mc/s consist of strip copper curved backwards towards the metal screening through which the anode lines of the valve protrude. These copper strips have been silver plated to improve circuit efficiency.

Fig. 7. Multiplier chassis 12·5 to 312·5Mc/s, top view

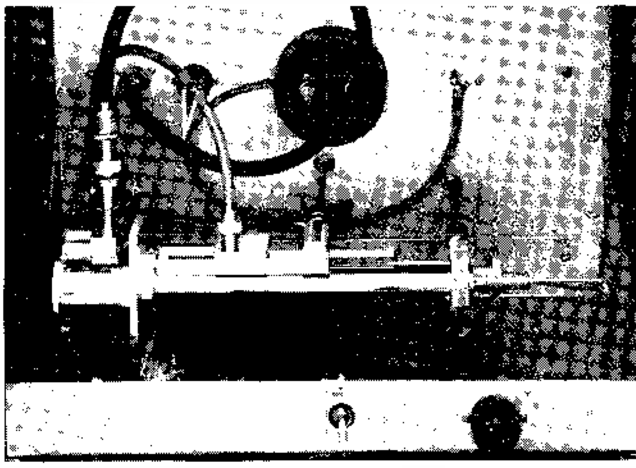


Fig. 8. CV397 coaxial line tripler unit

(4) Coaxial line tripler (Fig. 8). The grid input feeder has a small capacitance plate at the termination of the inner conductor to facilitate coupling; in practice the inner conductor is normally in contact with the cathode line of the tripler. The outer conductor of the coaxial line feeder is isolated from earth by capacitance in order that grid voltage may be applied to the tripler.
The power output connector is a 'Teleconnector' socket Type 53S/29M and again by means of capacitance feeds, no direct connexions are made to the live framework of the tripler unit.

(5) Mains Panel, with on/off switches, synchronous clock and indicator lamp.

Typical Operational Circuit Conditions

Crystal Oscillator and 12·5Mc/s Power Amplifier

The maximum power available from this unit is 4W at 12·5Mc/s with $I_{av9}=4·7$mA and $I_{av5}=31$mA. The set level control is adjusted to reduce this power level to about 3/4W, at $I_{av4}\simeq 37$mA, which is sufficient drive for the 5763 quintupler stage. Under these conditions the 5763 stage can still be sharply tuned in both anode and grid circuits. In the event of overdriving, the sideband content is greatly increased at 62·5Mc/s.

12·5 to 62·5Mc/s Quintupler, 62·5Mc/s Amplifier, 62·5 to 312·5Mc/s Quintupler, 312·5Mc/s Amplifier

In setting up V_{5-8} the preset bias on each stage is adjusted to prevent overdriving. For V_7, the 62·5 to 312·5Mc/s quintupler, an operating bias of about −230V is required to give a suitable angle of current flow and this is considerably in excess of the manufacturer's quoted maximum V_{g1}. The circuit has, however, operated for over 1 000 hours under these conditions. The readings are given in Table 1.

TABLE 1

V_5		V_6		V_7		V_8		Pout
I_{g1} (mA)	I_a	I_{g1} (mA)	I_a	I_{g1} (mA)	I_a	I_{g1} (mA)	I_a	(W)
1·78	34	0·81	92	2·6	65	4·2	102	16·1
—	—	—	—	—	—	5·0	106	17·6
The power output from V_7 is about 2 to 3W.								

312·5 to 937·5Mc/s CV397 Coaxial Tripler

In order to limit the grid dissipation of the CV397 (rated at 1W maximum), the bias on V_8 has been increased as shown in Table 2.

TABLE 2

V_8		V_8 POWER MONITOR	V_9			
I_{g1} (mA)	I_a	(W)	I_{g1} (mA)	I_a	Pout (W)	Pgrid (approx.) (W)
1·18	66·0	14·0	10·0	67·0	4.7	1·1
2·38	89·0	16·0	12·2	70·0	6·0	1·6

The power has been measured in a 12V 6W Philips festoon lamp in a tunable coaxial mount.

The grid input power and grid dissipation cannot easily be measured. An estimate of the grid dissipation is obtained for small angles of current flow by assuming that the current pulses are approximately rectangular. Under these conditions grid dissipation is the product of peak grid volts and average grid current.

Check on Sideband Content of 937·5Mc/s Output

The 937·5Mc/s output, suitably attenuated, was fed to a coaxial crystal tape CV2226 mounted in an X-band mixer unit and the X-band output was displayed by means of a spectrum analyser. An approximate measurement of sideband content was made for the 10^{th} harmonic 9 375Mc/s. The power in the 9 063Mc/s sideband was approximately 20dB down on the power at 9 375Mc/s.

Conclusions

If it had not been decided initially to use a 12·5Mc/s crystal oscillator as reference frequency, greater purity of output might have been obtained by using frequency triplers. In general triplers are easier to balance to earth and, due to their larger gain, inter-stage power amplifiers would probably be unnecessary. From the point of view of elimination of sidebands, the Q requirement of anode tuned circuits in tripler stages is far less stringent than for quintuplers.

Acknowledgments

The Author thanks the Admiralty, the Manager of Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this article.

REFERENCE

1. Norris, V. J. The Frequency-Multiplier Klystron. *J. Electronics 1*, 477 (1956).

A Continuous Amplitude/Time Analyser
for a Band of Low Frequencies

By W. A. P. Young*

The article describes a two channel bandpass amplifier and integrator for frequencies between 16c/s and 30c/s, and their use in the analysis of the faster waves in the electroencephalogram. Output pulses are written on the e.e.g. chart to give a continuous record of the amount of activity in the pass-band.

SOME recent work in the Department of Electro-encephalography and Electro-physiology has shown the need for a reliable device to measure the amplitude/time summation of brain rhythms in a band of frequencies around 20c/s. A knowledge of the absolute input frequency was not required, provided it was known to lie within specified limits. The requirements were briefly as follows:

(a) A pass-band from about 16 to 30c/s.

(b) An attenuation of at least 20 times for frequencies of 12c/s and below.

(c) Stable operation with freedom from frequent re-setting.

(d) As few 'running' controls as possible for the operator, preferably none.

(e) A display mechanism which would write unobtrusively on the moving paper of the e.e.g. with freedom from inking troubles.

(f) The apparatus to be easily portable, as it would have to be used on different e.e.g. machines.

(g) It should be possible to analyse two channels simultaneously.

It was considered that apparatus built up to conform to the above would be more suitable than the analyser[1] already in use, particularly as regards points (c) to (g).

The e.e.g. machines in use (Ediswan 8-channel) have an available output, amplified from the patient's head, and the analyser is designed to work from this.

General Description

The apparatus has two separate channels, each of which consists essentially of three parts, a band-pass section followed by an integrator which feeds a write-out mechanism. There is also a power unit supplying both channels. The whole is built into a box 22in by 10in by 11in high and weighs about 35lb.

Five parallel-T networks[2,3,5] each with its associated amplifier are used for the band-pass section with staggered tuning[4]. Each stage uses a pentode valve with negative feedback, giving a stable gain and a low impedance output to feed the network. By this method it was hoped to avoid the use of stabilized h.t. and heater supplies.

The band-pass amplifier feeds what is essentially a capacitor charging via a diode and resistor, but the d.c. starting level of the diode is kept nearly at the potential of the capacitor by means of a cathode-follower valve. This means that the capacitor continues to charge even with low voltage inputs. The capacitor voltage controls a trigger firing at a predetermined level and producing a pulse which discharges the capacitor via a diode switch. Thus the rate of firing is determined by the amplitude/time product of the input.

The discharge pulse is amplified and connected to the write-out mechanism, which uses two pen arms (one for each channel) of the same type as those in the main e.e.g. machines. They are attached to a simple 'door-bell' mechanism which responds to pulses from the amplifier, giving small 'pips' on the moving paper when energized. The ink is supplied by narrow-bore tubing from a small vertical cylindrical reservoir with a second close fitting cylinder for a lid. A small hole in the top of the lid is closed by the finger and this allows ink to be pumped into the pens to start them operating or sucked back when the apparatus is out of use. The pens are built on to a base plate with rubber feet which can stand on the e.e.g. machine so that the pens mark near the edge of the moving paper. The inkwell is attached to an extension arm which hangs over the edge of the machine top so that the level can be adjusted for correct ink flow.

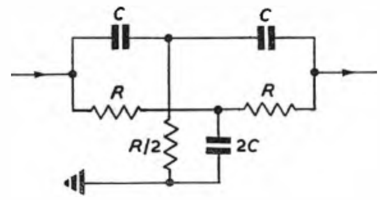


Fig. 1. Parallel-T network

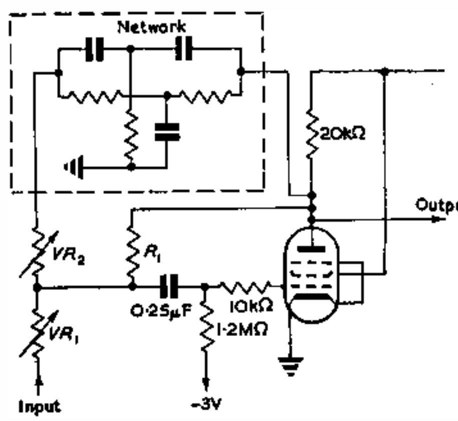


Fig. 2. Basic amplifier stage

The Circuit

The input from the e.e.g. machine is fed to the amplifier via a channel-selector, a switched attenuator and a pre-set gain control.

The twin-T networks were calculated from[2,3,5] $f_o = (1/2\pi RC)$ (Fig. 1) and the Q of the stage from $Q = (A+1)/4$ (approx.), where A is the gain of the amplifier.

The amplifier stages (Fig. 2) have negative voltage feedback applied via R_1 and the T-network loop gain is controlled by VR_2 which is preset; the input/output gain is set by means of VR_1. This arrangement allows of direct

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connexion to the input side of VR_1. Five similar stages are connected in cascade.

The resonance frequencies and Q's of the networks were calculated[4] from

$$d_1^2 = \frac{4+\delta^2-\sqrt{(16+6{\cdot}472\delta^2+\delta^4)}}{2} \qquad \delta^2 = (\alpha_1-(1/\alpha_1)^2+d_1^2$$

$$d_3^2 = \frac{4+\delta^2-\sqrt{(16-2{\cdot}472\delta^2+\delta^4)}}{2} \qquad \delta^2 = (\alpha_3-(1/\alpha_3)^2+d_3^2$$

with two stages staggered at $f_o\alpha_1$ and f_o/α_1 of dissipation factor d_1 two stages staggered at $f_o\alpha_3$ and f_o/α_3 of dissipation factor d_3 and one stage centred at f_o of band width β where

band-centre frequency (geometric mean) = f_o

overall bandwidth = β

and $\beta/f_o = \delta$ also $Q = (1/d)$

The results in this case where $f_o = 22{\cdot}5$c/s were

f (c/s)	30·8	27·5	22·5	18·4	16·5
$(=1/d)\ Q$	5·1	1·9	1·5	1·9	5·1
RC (ohms-Farads)	0·00517	0·00579	0·00707	0·00863	0·00964

From this it can be seen that the maximum gain required is of the order of 20 times, which is easily obtainable with a good margin for feedback. The twin-T networks were made up from available components (5 per cent high stability resistors in the region of 100kΩ and 20 per cent capacitors) which were matched up on a small bridge to about ± 1 per cent.

Battery bias is used and with this the finished circuit gives a band-pass curve in close agreement with that calculated (Fig. 3). Small differences can be adjusted out easily by means of the 'Q' controls.

It was decided that a single 3V battery to provide bias for both channels (10 valves) was simpler and less bulky than an automatic bias arrangement. The latter was tried with a common cathode resistor, 75Ω, and 2 000μF capacitor for 5 stages together, but this produced a 'dip' at the low frequency end of the flat top as would be expected. This could, however, be trimmed out by adjustment of the Q controls.

The integrator section (see Fig. 4) consists basically of the capacitor C_3 charging via R_4 and V_{13}, the input being supplied from the cathode-follower V_6 via C_2. The cathode-follower V_7 keeps the cathode of V_{13} slightly positive to the voltage on C_3 via V_{14}. The value of the resistor R_5 is fairly critical, since if it is too high the bias of V_7 is high and C_3 will not charge from a low voltage, and if too low, grid current charges C_3 in the absence of a signal. Using the value shown, the capacitor will charge at less than 2V p-p input and the drift rate is less than 1 per cent of the maximum rate. V_7 also feeds the cathode-coupled trigger $V_{9\text{-}10}$ and VR_7 is set for this to fire at about +125V. The pulse from V_9 (about 750μsec duration) is fed to the grid of the cathode-follower V_8 where it is limited to +150V by V_{17}. This pulse is used to discharge C_2 and C_3 by means of the diodes $V_{15,16}$ and set their voltage to +150V, the starting point. The cathode-followers $V_{6,8}$ are necessary in order to form a low resistance discharge path for C_2 and C_3 so that these can be fully discharged in a short time. This time (about 750μsec) is so short compared with the maximum trigger rate used (about 17/sec) that the total time lost due to discharging introduces negligible error. The regulator tube V_{11} provides a convenient starting voltage for C_2 and C_3 and a stable reference potential for the trigger. The peak-to-peak maximum input voltage at C_2 is approximately 125V, i.e. 150V less the minimum voltage of the cathode of V_7, since this d.c. restores C_2.

The voltage from V_7 cathode is shaped and amplified in V_{12} to provide a pulse for the write-out mechanism. The screen of this valve is interrupted in the output socket to avoid damage should the apparatus be switched on with this plug disconnected. The heaters of $V_{6\text{-}10}$ and $V_{13\text{-}17}$ are all connected to +150V. This also helps to minimize the effect of heat-to-cathode leak on $V_{15\text{-}16}$.

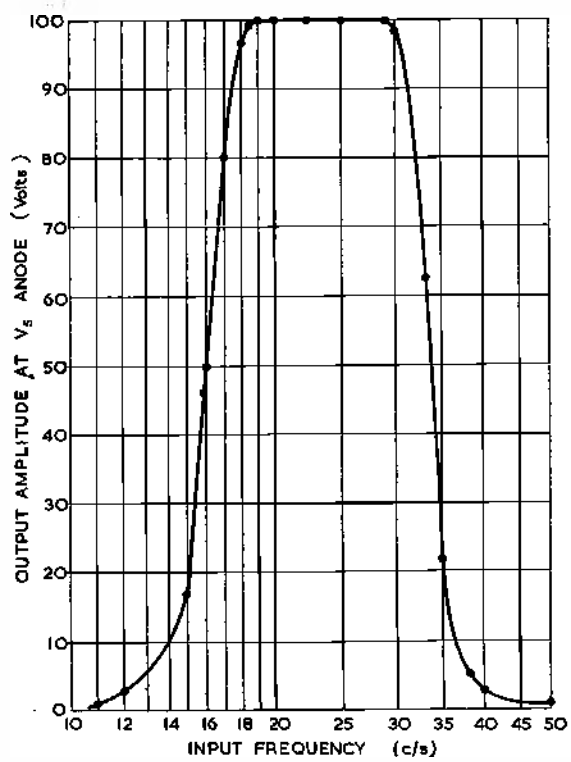

Fig. 3. Band-phase section frequency response

The power supply in conventional, and separate decoupling is used for each band-pass section to avoid interaction between the channels.

Setting Up

The 'Q' values of the bandpass section are set up with an oscillator which is connected via a resistor of about 2MΩ to the junction of $R_{2,3}$ (Fig. 2) for each valve in turn. The oscillator frequency is observed when the output at the anode of the stage being adjusted has fallen to 0·707 of maximum, hence Q, from $Q = f_o/\Delta f$ and R_2 is adjusted for the correct value.

The integrator section is adjusted by means of VR_7 so that the d.c. level observed at the cathode of V_7 runs from 150V to 125V or so. This is not at all critical, the limits being set by uneven firing of $V_{9,10}$ when the voltage swing is too small and the maximum permissible p-p input voltage is reduced as the swing increases.

The input attenuator is marked '½, 1, 2' and with this in the centre position it has been found convenient to set the input preset gain controls to give about 10 'pips'/sec second for a steady 20V p-p input in the centre of the pass-band.

Performance

The stability seems good, as no resetting has been necessary during two months' use.

When tested, a mains change of 10V produced less than 1 per cent change in the output rate when the input signal was set in the pass band, and when set on a steep part of

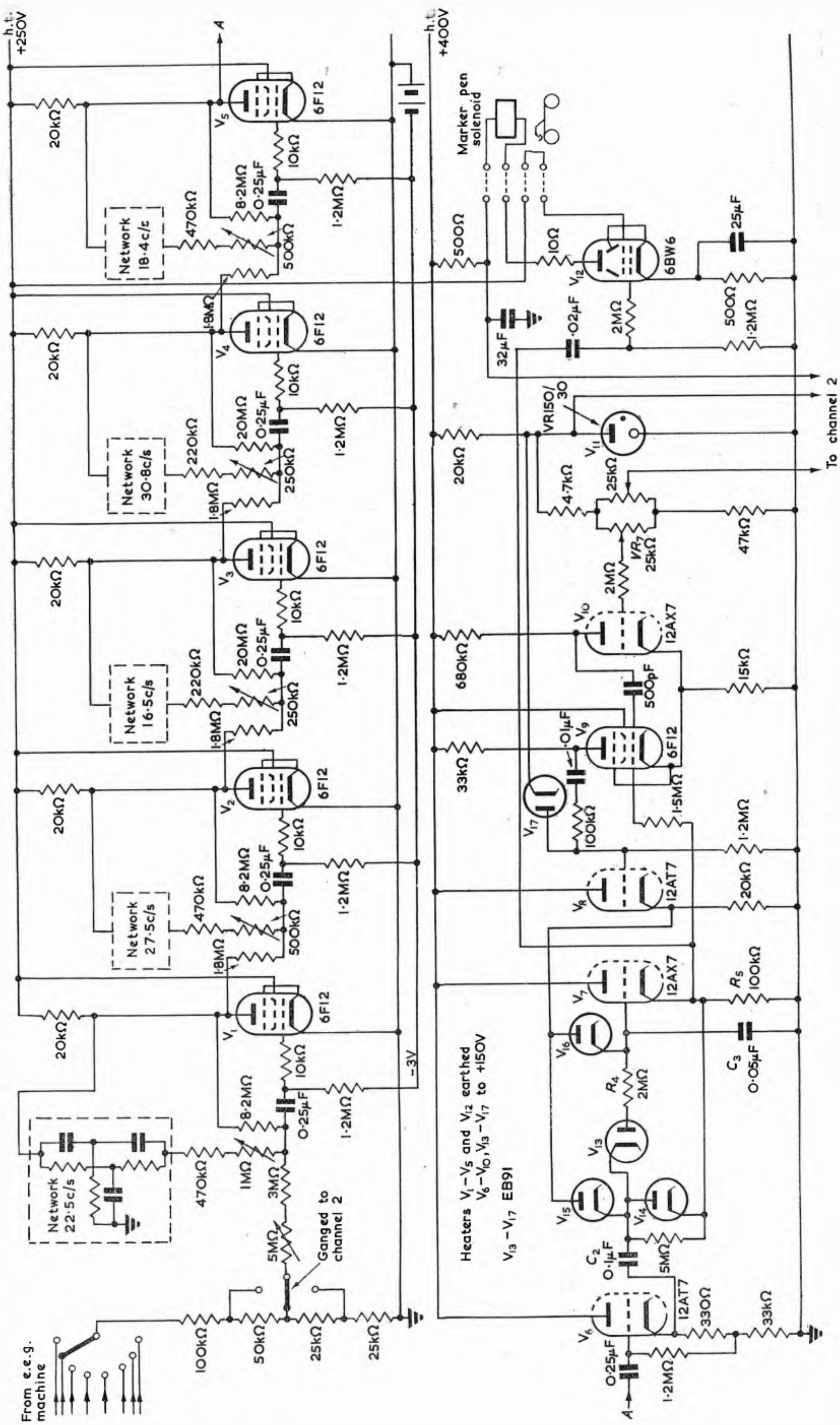

Fig. 4. The complete circuit

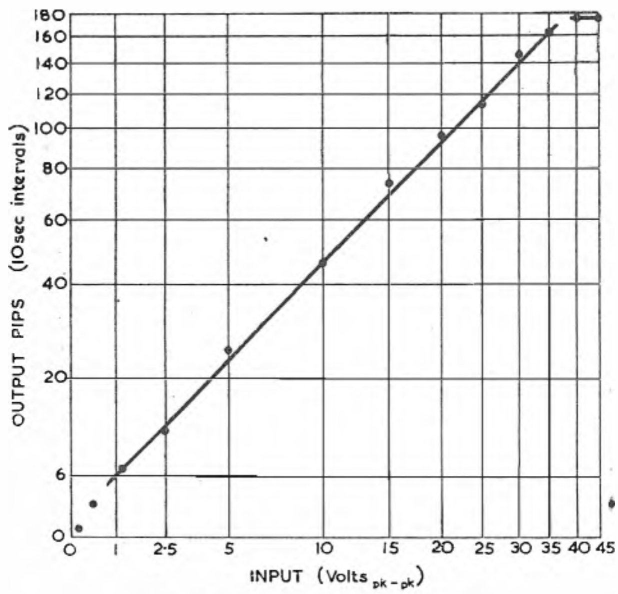

Fig. 5. Overall amplitude response

the curve (6dB down) there was less than 2 per cent change.

Fig. 5 shows the input/output linearity, and it can be seen that a range of about 40:1 is provided.

Operation

The operator selects the two channels to be analysed and after a few minutes of recording sets the input attenuator to give a convenient output in the range 15 to 30 pips in 10sec. Since these 'pips' have to be counted visually, and the expected increase is about four times, these limits are convenient in that the lower gives counts of sufficient accuracy while the upper is not too laborious.

The present use of the apparatus is to determine the increase of activity in the beta frequencies (e.g. waves above 14c/s) while giving drugs intravenously to psychiatric patients. A standard solution of the drug is given, injection being made rapidly at 60sec intervals, with the channels set to analyse the output from two specific areas of the patient's head. After a convenient interval, one channel is set to a different position to compare with the other (standard) channel.

After the record is finished, the number of 'pips' over a period (say 10sec) can be counted and in this way the activity over particular parts of the skull can be compared with the standard, or each other, at any time after any particular injection.

Acknowledgments

Thanks are due to Dr. E. C. Turton and Dr. C. P. Seager for their advice and encouragement, to Mr. M. C. Jackson for his assistance with the construction of the device and to the staff of the e.e.g. department for their patience in using it.

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The Metrovick Digital Computer

A new electronic digital computer has recently been introduced by the Metropolitan-Vickers Electrical Company at Trafford Park, Manchester.

The new unit, called the "Metrovick 950", is a general-purpose model capable of application to a wide variety of mathematical problems in research and engineering design. The first "950", completed some months ago, is now in experimental use in the Company's Electronics Department, whose product it is.

The design of this computer incorporates interesting features of modern technique, of which two in particular are worth special mention: first, the use of printed wiring on the plug-in boards; and second, the extensive use of transistors instead of thermionic valves. These and other features of design and manufacture allow this computer to be offered at a competitive price.

The following description is arranged according to the five basic parts in any digital computer, viz.—

(1) Input equipment, for feeding information into the machine

(2) The store, for holding instructions and data

(3) Arithmetical circuits, for performing the calculations

(4) Output equipment, for giving the results

(5) Control unit, to co-ordinate the whole machine so that instructions are obeyed correctly and in the right order.

(1) Input. The input of numerical data and the programme of instructions are coded into five-hole punching on a paper tape, which is fed into the computer and is there read off photo-electrically at high speed.

(2) The Store. A magnetic drum constitutes the main store. A total of 4096 'words' can be stored on its 128 separate circumferential tracks, each track storing 32 words and the mean access time to any address on the drum is one-half of a revolution period, i.e. 10 milliseconds.

In addition to the main store of 4096 words, several stores with capacities of one or two words are needed in the arithmetical unit.

A further store holding eight words on a regenerative track is provided to enable instructions to be modified after they are taken from the main store and before they are obeyed.

(3) Arithmetical Circuits. These circuits, which use transistors, include an adder-subtractor unit and a multiplier.

Multiplication in the binary scale is merely a series of additions and shifts. For one multiplication the Metrovick 950 multiplier performs eight additions in parallel and does this four times, the sum of the four answers being the product required.

(4) Output. Output is by means of five-hole punched paper tape, or a printed page.

(5) Control Unit. The basic timing of the computer is controlled by signals obtained from the tracks on the magnetic drum. Transistor circuits control the sequence of operations and the flow of instructions and numbers.

All the operator's controls are conveniently arranged on a panel with a desk table in front. Two cathode-ray tube monitors enable the contents of any track of the main store, the B store, and all arithmetical registers to be examined. A set of 32 keys is provided which enable a number or an instruction to be set up, and further keys enable this to be placed in any register as required.

The computer is housed in a steel frame which is assembled in three units, viz. control desk, control and power supply rack, and computing rack. Each of these is designed to pass through a standard doorway.

The computing rack is divided into two sections: the drum store bays contain the magnetic drum store and all its associated circuits; the transistor bays contain the circuits associated with the logical functions of the computer. All these circuits, which comprise basic waveform generators, staticizers, arithmetical units, coincidence units, etc., use transistors and are produced on bakelite boards 8in. by 7in. by the 'printed circuit' technique. The printed boards are carried in ten panels and are plugged into sockets connected in the permanent wiring. This arrangement permits quick removal of the boards for servicing.

The whole of the section of the machine in which printed boards are used has transistors as switching elements in place of the more conventional thermionic valves.

A Two-Phase Low-Frequency Oscillator

(Part 2)

By E. F. Good*, M.A.

The Type CT365 Oscillator

Tuning

Ignoring the small lowering of frequency caused by finite gain in the amplifiers, the frequency of oscillation when the T's of both integrators are varied and kept equal is given by

$$\omega = 2\pi f = 1/CR \quad (27)$$

the same relationship as for a Wien-bridge oscillator. Thus with $C = 1\mu F$ and $R = 1M\Omega$, $\omega = 1$ and $f \simeq 0{\cdot}16c/s$; and with $10\mu F$ and $10M\Omega$, $f \simeq {\cdot}0016c/s$, which is, perhaps, as low a frequency as it is reasonable to generate with this type of oscillator.

At low frequencies the simplest method of tuning is to make the two R's variable; and in a practical arrangement each might consist of a variable resistor in series with a fixed resistor of slightly less than one tenth the value. In this way a continuous frequency sweep of a little over 10:1 is obtained, and several consecutive ranges may be covered by changing the two C's in 10:1 steps. A convenient value for the variable part of R is 155kΩ, so that used with values of C of 10μF, 1μF, and 0·1μF, frequency ranges of 0·1 to 1c/s, 1c/s to 10c/s, and 10c/s to 100c/s, with small overlaps are obtained. When higher frequency bands are used, stray capacitance may become troublesome as already explained, and as a first precaution it is advisable to mount the fixed resistors close to the terminals of the variables.

To obtain low frequencies with low values of capacitance it is necessary to use high values of resistance, and suitable (wirewound) high value variable resistors, say of the order of 1MΩ, are either very large and expensive or non-existent. Methods of overcoming this difficulty, in which potentiometers of moderate value are used in conjunction with high value fixed resistors have been suggested to the author by Mr. P. J. Baxandall, and are shown in Figs. 17 and 18. In the first arrangement, Fig. 17, the introduction of the potentiometer allows the effective T of the integrator to be increased, and so allows large values of T to be obtained without the use of very high value components. A disadvantage, however, is that the reduction in zero-frequency loop gain will result in an increase in d.c. level drift at the output terminals of the amplifiers. In the second arrangement, Fig. 18, this disadvantage is not present; but since the potentiometer reduces the effective T of the integrator, the arrangement is not quite so advantageous for obtaining very low frequencies. In both arrangements there is a 'throwing away' of loop gain in the sense that $q_{(max)}$ is reduced (equation (20)); and therefore when these methods are used, the amplifier gain should be very high compared with the required value of q.

In many applications continuous tuning is unnecessary or even undesirable, and the system of decade tuning (Muirhead-Wigan)[2] suggests itself as suitable. Using stable-carbon (Grade 1) resistors, values of R up to 10MΩ or higher may be used, and very low frequencies obtained with quite moderate sized capacitors. A particular advantage of any system of switch tuning is the high resetting accuracy obtainable.

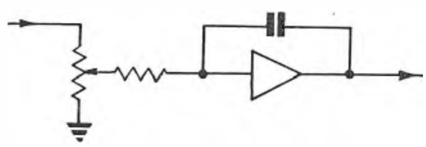


Fig. 17. Method of using a potentiometer to increase the T of a Miller integrator

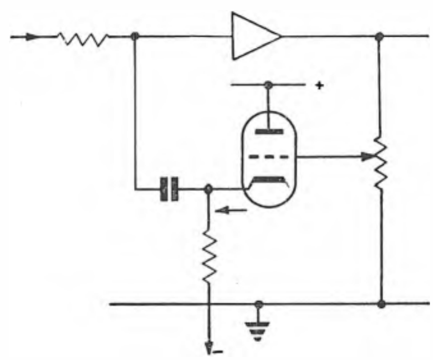


Fig. 18. Method of using a potentiometer to reduce the T of an integrator

When very low frequencies are not required it is practicable to make the C's continuous variables, and this has the advantage that over each frequency sweep the impedances of the tuning components are constant. It is necessary for

* *R.R.E. Malvern*

both terminals of each tuning capacitor to be isolated from each other and from earth, although since the output terminals of the integrator amplifiers, and the inputs (the virtual earths), are low impedance points, there is no need to provide particularly high insulation between the terminals of the tuning capacitors and earth. Leakage resistance between the terminals of the tuning capacitors, however, causes damping as already explained (Fig. 7(b)), and should be avoided.

As equation (9) shows, the resonant frequency of the main loop, and hence the frequency of oscillation, may be varied by varying a single component: for example the gain of the phase invertor may be varied from the usual value, unity, or the T of only one of the integrators may be varied. However, when this method of tuning is used the integrators no longer work always with $R=1/\omega C$, and the amplitudes of oscillation at the outputs of the two integrators and of the phase invertor do not remain equal. It should also be noted that when only one component is varied a relationship of the type

$$\omega \propto 1/\sqrt{R} \quad \dots\dots\dots\dots\dots\dots \quad (28)$$

is obtained.

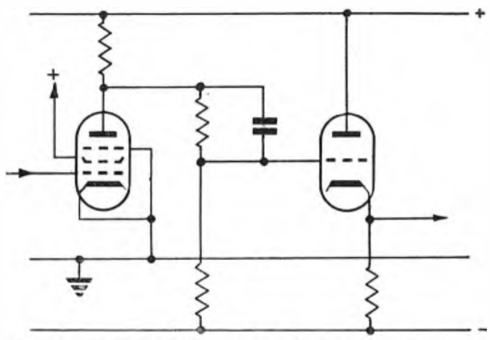


Fig. 19. Simple amplifier with cathode-follower output

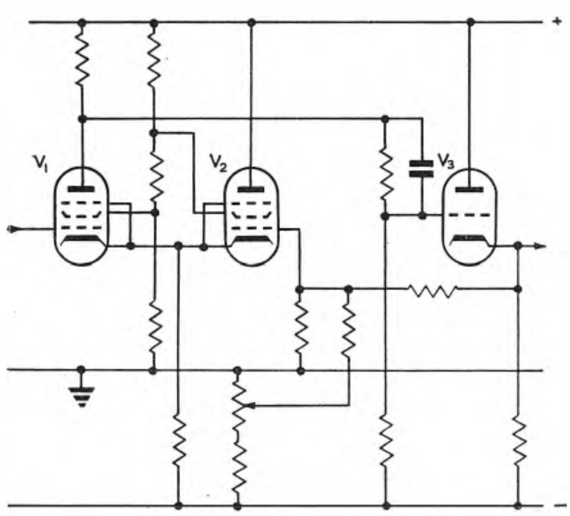


Fig. 20. Amplifier with long-tailed-pair input stage, positive feedback and bias adjustment

Amplifiers

In some applications a single valve will serve for each integrator; but a more generally useful amplifier will include a cathode-follower as shown in Fig. 19, since this allows the input and output of the amplifier to be at earth potential approximately, and provides a low-impedance output for connecting to the several loads. A further refinement is to replace the single input valve by a long-tailed pair, Fig. 20, the spare grid so introduced (V_2) being useful for introducing (a) bias adjustment so that the input grid (V_1) may be brought exactly to earth potential (d.c.), (b) positive feedback from the output of the amplifier to make the amplifier gain, A, approximately infinite. The principal advantages of the latter modification (having $A \simeq \infty$) are that the Miller integrators can now give almost exactly 90° phase shift without having a very high natural internal gain, and that in the phase invertor interactions between the various feedback paths are reduced. The long-tailed pair will also give a useful amount of zero-drift compensation, which will be helpful when a reduction in d.c. level drift is required at one or more of the output terminals.

At sufficiently low frequencies, amplifiers using more valves and having very high gains may be used (i.e. amplifiers of the d.c. computing type). It is doubtful, however, if the use of such amplifiers will give any appreciable improvement in performance, except in those cases where methods of tuning are used which 'throw away gain'. Very high amplifier gain will then be useful in maintaining a constant q and a low output impedance over the tuning range.

The Limiter

When in addition to the square wave fed back to the phase invertor a square-wave output of larger amplitude is required, a squaring amplifier with or without regeneration may be needed; but in other cases a diode limiter similar to that shown in Fig. 14 will usually be sufficient, and will probably take the practical form shown in Fig. 21, the voltage drops across the two resistors r replacing the batteries in providing bias for the diodes. The variable resistor connected as a current diverter across the two resistors r serves to vary the bias, and provides a good means of adjusting the amplitude of oscillation.

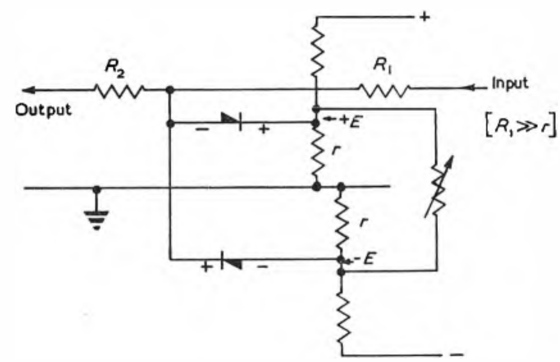


Fig. 21. Diode limiter with bias adjustment for amplitude control

Such a limiter has essentially two states. For low amplitudes of sine wave fed in, the diodes do not conduct and the current fed back is determined by the sum $R_1 + R_2$. For an oscillation to build up, the relative value of this current must be sufficient to overcome the positive damping in the circuit. The other state is when one or other of the diodes is conducting, and is the state existing for most of the time when the amplitude of sine wave fed in is large. In this state a resistance intermediate in value between r and $r/2$ is connected between the junction of R_1 and R_2 and earth, and this resistance should be sufficiently low for the current fed back to the phase invertor to be less (and preferably much less) than the amount needed to overcome the damping existing in the main loop—otherwise the limiting will not be effective, and the oscillation will build up in amplitude until overloading occurs in the amplifiers.

When connecting the limiter to the rest of the circuit, it is desirable to arrange for there to be no bias at the junction of R_1 and R_2 sufficient to cause one or other of the diodes to conduct when the circuit is quiescent. Otherwise the feedback along this path may not be sufficient to ensure that the net damping is negative, and the circuit may fail to oscillate.

Because of the low values of bias used (often only one or two volts), thermionic diodes are not a very good choice for the limiter, since changes in 'contact potential' would give appreciable changes in effective bias, and hence in oscillation amplitude. Point contact germanium diodes are a convenient and economical choice, but these also can give drift in amplitude, in this case because of fall in back resistance with rise in temperature. The best choice at present seems to be the silicon junction diode, which gives a very high back resistance even at temperatures up to 70°C, coupled with a low forward resistance; and does not, of course, suffer from 'contact potential' changes caused by heater voltage variations.

H.T. Supplies

An obvious choice for h.t. is a double stabilized supply giving approximately equal positive and negative voltages (Fig. 22), and such a supply is indeed very suitable. In many cases, however, unstabilized h.t. supplies such as shown in Fig. 23 are quite satisfactory, and are, of course, a considerable economy. It is clear that the smoothing capacitors, while effective in by-passing 'hum' frequency currents, do nothing to reduce the output impedance of the supplies at very low frequencies. And the fact that such supplies are often satisfactory is explained by (a) the relatively small amount of signal-frequency currents in the h.t. lines, (b) the low sensitivity of amplifiers operating with a high degree of feedback to voltage fluctuations on the h.t. supplies.

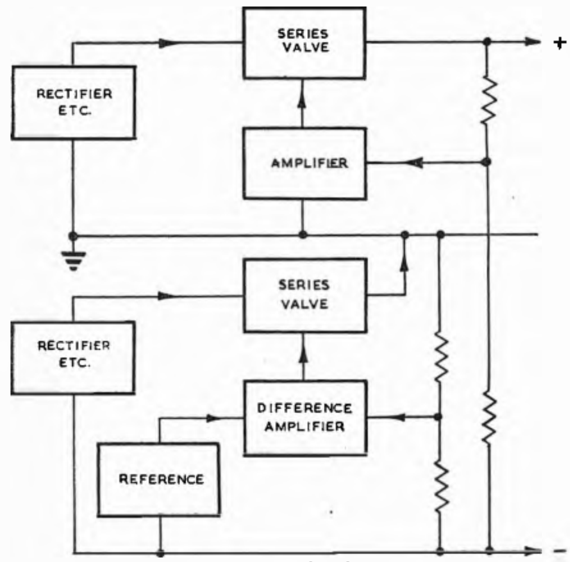
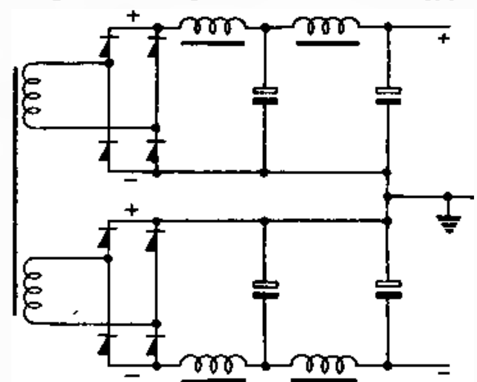


Fig. 22. Block diagram of stabilized h.t. supply

Fig. 23. Block diagram of unstabilized h.t. supply



A Complete Oscillator

Fig. 24 shows an inexpensive version of the oscillator using three double triodes and a double diode. The q is determined by the gain of the amplifiers, and is therefore not very closely controlled—also the phase difference between the two outputs differs from 90° by two or three degrees, even assuming the second integrator capacitor has zero power factor. However, a reliable sine wave output is obtained, at about 20V r.m.s. amplitude, and with less than 0·5 per cent harmonic content at the best output.

A small cathode-ray tube is connected to both outputs, and the circle or ellipse which is drawn is a useful check that the circuit is oscillating and that the waveform has not been spoiled by connecting a too heavy load. The amplitude control is perhaps not connected in the best position, since an appreciable amplitude of oscillation exists at the input of the phase-invertor amplifier, and adjustment of the amplitude control causes some frequency shift.

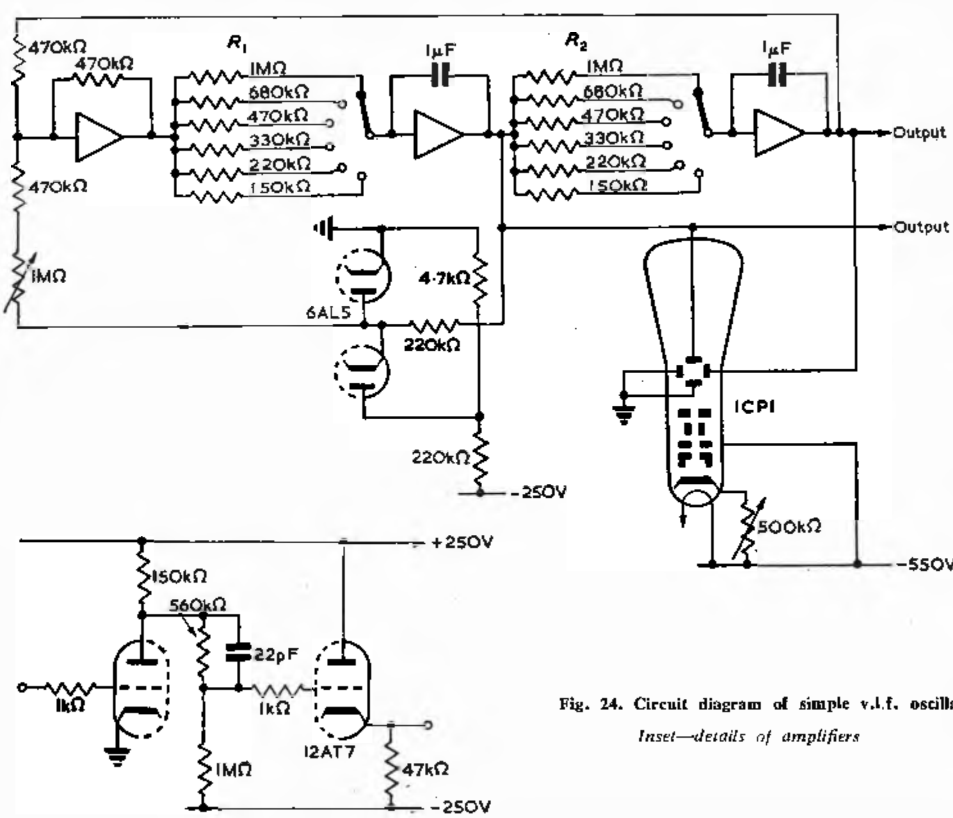


Fig. 24. Circuit diagram of simple v.l.f. oscillator

Inset—details of amplifiers

Conclusion

The two-integrator circuit is well known as a servo system or servo simulator (for example Whiteley[3], and Williams and Ritson[4]), but does not appear to have been used previously as a selective circuit. Its principal advantage over *RC* tuned resonant circuits of the parallel-T type, for example, is that there is no magnification of errors in component values, and that high selectivity can be combined with a predictable and stable value of peak gain or *q*. A possible disadvantage is that rather more valves are needed.

The high selectivity and stable gain make the circuit very suitable for use in a practical variable-frequency sine-wave oscillator in which the amplitude of oscillation is controlled by a square-wave limiter, and such an oscillator is especially advantageous at very low frequencies. The attenuation of the selective circuit for frequencies above resonance is very high, and despite the use of a square-wave limiter, oscillators of the type described can easily be designed to have a lower harmonic content than is usually found in *RC* oscillators of the linear type. An additional advantage of the two-integrator circuit is that quadrature outputs are available.

Acknowledgments

The author is grateful to several of his colleagues at R.R.E. whose interest in the circuits described has been of great value; and particular acknowledgment is due to Mr. E. N. Shaw, who designed the low-frequency oscillator, CT365, which is the first embodiment of the new oscillator circuit to go into production, and which is now being manufactured and marketed by Dawe Instruments Ltd.

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A New Roving Eye

Since the BBC's first Roving Eye was designed and built by the BBC Engineering Designs Department about three years ago, a great deal of experience has been accumulated. This has shown that among its functions it has much more frequent use in doing modest broadcasts such as short inserts into programmes like "Sportsview" and "Panorama", than in doing broadcasts while roving. Roving Eye II therefore, is designed as a small self-contained two-camera mobile unit, but the facility to rove when required, has been retained. An additional feature has been added by the inclusion of a 45ft pneumatically operated extending mast, so that the radio links, which are necessary for roving, can also be used over a considerably increased range when operated from a fixed point.

The new vehicle has been designed to fulfil the following purposes:

1. Transmission of television picture and sound signals from a moving vehicle.
2. Operation as a self-contained two-camera outside broadcast unit.
3. Transmission of pictures at very short notice (e.g. from the scene of a fire or accident).
4. Provision of additional camera channels in conjunction with a normal three-camera outside broadcast unit.

The inclusion of a 'vision mixing' unit permits the selection of the output of the required camera. The picture signal can be transmitted, together with the sound programme signals, to a fixed base while the vehicle is on the move. A built-in power supply is carried, together with the radio equipment for transmitting vision and sound signals to base, and maintaining two-way communication for control purposes.

Experience with the earlier single-camera vehicle has shown that this is admirable for providing an additional camera on an outside broadcast, in conjunction with a normal three-camera outside broadcast unit. However, with independent operation the production techniques are greatly enhanced by using the two-camera operation possible with this new unit.

The cameras used are Marconi Mark 1B Image Orthicon type, these being smaller and lighter than later models. One of the cameras is brought into its operational position by raising it through the roof. The cameraman stands on a fixed platform, which does not rise with the lift, but which is above the floor level of the vehicle.

The second camera is mounted at the near side front of the vehicle alongside the driver, and it can look forward through the windscreen or through the nearside window.

The vision mixing unit makes it possible to fade or cut from one camera to the other, or to mix the outputs of the two cameras. A line sending amplifier is provided for the feeding of vision signals via Post Office lines, when the equipment is operated from a fixed point.

There are three picture monitors (besides those which are incorporated in the camera control units). One of these is at the commentator's position at the front of the vehicle. The other two, which are mounted above the camera control units, include multi-channel receivers, so that they can also be used as radio check monitors. Provision is made for eight microphone channels.

For normal outside broadcast operation, both cameras can be used away from the vehicle, to which they can then be connected by cables, which may be up to a thousand feet in length. Alternatively, the whole of the apparatus comprising the camera channel and sound and vision control equipment can be conveniently removed from the vehicle and set up elsewhere, using a spare cable 'harness'.

Vision signals are transmitted from the Roving Eye to base by a frequency-modulated transmitter operating in the band 610 to 660Mc/s and feeding a six-element Yagi array at an overall height of 16ft when roving. In order to keep the vision transmitting aerial accurately beamed on base as the vehicle manœuvres, the array is controlled by a gyro-compass, which keeps it on a constant bearing through a servo system. The bearing of the aerial can also be varied independently of the gyro-compass to correct for the alteration in bearing, caused by change of position of the vehicle in relation to the base.

Sound programme signals are transmitted to base by an f.m. transmitter which uses an omni-directional roof-mounted whip aerial on a frequency in Band I. This radio link also carries an engineering control speech channel as upper-sideband modulation of a 10kc/s subcarrier, and a 15kc/s subcarrier, which can be keyed for signalling purposes. During rehearsals the sound programme channel is used for 'production control' talk-back from vehicle to base. Another whip aerial receives a Band I signal from base, and frequency-modulated subcarriers on this signal provide production talk-back, engineering control, signalling and a mains frequency locked to the 'grid'.

When the vehicle is not roving, the Yagi aerial and also the single whip aerial serving both the transmitter and receiver on Band I, can be mounted on a telescopic mast which gives an overall height of 45ft. A combining unit enables the Band I transmitter and receiver to share the whip aerial, and includes a 60dB rejector to prevent break-through from transmitter to receiver. The range is of the order of two miles when roving, but when standing still and using the high mast, it is of the order of ten miles.

The complete vehicle

Forming Musical Tone Colours From Complex Waveforms

By Alan Douglas, M.I.R.E.

In this article a survey is made of the principal methods used to derive imitative or 'new' tones from single complex waveforms. Apparatus which makes use of pre-formed tone-colours or is dependent on toneforming by addition of sine-waves is not included.

THE following remarks apply in the main to music generators using valve oscillators. Taking into account one or two discontinued instruments and the electromechanical systems which do not come into this category, there are seventeen examples of valve oscillator music generators available today, some in more than one model. It is not proposed to describe any of these instruments beyond specifying the basic waveforms obtainable therefrom. All use tone circuits embodying the methods described below.

It does not substantially matter whether the object is the formation of new or established tonecolours; modern tendency for some years has been to produce these by suitable shaping circuits from a single complex waveform per key or other pitch source. In this article it is convenient and sufficiently correct to use the expressions pitch and frequency interchangeably. Similarly, although these shaping circuits are very rarely true filters, they are always referred to as such since it has become the common nomenclature in this art.

The main reasons for designing oscillators to provide a pitch note together with a retinue of harmonics are that the method is economical, and that the harmonics are truly in tune with the fundamental. Opinion is divided as to whether the conditions for oscillation should be such as to over-drive the circuit and thus cause harmonic distortion, with the attendant possible lack of stability and dependence on accurate line voltage control; or whether it is better to generate a sine wave and distort the output by additional circuits. Both methods are in use, but the object is the same —to produce the pitch note and as many harmonics as are thought economically desirable.

The circuits notwithstanding, it is highly improbable that any acceptable musical sounds could be produced by economical electrical means were it not for the remarkable accommodation, tolerance and powers of synthesis possessed by the human hearing system. So long as the ear is presented with anything which it is capable of analysing, it will not be unduly irritated. Further, the percentage of people having either accurate or sensitive tonal perception is extraordinarily small.

Musical timbre or tone quality depends on the number and relative amplitude of the harmonics accompanying the pitch note and not on their phase relationship, so long as this state is steady. For this reason, the application of toneforming (or wave-shaping) circuits is made much easier. Electrical tone-forming can be divided into two parts, one based on resonant circuits and one on broader forms of control.

It is surprisingly difficult to produce 'new' tonecolours. Analysis of sounds falling into this category shows that any unusual effect is largely due to the sound starting or stopping in a particular way; a lack of change in timbre with a change in loudness; too great a degree of steadiness in the sound, the use of an invariable vibrato, or the fact that this is often confined to amplitude modulation. An extension of pitch range to very high or very low frequencies may add to the lack of realism, and sometimes extreme purity of the waveform contributes to this effect. But the ear will always attempt to find some similarity between such sounds and one to which it is long accustomed so that the way of the 'new tone' protagonist is hard. At best, most new tones seem unreal rather than new and this is often because they are produced at an excessive loudness level. Very little work has been done on multinote generators in which only new tones are produced as chords; but there is considerable scope here since the effect is entirely different from the sounding of single notes. It must be concluded that at the present time there is not sufficient demand for new tone polyphonic systems. Typical simple examples of unusual sounds are the pure sine wave and the characteristic waveform from a multivibrator.

When we turn to imitative tones, those based on resonant circuits can usually be manipulated so as to have a high degree of realism in a steady state. At the same time, the very properties of such circuits prevent the maintenance of the effect over a wide range of notes. Very often it is desirable to have a range exceeding three octaves, but it must be remembered that the corresponding range of the physical instrument imitated would be unlikely to exceed three octaves; the demand for a wider compass is an artificial situation brought about by the organ pipe, where there is little limit to the range. This problem is entirely different because each independent pipe supplies the required harmonic series and further, is capable of considerable adjustment without this having any effect on other pipes of the same tone quality.

The resonant circuit is an analogue of the formant found in physical instruments. This band of tone-characterizing frequencies is found in all orchestral instruments dependent on resonance, and has been discussed in detail elsewhere[1,2,3]. The closer the coupling of the air to the vibratile element, the narrower and more pronounced the formant band. While this has often been shown in a general way before, Fig. 1 gives an analysis of these bands for some more important orchestral instruments. In this diagram the relative intensities of the instruments are not to the same scale, the figures on the ordinates representing the magnitude of a particular frequency compared with the fundamental; e.g. in the case of the oboe, 880c/s has five times the intensity of the fundamental and for the trombone, 260c/s has four times the intensity. These values are not in any way related to the comparative total energy for each instrument. The formants persist so long as enough power is available to fully excite the resonator, but very often at the upper frequency limits this is not possible in wind instruments; while at the other end, the coupling to the air generally falls off so that the pitch note becomes weak while the formant band remains.

These remarks underline the so far insuperable difficulty in duplicating the light and shade accompanying a general tonal structure and account for the suitability of electrical generators for organesque tones rather than the much more expressive orchestral instruments. With every single note, indeed at almost every instant of sounding, the relative levels of the harmonics and fundamental undergo changes

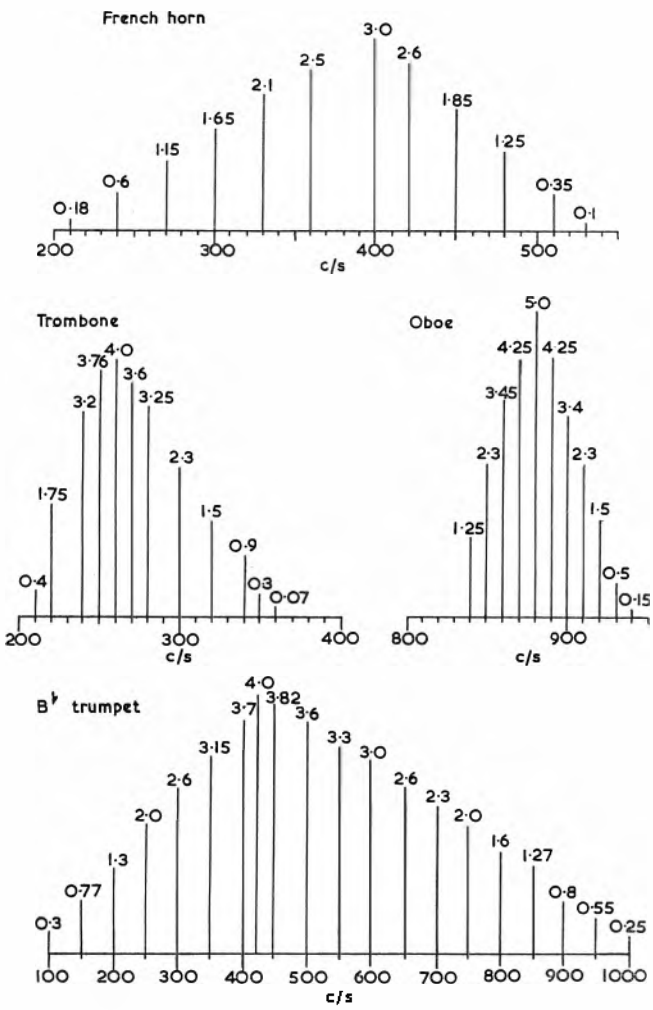

Fig. 1. Analysis of formant frequencies for some orchestral instruments

in intensity. Fig. 2 shows this for a flute, where the removal of just one harmonic markedly influences the tone; whereas the removal of one or more harmonic from a tone such as that of a trombone would make no audible difference whatever. This of course is why formant synthesis of steady state reed and string tone is successful. The actual physical formants are unsymmetrical, but the electrical counterparts are symmetrical. The small discrepancy in harmonic reinforcement due to this property does not unduly affect very complex tones. But, with simpler tones such as the flute, electrical formants are not successful.

Nearly all orchestral instruments (and the organ) contain a narrow band of noise in addition to the tone spectrum. While experiments in injecting noise appear to add some realism to certain kinds of sounds, it is impracticable to do this commercially at a reasonable cost and the degree of noise changes according to the extent of resonator excitation; no satisfactory method for measuring the preferred amount of noise has been devised.

Faced with these difficulties, we have to consider how far and how accurately electrical synthesis can be carried out. For ease of adjustment a sawtooth waveform is desirable. This need not be perfect, the synthesis from two shapes of sawtooth for an oboe being shown in Fig. 3.

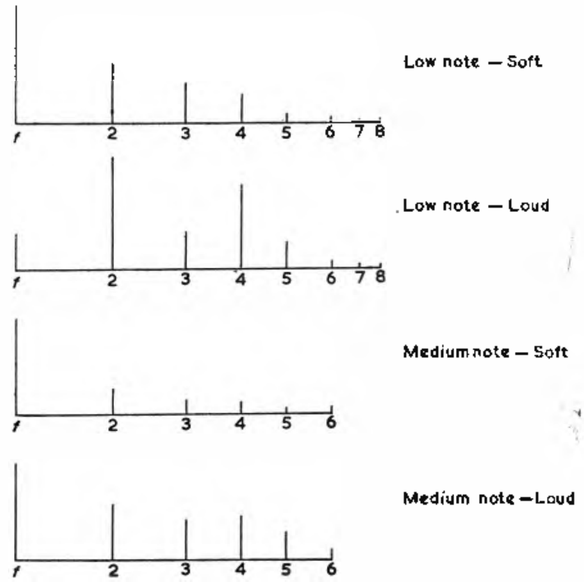

Fig. 2. Analysis of orchestral flute tone

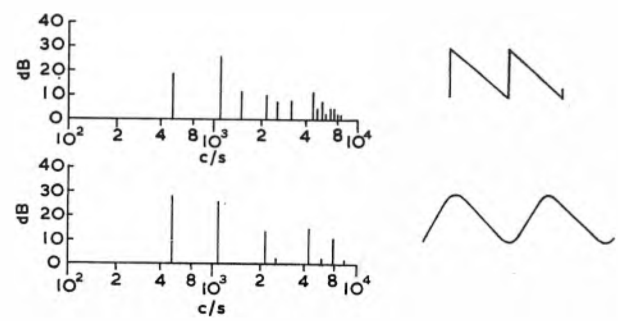

Fig. 3. Synthesis of oboe from perfect and imperfect sawtooth waveforms

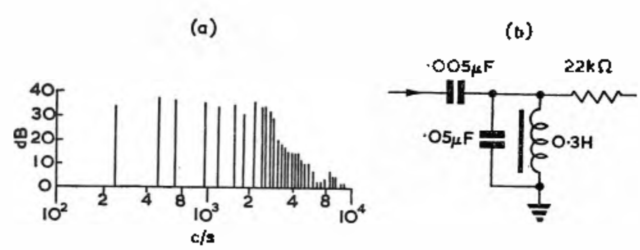

Fig. 4. (a) Oboe, reed alone
(b) Electrical circuit for oboe tone

The difference to the ear is not greater than that between two commercial makes of oboe. The fact is that in this particular instrument the formant band is so strongly defined that the accentuated harmonics obscure any others which may be present. This well illustrates formant theory, for the waveform of an oboe reed *alone* is given in Fig. 4(a). The electrical circuit to form this tone is shown in Fig. 4(b). It resonates at about 880c/s and if the waveform from the generator contains enough harmonics, the response will be quite good over four octaves. Above and below this range, the tone will suffer, although this could be rectified by the addition of extra formant circuits in shunt.

The thin, acid tone of the oboe is very characteristic but there is a marked similarity between the sound of a cornet, trumpet and trombone, so much so that electrical synthesis is comparatively easy. In the spectrum of Fig. 5(b), it will be noted that a large number of harmonics appear at almost the same level. In fact the raising or lowering of any of these by about 6dB does not affect the sound. But a more interesting fact is that we do not require anything like this number of harmonics to produce good reed tone. The general effect of adding even harmonics is to brighten the sound without much affecting its quality; whereas addition of odd harmonics adds stridency and gives an assertive quality. Odd harmonics lying close together within an octave contribute little to a loud sound, but those spaced further apart and in particular, the 7th, 11th and 13th, provoke a great deal of stridency. It could reasonably be said that one could not have too many harmonics if a brilliant sound is the aim, but only if the loudness is correct; for if such a spectrum is greatly reduced in level,

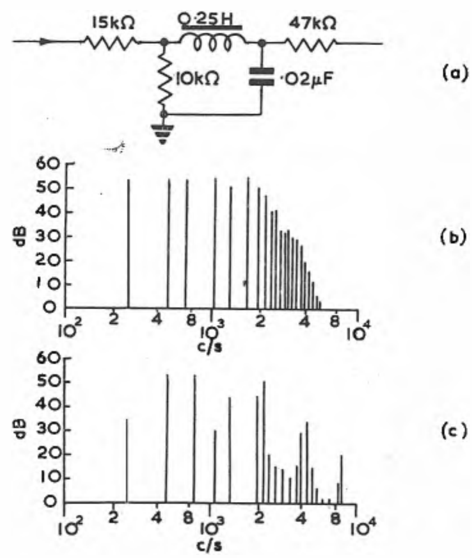

Fig. 5. (a) Electrical formant circuit for trumpet tone
(b) Analysis of trumpet waveform
(c) Analysis of electrical trumpet waveform

the tone quality cannot be distinguished from the upper notes of a 'cello; and for one particular steady tonal state, the tonal spectrum of the 'cello cannot be distinguished from that of a cor anglais in certain parts of the compass. It is the modulation from vibrato and the difference in the noise content which constitute the main differences.

A high Q resonant circuit like that of Fig. 5(a) will simulate the required formant band. Fig. 5(b) is the wave analysis of an actual trumpet and Fig. 5(c) is the electrical counterpart. The similarity is sufficiently close to be deceptive. Although even harmonics are present in this spectrum, the over-riding effect of the odd harmonics is so great that a perfectly good trumpet synthesis can be obtained from a square wave containing no even components at all. The presence of even harmonics makes the tone smoother and a better mixer with other sounds.

The French horn is an interesting problem. There is a greater possible variation in harmonic texture according to the manner of playing than in any other orchestral instrument. Usually it is envisaged as a smooth, agreeable sound and since one can only select a particular steady state for synthesis, this condition (in which the full harmonic range is represented with only slight formant accentuation) is generally chosen; a wise choice, since the more agressive horn tones are difficult to imitate. A sawtooth wave is essential here for if the harmonic development is curtailed the sound will be harsh. One successful electrical circuit is shown in Fig. 6. An extra low-pass RC filter section is introduced here to 'roll off' the top response over the most satisfactory range of the circuits 130·8 to 1 046·5c/s.

Some useful reed-like tones can be obtained from double formant circuits like that of Fig. 7. The harmonic spectrum is so complex that the tone becomes bassoon-like and if kept at a low level it is indeed extremely like that instrument. The principle can be extended to embrace a chain of resonant circuits as in Fig. 8, in which the resonators may be damped by shunt resistors to flatten the peaks.

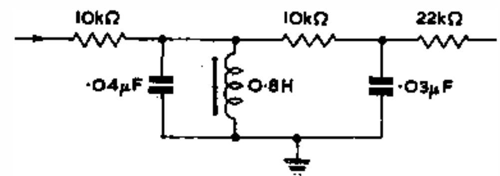

Fig. 6. French horn filter

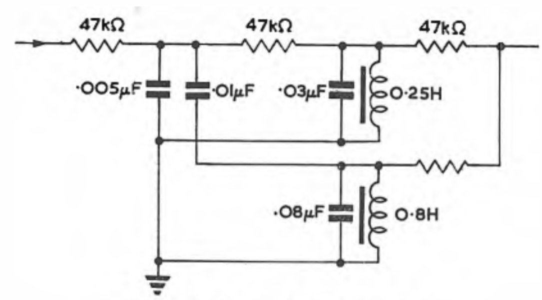

Fig. 7. Double formant circuit

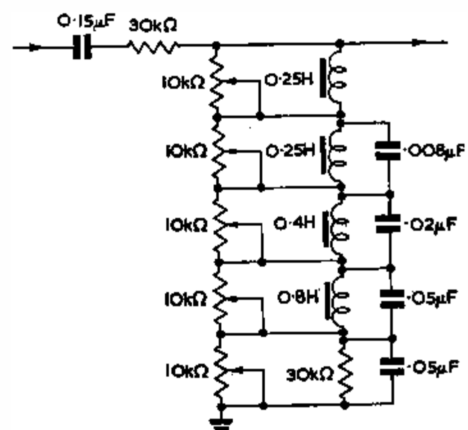

Fig. 8. Cascaded formant circuits in shunt arrangement

Such an arrangement can be made to produce a reasonable imitation of a saxophone, probably because there is no definite harmonic spectrum for this instrument; the coupling of the reed to the air is so loose that the player can make the tone quality veer between that of a clarinet and a horn at will; when a vibrato is added, the spectrum defies analysis.

By complete suppression of even harmonics, that is by using a square wave with a 50/50 pulse width, similar circuits will produce imitations of the clarinet. If tone-colours octavely related to the above examples are contemplated, it must be remembered that the relative levels of the harmonics are *not* the same in those instruments of the same nominal tone quality but lower in pitch, e.g, trumpet and trombone; clarinet and bass clarinet; alto and tenor saxophone, etc. Therefore an elaborate generator system is really necessary, although in commercial instruments a sub-octave coupler is in general use. The results are far from accurate imitations. It is, however, true to say that the positive indentification of a specific reed instrument depends much more, if not entirely, on the starting characteristics of the tone than on the actual waveform and a few simple experiments with a tape recorder will prove this.

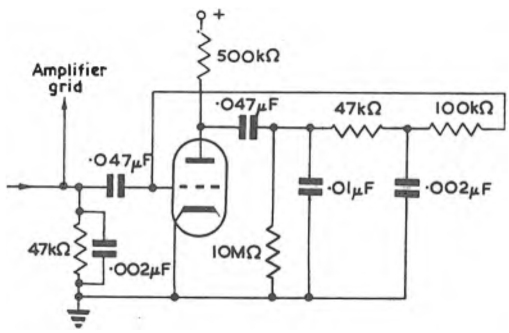

Fig. 9. Phase shift tone shaping circuit

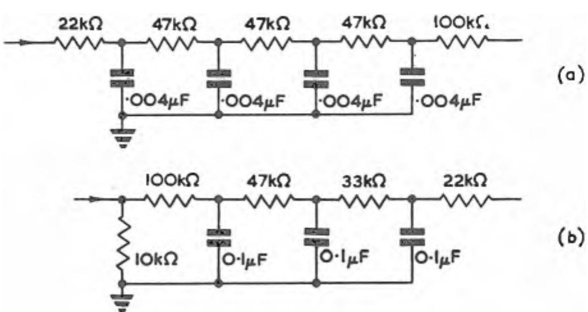

Fig. 10. Organ flute filter

In between the resonant and non-resonant filters we find a different approach in the phase shift circuit of Fig. 9. The object of this device is to produce an impedance between input and earth which changes to a marked extent with frequency. Over a certain range of frequencies the phase of the feedback signal at the grid is such as to reinforce the input signal. At one frequency in this band the reinforcement is a maximum, a peak is produced in the response and the shunting effect on the input is a minimum. At low frequencies the feedback signal phase-shifts so as to oppose the incoming signal, the grid-earth impedance drops and the response falls off. This will continue until the reactance of the input coupling capacitor becomes so high that the signal level begins to rise again. The overall effect is to produce a peak in one frequency band and a dip in another, and by choice of suitable components these may be adjusted over a wide range. The effect is not so sharp as a tuned inductive filter and it may be considered more as a frequency tilting circuit.

Turning now to filters for other purposes, the simplest application is the graded low-pass filter. For most tones of a flute-like nature it is only necessary to reduce upper harmonic development, given a sawtooth input waveform. The spectrum of the orchestral flute is more complex than is usually assumed, even if it is simpler than that of most other instruments; but we have to limit ourselves to the pipe organ types of flute. Clearly, if there is too much reduction of the upper frequencies, the top register will be weak, but if there is not enough reduction of these the sound will be stringy. Over about four octaves good results are obtained with the circuit of Fig. 10(a), and since this is an important filter for organ pedal tones the values to cover from CCC 32·7c/s to F 174·6c/s are given in Fig. 10(b) for a bourdon type of tone.

In certain types of generator it is possible to increase the output from the oscillators at intervals, usually one octave apart, as the pitch range of the keyboard rises. By this means the cutting effect of such filters in the upper registers is reduced and the brilliance maintained. Another way of doing this is to inject frequencies one octave higher into the filter, as in the diapason tone network of Fig. 11. Extra keying contacts add to the complication of this method.

The least satisfactory group of tone filters comprise those used for string tones. Yet this is not so much a consequence of their properties as the fact that string tone is never uniform for a moment. Even in a steady state there is nearly always modulation by vibrato. Organ string pipes bear no resemblance to orchestral strings although they represent a very important group of voices; but their aesthetic appeal is based on their use in chords, and since the tuning is not perfect there are many continual phase changes in the combined spectrum which give a breadth and spacious effect to the sound. The best synthesis which can be obtained derives no benefit from this chorus effect unless there are quite independent oscillators for each note; in this case, fractional mistuning may be permissible and the results are moderately good. For notes fairly low in the scale, the high pass filter of Fig. 12(a) proves adequate. This follows from the fact that the fundamental pitch note is weak in orchestral string instruments towards the bottom of their range. For notes towards the upper end of the range, a resonant circuit as in Fig. 12(b) is probably more effective, but it is important that the level should be low; it is assumed that a sawtooth waveform is available. The pure sawtooth

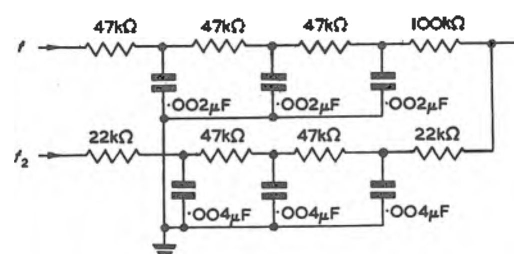

Fig. 11. Open diapason filter

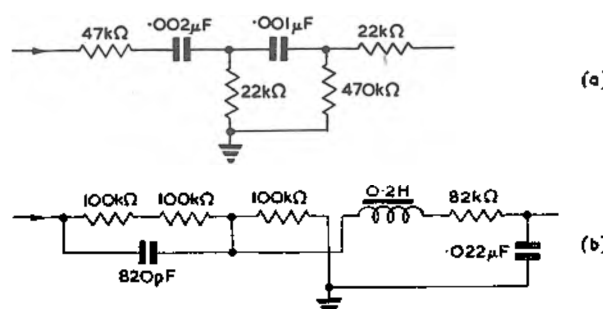

Fig. 12. String tone filter

is not unrealistic, but it is better to remove some or all of the lower components by high-pass filters.

The foregoing represent most of the general methods in use today for forming musical sounds, although there are many other possible circuits. There is no doubt that the work of the Cologne school of research on electrical music will accurately define the minimum permissible spectrum for tonal synthesis in due course, and it might be mentioned in passing that when dealing with one single note, it is already possible to fully correct for all the faults of intonation in orchestral instruments. But where chords are required, no amount of ingenuity can supply the complicated transients which accompany the starting and stopping of a tone. If these were steady and well-defined, it might be possible, but since they are not due to any inherent property of design but rather depend on the manner, rate and degree of the exciting force, it is evident that they constantly change in character as may easily be heard by listening carefully to a single orchestral instrument of any kind. It is the transients which give the expressive intonation to musical tones and we must be content, in a polyphonic instrument, with one kind of transient synthesis. It is most essential to regulate the loudness of the respective circuits to approximate to that of the instruments imitated, or the results may be unsatisfactory; this is because by training and usage we are accustomed to hear certain sounds at well-defined loudness levels, and because the physical instruments themselves have a maximum power output. The ability to increase this by amplification must not be mis-used.

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A Millimicrosecond Time-Base

By M. W. Jervis†, and R. T. Taylor*

A hard valve time-base with thyratron drive is described. The output has a rate of change of voltage 5 × 10^{11}V/sec with reasonable linearity. This is suitable for sweeping typical single stroke cathode-ray tubes in a few millimicroseconds.

RECENT developments have produced cathode-ray oscillographs for displaying single transients with very high writing speeds[1,2,13]. In order to exploit these, time-bases are necessary which will sweep the screen in times of the order of millimicroseconds. A typical oscillograph might have an X-plate sensitivity of 2V per spot width, so that for a trace length of 200 spot widths in 1mμsec a time-speed of 4 × 10^{11}V/sec would be required.

Time-bases of this order of speed have been described, the present article describes a modification of the type originated by Fletcher[3].

In the limiting case for the fastest time-base, the stray and deflector plate capacitance is first charged and then discharged as fast as possible, a valve being used as a discharge device. If a constant discharge current I, is assumed and the total capacitance is C, then:

$$I = C\,(dv/dt)$$

dv/dt being the time-base speed.

For $C = 20$pF

and $(dv/dt) = 10^{12}$V/sec

$I = 20$A

Such currents are easily obtained from thyratrons, and millimicrosecond time-bases using them have been described[4]. Even with hydrogen thyratrons, however, there is a firing delay of about 50mμsec and as this varies slightly it causes a 'jitter' of several millimicroseconds in the triggering[5].

The Hard Valve Circuit

PRINCIPLE OF OPERATION

Hard valves are available which are capable of passing large anode current pulses[6] and these have the advantage of reduced delay and more stable triggering.

The basic time-base circuit is shown in Fig. 1. The grid of the valve, usually a tetrode, is held cut off by a large negative voltage. The anode is fed through a high value resistor from a high voltage supply, the screen supply being fixed and decoupled. A positive pulse is applied to the grid and the anode to earth capacitance is discharged by the anode current. If the tetrode is operated above the knee in the I_a/V_a characteristic, a fairly constant discharge current is obtained, so giving a linear sweep.

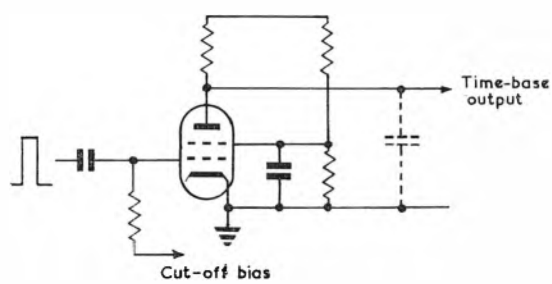

Fig. 1. Basic time-base circuit

DRIVING PULSE

The driving pulse applied to the grid of the discharge valve must be large enough to overcome the cut-off bias and of low output impedance to drive the grid positive by about 200V. With large discharge valves the cut-off bias will be between 150 and 500V and the grid current several amperes, the values depending on valve type and electrode potentials.

The rise-time of the driving pulse must also be short, otherwise the valve will start to conduct before maximum drive is available. This would introduce a delay and has the effect of reducing the anode voltage when maximum drive is applied.

A suitable driving pulse can be obtained using transformer or thyratron arrangements.

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The transformer arrangement has the advantage that the discharge valve can be pulsed between grid and cathode (bootstrap drive) with a resistor connected in the cathode lead. The voltage across this resistor provides a positive going sweep so giving a push-pull output. The transformer can be fed from an amplifier[7] or made part of a blocking oscillator[8,9]. With such an arrangement a time-base with a delay of 45mμsec and a jitter of 0·1mμsec has been obtained[9]. In an arrangement described by Connor[10] the transformer feeds a cathode-follower which linearizes the sweep. The cathode-follower output feeds a push-pull output stage consisting of a tetrode with cathode and anode

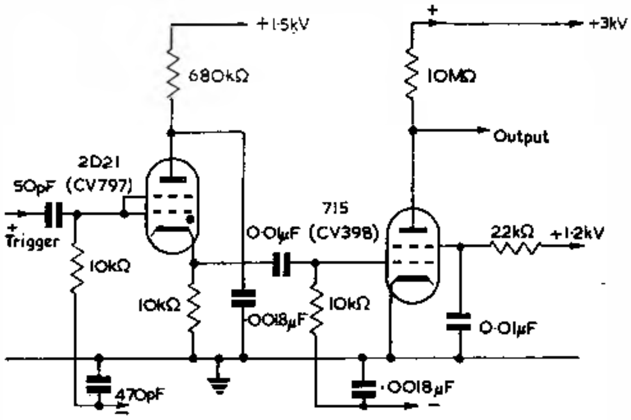

Fig. 2. Final time-base circuit

loads, so giving positive- and negative-going sweeps. A 600V scan of 2mμsec is obtained with this circuit.

The thyratron drive described by Fletcher[3] and Smith[11] has the merit of simplicity, however. The basic circuit is shown in Fig. 2, a small tetrode thyratron type 2D21 (CV797) being used to drive a 3D21 discharge valve (see next section). The rated anode voltage is 650V, but this can be considerably exceeded and circuits have been described[12] with anode voltages up to 1 850V. With 930V applied, Fletcher found the firing delay to be about 50mμsec with a jitter of ±0·1mμsec.

The Discharge Valve

The discharge valve is required to pass a large peak current for a reasonable grid drive (including cut-off voltage) together with a small output capacitance. It is also desirable that the heater power shall not be excessive.

Table 1 summarizes the characteristics of the more suitable types; larger valves are reviewed by Glasoe[6].

From this data it can be seen that in spite of the rather large cut-off voltage and heater power of type 715, there is a considerable advantage in the anode current available.

TABLE 1

COMMERCIAL NUMBER	CV NUMBER	RATED PEAK CURRENT (A)	ANODE CAPACITANCE (pF)	RATED ANODE VOLTAGE (kV)	HEATER POWER (W)	APPROX. CUT-OFF VOLTAGE at V_A =3kV (V)
11E3	CV73	3·5*	7·5	3·5	10·5	150
3D21	CV2659	5·0*	10·0	3·5	10.7	150
KT67	CV437	5·7†	9·0	0·6	9.5	150†
QQV06—40	CV2797	5·7†	6·4	0·6	11·5	150†
715	CV398	15·0*	7·5	15·0	52·0	500

* Manufacturers' curves show that currents about 50 per cent higher than this are obtainable under some conditions.

† Data given by Connor[10].

This was confirmed experimentally by measuring the anode current with the thyratron drive arrangement used in the final time base (Fig. 2).

Time-Base Circuit

The 2021 thyratron drive and 715 tetrode were used in the circuit shown in Fig. 2, the voltages being as shown. The available sweep is 1·25kV.

The speed of the time-base was measured from the oscillogram shown in Fig. 3. The oscillograph[13] used was one in which no precautions were taken to prevent light from the electron gun from reaching the photographic plate. This light is responsible for the white patch in the middle of the trace.

The Y-deflexion was the output of a 9·1kMc/s magnetron and the X-plate sensitivity is such that one cycle on

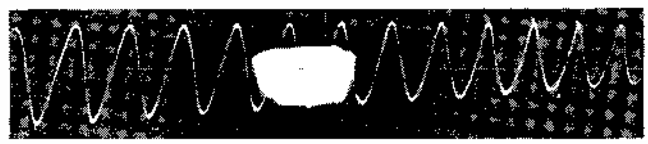

Fig. 3. Resultant oscillogram

the oscillogram corresponds to 60V. The time-base speed is thus $60 \times 9{\cdot}1 \times 10^9 = 5{\cdot}5 \times 10^{11}$V/sec.

Assuming a total capacitance of 20pF and a discharge current of 15A, the theoretical speed would be $7{\cdot}5 \times 10^{11}$V/sec. Using CV73 or CV2659 valves, speeds lower than this by a factor of 2·5 were obtained.

The speed of the time-base can be reduced by inserting a resistor in series with the grid of the discharge valve. If a variable resistor is used, this provides a convenient time-base speed control. With the circuit described, a 5kΩ resistor reduced the speed by a factor of 10.

Conclusion

The time-base described gives a reasonably linear sweep of 1·2kV with a speed suitable for sweeping typical cathode-ray tubes in a few millimicroseconds. The measured speed is in fair agreement with the theoretical figure.

Acknowledgment

The authors are indebted to Dr. T. E. Allibone, F.R.S., Director of this Laboratory, for permission to publish this article.

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Dynamic Methods of Testing Semi-Conductor Rectifier Elements and Power Diodes

(Part 2)

By A. H. B. Walker*, D.I.C., B.Sc.(Hons), A.C.G.I., M.I.E.E.,
and R. G. Martin*, B.Sc.(Eng.), A.C.G.I., B.Sc(Econ). Grad.I.E.E.

Some Approximate Dynamic Test Methods

The special instruments described in Part 1 permit the accurate measurement of true dynamic characteristics under any desired operating conditions. However, simpler methods using normal instruments are very often necessary for general acceptance testing. Since a standard of comparison is now available, these approximate methods can be considered and the errors which they are likely to introduce can be estimated. In several cases direct comparison of methods has been undertaken and it has been shown that it is impossible to form a universal judgment on any 'approximate' dynamic test method, since the errors which it introduces depend very much on the characteristics of the rectifier being tested. This of course emphasizes the danger of such approximate methods; successful results with one type of diode may easily produce over-confidence in the method and serious errors may later arise in using it to test other diodes. In the following discussion it is assumed in all cases that each arm of the arrangement comprises one element or diode only.

Approximate Methods Requiring Four Diodes (Bridge Methods)

Where the diodes are supplied, or are to be used, in a bridge circuit this method is attractive since it provides the overall dynamic characteristic of the four diodes (or four complete arms) and this is frequently all that is required. However, if the correct measurements are taken, the readings can be reduced to the standard dynamic characteristics 'per diode', although of course an average characteristic of four diodes is necessarily obtained.

Open-Circuit Bridge (Reverse Test)

First the bridge is completed electrically (in many conventional stacks this will involve the common connexion of the two outer connectors) and the rated a.c. supply voltage is applied to the a.c. terminals, the d.c. terminals being open-circuited. The mean a.c. reverse current is measured using a rectifier milliammeter. The scale reading of a normal instrument, which will have been calibrated on sinusoidal current, must be divided by 1·11 to obtain the mean value of the current to which, of course, the instrument responds.

To obtain the mean reverse current per diode, in order to express the characteristic in standard terms, the total mean reverse current of the bridge, as measured, must be divided by four.

If a current transformer is used to feed the milliammeter there is a risk of error should the arms of the bridge be seriously unbalanced in reverse characteristic. In this case there will be a d.c. component in the transformer primary current which will not be reproduced on the secondary side and the mean reverse current reading will be inaccurate. The presence or absence of such an unbalanced d.c. component can be checked by the connexion of a d.c. moving-coil instrument in series with the a.c. supply to the bridge. If this instrument reads zero the current transformer will not introduce an error.

The value of mean reverse current obtained by this method may be lower than the true value given by a full dynamic test (or in actual operation) since no forward current is flowing and there is no junction heating from forward losses. The effects of cyclic deforming and hole storage are also absent.

Short-Circuit Bridge (Forward Test)

The d.c. terminals of the bridge are short-circuited and the a.c. terminals are energized from an a.c. supply until full rated bridge current flows, i.e. twice the diode rating. This current may be measured either as a.c. input current or d.c. output current using a low resistance ammeter, there being merits in both methods as will be seen.

The alternating voltage at the input terminals of the bridge is measured.

Three main alternatives are possible:

(a) Low impedance a.c. supply—r.m.s. or mean input voltage to rectifier measured.

(b) High impedance a.c. supply—mean input voltage to rectifier measured.

(c) High impedance a.c. supply—r.m.s. input voltage to rectifier measured.

In method (a) the rectifier input voltage is sinusoidal and it is immaterial whether the mean or r.m.s. value is measured. However, since the shape of the current wave will be determined solely by the curvature of the forward characteristic of the rectifier, there being no other impedance in the circuit, the operating conditions do not correspond to those obtaining when the rectifier is handling power in service, and the readings obtained for mean forward voltage drop tend to be very low.

By contrast, in methods (b) and (c) the current is forced to be sinusoidal. The mean input voltage measured in (b) can therefore be divided by four to give the standard voltage drop per diode but the r.m.s. input voltage given by (c) cannot be converted into standard terms and therefore cannot accurately predict the performance on load.

Of these three methods, therefore, (b) is recommended and a resistive ballast in the a.c. circuit of 50 times the mean bridge input voltage is suggested in order to ensure sinusoidal current drive.

Several errors, however, still result from this method. First, as in all such short-circuit tests, the rectifier is not subjected to reverse voltage and, particularly in the case of selenium rectifiers, the forward voltage drop tends to be low for this reason.

Secondly, the measurement of the mean input voltage in (b) may involve the use of a potential transformer feeding a rectifier instrument (as in normal multi-range test sets) and if the rectifier is seriously unbalanced there will be an error due to the loss of the d.c. component as discussed previously. If method (a) is preferred, the use of a d.c. ammeter to short-circuit the bridge introduces a variable degree of ballasting and as in all methods, this must be suitably corrected for voltage drop. If an a.c. ammeter is used with method (a) it will either cause excessive ballasting of a distorting nature or if it is fed from a current transformer

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it will be subject to error due to loss of the d.c. component.

Finally, for larger equipment in which three-phase rectification is employed, all such short-circuit methods will give low readings for mean forward drop and forward losses, for another fundamental reason. This is that under short-circuit test conditions the total circuit voltage falls to a very low value just sufficient to pass full forward current, but the reactance voltage drops in the supply transformer etc. remain at their full value. As a result the commutation between phases changes, the overlap angle increases, the effect of circuit reactance becomes dominant and the forward conduction angle tends to increase from 120° to 180°. Thus the mean forward voltage drop and the forward losses on short-circuit fall considerably below those which would occur in full power operation at the same load current.

LOADED BRIDGE (FORWARD TEST)

In order to overcome the disadvantages of the previous method the bridge may be operated at full voltage and current on resistance load. This produces the standard loading conditions and the mean dynamic forward drop in the bridge can be obtained by accurately measuring the input and output voltages. If the r.m.s. input and d.c. output voltages are V_{ac} and V_{dc} respectively, then

$$\text{Mean dynamic voltage drop per element} = \frac{(0{\cdot}9\ V_{ac}) - V_{dc}}{4}$$

The mean current per element is, of course, one half of the d.c. output current.

Although this method is free from the errors discussed above, it is greatly dependent upon the accuracy of the instruments since the result is obtained as the difference of two large quantities. This instrument error can be reduced considerably when testing bridges of greater than about 50V output, by using a transformerless rectifier voltmeter as a transfer instrument between the a.c. and d.c. terminals of the bridge. With this instrument, the total mean dynamic drop in the bridge is given by 0·9 × (difference between a.c. and d.c. readings) and the mean dynamic voltage drop per element is one quarter of this.

Approximate Dynamic Methods Requiring Two Diodes

The following methods are logical derivatives of the bridge methods and therefore exactly the same comments regarding sources of error are applicable.

OPEN-CIRCUIT BACK TO BACK PAIR (REVERSE TEST)

Two diodes are connected back to back and the rated a.c. voltage is applied. The mean reverse current is measured by a rectifier milliammeter. If a transformer coupled milliammeter must be used, a check should be made for absence of d.c. component. The standard mean reverse current per diode is obtained by dividing the total mean reverse current by two.

PARALLEL OPPOSITION PAIR (FORWARD TEST)

Two diodes are connected in parallel, but with reversed polarity and are connected in series with an a.c. supply and a rectifier a.c. ammeter. The applied voltage is increased until the mean value of the current is equal to twice the half-wave current rating of each diode (i.e. equal to the bridge rating). The a.c. voltage appearing across the two diodes is measured.

Three alternatives are possible:

(a) Low impedance a.c. supply—r.m.s. or mean rectifier voltage measured.
(b) High impedance a.c. supply—mean rectifier voltage measured.
(c) High impedance a.c. supply—r.m.s. rectifier voltage measured.

Method (b) is preferred for the reasons discussed previously. The standard dynamic forward characteristic (which, of course, appears as the average of the two diodes) is obtained by dividing both the total current and the mean rectifier voltage by two.

Approximate Dynamic Methods Requiring Only One Diode

The main advantage of these methods is that they provide readings relating only to the particular diode tested; no averaging with other diodes being involved. They are particularly useful as they may be applied independently to individual diodes or arms of a complete rectifier arrangement.

FULL-WAVE RECTIFIED SINE WAVE (FORWARD TEST)

Forward current derived from a ballasted full-wave single-phase unsmoothed rectifier is passed through the diode. The mean value of this current and the mean d.c. voltage across the diode are measured by moving-coil instruments. To convert to standard values, both the mean current and the mean voltage are divided by two.

The following errors are introduced:

(a) If double the rated mean forward current is passed, but, of course, without idle or reverse half cycles, the excursion over the characteristic will be the same as in normal operation, i.e. the peak current will be correct, but the junction temperature will be excessive unless the test is made extremely brief (e.g. 1 second). The reading obtained will tend to be low.

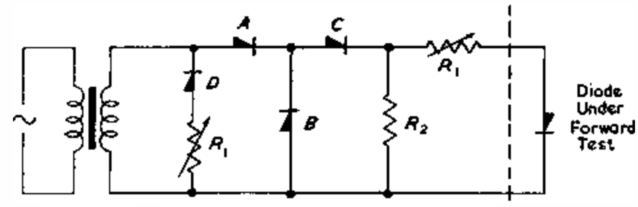


Fig. 10. Circuit for generating half sine-wave forward current

(b) If normal rated mean forward current is chosen, the excursion over the characteristic will be only half normal, the peak current will be half normal, and the dynamic characteristic can only be plotted to half rating.
(c) There will be a further tendency for the reading to be low due to the absence of reverse voltage stress (particularly in selenium rectifiers).

FULL-WAVE RECTIFIED SINE WAVE (REVERSE TEST)

Reverse voltage derived from an unsmoothed single-phase full wave rectifier is applied to the diode. The mean d.c. input voltage is made equal to 0·9 times the r.m.s. diode rating. The mean reverse current is measured. To convert to standard values, input voltages are divided by 0·9 and reverse current readings are divided by two.

The following conflicting errors are introduced:

First, the reading of reverse current will tend to be low due to the absence of forward current on alternate half cycles, and secondly the reading will tend to be high owing to the absence of any elements which must carry the peak reverse current of the test diode in the forward direction.

HALF SINE WAVE (FORWARD TEST)

Forward current of a special waveform comprising half sine waves separated by half cycles of zero current, having a mean value equal to the diode rating, is passed. The mean value of the voltage drop across the diode is measured. The mean instruments read directly in standard values. The only error introduced is that particularly with selenium rectifiers the forward drop measured will be low due to the absence of reverse voltage.

Half-Sine Wave (Reverse Test)

Reverse voltage having the same special waveform as above is applied. The mean d.c. value of the applied voltage is measured and is made equal to 0·45 times the r.m.s. diode rating. The mean reverse current is measured. To convert to standard values, input voltages are divided by 0·45, but the mean reverse current is read directly. The main source of error is that owing to the absence of forward current the reverse current measured may be low.

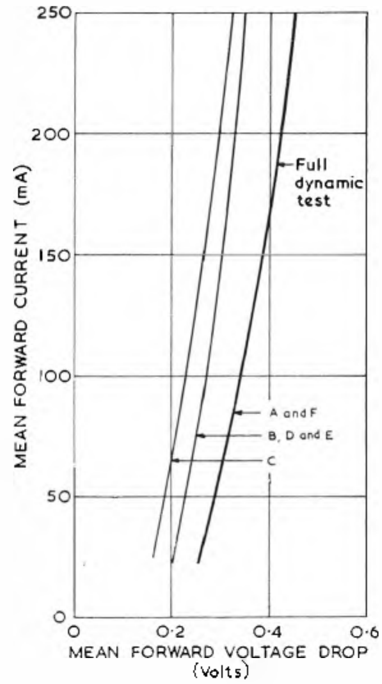


Fig. 11. Selenium dynamic forward characteristics

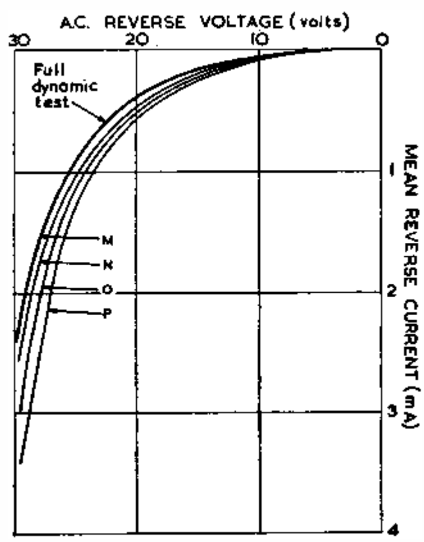


Fig. 12. Selenium dynamic reverse characteristics

Power Source for Half Sine Wave Tests

The circuit shown in Fig. 10 is suitable for generating half sine wave forward current.

Resistor R_1 is chosen to drop about 50 times the mean forward drop in the test diode. Diodes *A*, *B* and *D* are rated to withstand the full a.c. voltage, but diode *C* can be of low voltage rating. For current rating diodes, *A*, *C* and *D* must carry the full load required. The shunt resistor R_2 is chosen to reduce the unwanted reverse half cycle to negligible proportions.

A similar circuit using low current high voltage rectifiers is used for generating the half sine wave reverse waveform.

For both forward and reverse waveforms it is essential to check by oscillograph or unipolar voltmeter that the unwanted half cycle is effectively suppressed.

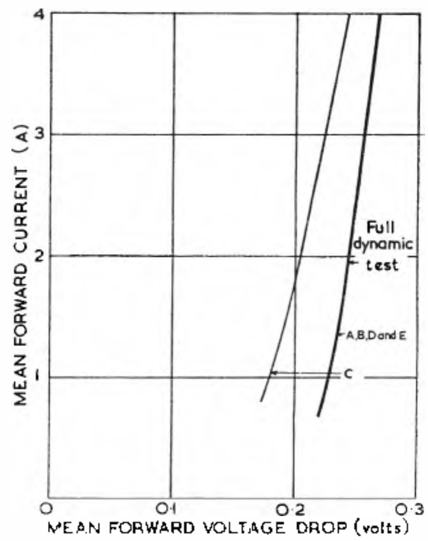


Fig. 13. Germanium dynamic forward characteristics

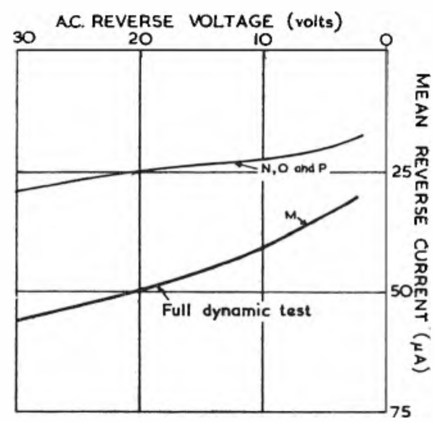


Fig. 14. Germanium dynamic reverse characteristics

Comparison of dynamic characteristics measured by various methods for typical selenium and germanium rectifiers

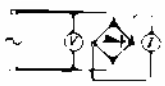
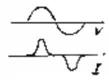
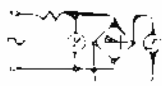
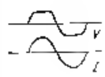
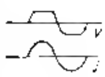
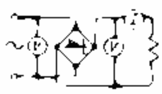
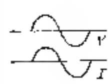
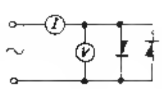
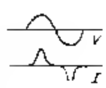
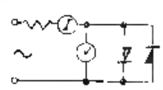
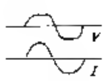
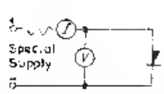
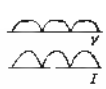
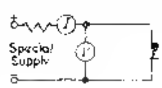
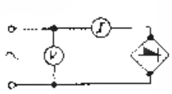
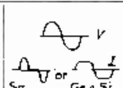
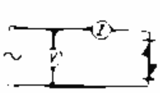
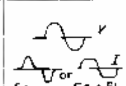
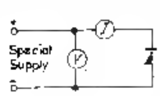
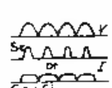
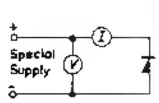
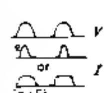
FORWARD CHARACTERISTICS:
A = Full dynamic test
B = Short circuit bridge test. High impedance supply. R.M.S. voltage measured
C = Short circuit bridge test. High impedance supply. Mean voltage measured
D = Short circuit bridge test. Low impedance supply
E = Full wave rectified sinewave test
F = Half sinewave test
G = Loaded bridge test

REVERSE CHARACTERISTICS:
M = Full dynamic test
N = Open circuit bridge
O = Full wave rectified sinewave
P = Half sinewave

Comparison of Approximate Dynamic Methods With Full Dynamic Methods

Some indication of the degree of error which can be introduced by the above approximate methods is given in the curves of Figs. 11, 12, 13 and 14, in which the dynamic forward and reverse characteristics of typical germanium

APPROXIMATE DYNAMIC TEST METHODS FOR SEMICONDUCTOR RECTIFIERS

	METHOD	NO. OF DIODES REQD.	CIRCUIT	INPUT WAVEFORMS	MEASUREMENTS	CONVERSION FACTORS TO STANDARD DYN. CHARS PER DIODE	INHERENT SOURCES OF ERROR	COMMENTS
FORWARD TESTS	Short-circuited Bridge Low Impedance Supply	4			Mean or r.m.s. input voltage Mean output current	Divide mean input voltage by 4 Divide output current by 2	No reverse stress is applied Current waveform is distorted	Voltage drop in ammeter must be allowed for. An a.c. rectifier ammeter can be used subject to precautions discussed in text.
	Short-circuited Bridge High Impedance Supply	4			Mean input voltage Mean output current	Divide mean input voltage by 4 Divide output current by 2	No reverse stress is applied	Voltage drop in ammeter must be allowed for. An a.c. rectifier ammeter can be used without error as an alternative.
	Short-circuited Bridge High Impedance Supply	4			R.M.S. input voltage Mean output current	Cannot be interpreted	No reverse stress is applied	Readings cannot be corrected to standard dynamic characteristics and therefore cannot accurately be related to performance on load or to rectifier losses (see text).
	Loaded Bridge	4			R.M.S. input voltage Mean output voltage Mean output current	Subtract mean output voltage from 0·9 × input voltage. Divide remainder by 4. Divide output current by 2	None	Risk of inaccuracy as result is obtained as difference of two larger quantities. Errors can be reduced by use of transfer instrument (see text).
	Parallel Opposed Pair Low Impedance Supply	2			Mean or r.m.s. input voltage Mean input current	Divide mean input voltage by 2 Divide mean input current by 2	No reverse stress is applied Current waveform is distorted	A.C. ammeter gives some current ballasting and resultant distortion of input voltage. Risk of error in a.c. ammeter and a.c. voltmeter due to d.c. component resulting from unbalanced diodes.
	Parallel Opposed Pair High Impedance Supply	2			Mean input voltage Mean or r.m.s. input current	Divide mean input voltage by 2 Divide mean input current by 2	No reverse stress is applied	No error in rectifier ammeter but risk of d.c. component error in input voltmeter if transformer coupled (see text).
	Full Wave Rectified Forward Current	1			Mean input voltage Mean input current	Divide input voltage by 2 Divide input current by 2	Mean current density is twice normal for a given peak value No reverse stress is applied	Auxiliary full-wave rectifier required to generate input waveform. Characteristics can only be taken up to half normal rating without overheating.
	Half Sine Wave Forward Current	1			Mean input voltage Mean input current	Reads directly in standard values	No reverse stress is applied	Special rectifier arrangement required to generate input waveform. Risk of error from imperfectly generated waveform. As opposed to above method the characteristics can be taken up to full rating.
REVERSE TESTS	Open-circuited Bridge	4			R.M.S. input voltage Mean input current	Divide mean input current by 4	No forward current flowing	If transformer coupled rectifier milliammeter is used, there is risk of error due to d.c. component if bridge is unbalanced (see text).
	Back to Back Pair	2			R.M.S. input voltage Mean input current	Divide mean input current by 2	No forward current flowing	If transformer coupled rectifier milliammeter is used, there is risk of error due to d.c. component if diodes are unbalanced (see text).
	Full Wave Rectified Reverse Voltage	1			Mean input voltage Mean input current	Multiply mean input voltage by 1.11 Divide mean input current by 2	No forward current flowing. Recovery period between reverse half cycles is half normal.	Auxiliary full-wave rectifier required to generate input waveform; preferably loaded. Reverse loss double normal but this not serious as owing to absence of forward heating, junction is cooler than normal. Errors depend greatly on type of rectifier.
	Half Sine Wave Reverse Voltage	1			Mean input voltage Mean input current	Multiply mean input voltage by 2·22 Input current reads standard value directly	No forward current flowing	Special rectifier arrangement required to generate input waveform. Risk of error with imperfectly generated waveform. Reverse losses normal apart from effect of absence of forward current.

and selenium rectifiers measured by many methods, including full dynamic test, are shown.

It can be seen that by the use of approximate methods, errors of as much as 30 per cent can occur in the forward direction, while in the reverse direction even larger errors can arise particularly with germanium rectifiers.

It has also been found that rectifiers of the same general type, supplied by different manufacturers, show different errors in the various approximate dynamic test circuits and that they may, therefore, appear in a different order of merit depending on the circuit used. A summary of approximate dynamic methods is given in Table 1.

Conclusion

Approximate dynamic test methods must be used with caution and a knowledge of their weaknesses. In general, they are only suitable for routine testing under fixed conditions when allowance can be made for the errors introduced by the test methods chosen, when used with the particular type of rectifier concerned.

However, if a standard dynamic testing technique is required which will give accurate and reproducible results not only on different types of rectifiers but even on different makes of the same type, whether as a standard type test or as a check on simpler methods, a full dynamic test is essential. For this purpose it has been found that the simple half wave resistance loaded circuit which can be easily and reliably reproduced, used in conjunction with the special instruments described herein, will meet the strict requirement of such a standard test and is applicable to semiconductor rectifiers of all types.

Acknowledgment

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A Feedback Modulator

By J. L. Douce*, Ph.D., and M. Flinders*, B.Sc.

A reliable accurate unit is presented which modulates the mark-to-space ratio of a 10kc/s pulse train in accordance with an input voltage. Feedback is employed to obtain high accuracy, together with good long-term stability. Some applications are briefly discussed.

In the recording and processing of analogue information, it is often convenient to encode the transmitted data in order to minimize the effect of noise or varying gain in the system. For example, when recording signals on magnetic tape, some modulating mechanism must be employed if very low frequencies are being handled. Amplitude modulation of a high frequency carrier signal may be employed, but this technique suffers the disadvantage that variations in amplifier gain cause the output amplitude to vary. Where the data is stored on a magnetic drum[1], drum eccentricity causes the output signal to vary in amplitude as the drum rotates. These disadvantages are eliminated by modulating a carrier signal in such a manner that the amplitude of the carrier is unimportant. Two such methods may be considered, frequency modulation and pulse width modulation. In frequency modulation, the frequency of a carrier signal is varied in sympathy with the data to be encoded, giving a sinusoidal signal of varying frequency. Evidently if the data is recorded on tape or film, the play-back speed must be accurately equal to the recording speed to eliminate errors. Pulse width modulation does not suffer from this

* *University of Manchester.*

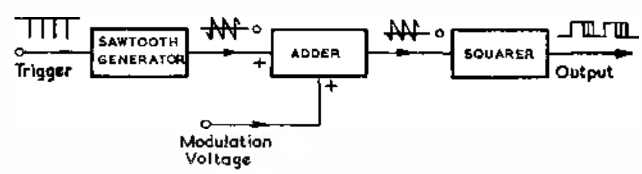


Fig. 1. Schematic of single modulator

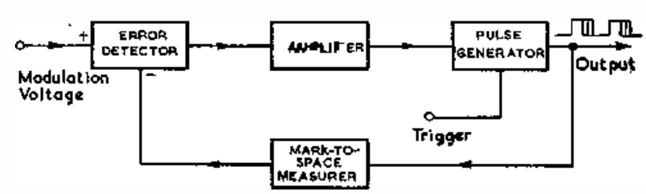
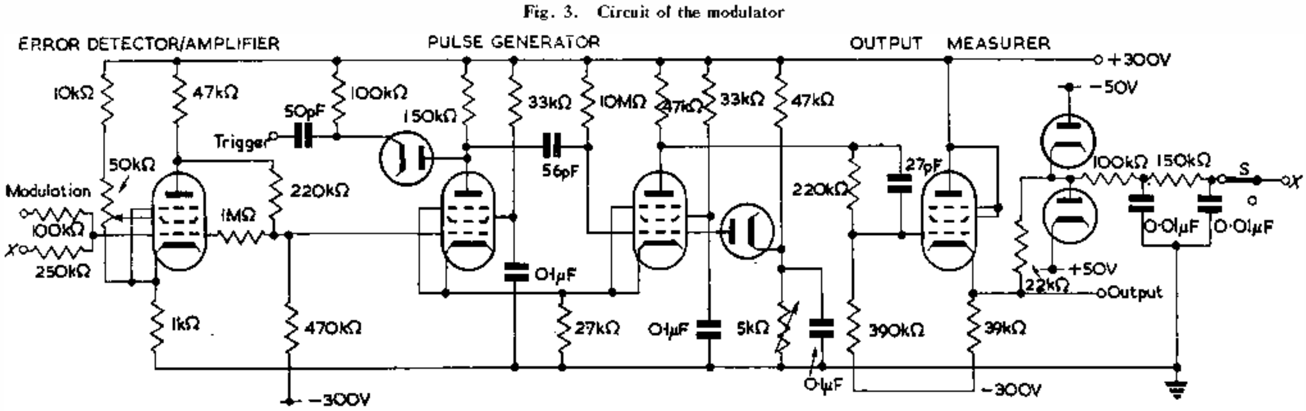


Fig. 2. Schematic of the feedback modulator

Fig. 3. Circuit of the modulator



disadvantage, which becomes an important feature when accurate speed control is difficult.

A further advantage of pulse-width modulation is that signals encoded in this form may be fed directly into an electronic multiplier, giving a simplification in an analogue computer[2].

The modulator described has been incorporated in an analogue computer which includes a magnetic drum as a storage element, information being stored on the drum as a series of 10kc/s pulses, of variable mark/space ratio. Techniques have been developed to enable this storage element, in conjunction with an electronic multiplier, to simulate and analyse the behaviour of complex feedback systems.

The features required in a modulator for use in this equipment are reliability, good zero stability, linearity and compensation against variation in pulse repetition frequency.

It is shown how the feedback circuit developed satisfies these requirements in a simple manner. Measurements verify the satisfactory behaviour of the unit.

Principle of Operation

The simplest form of pulse width modulator is shown schematically in Fig. 1.

Pulses of the modulation frequency trigger a sawtooth generator. This sawtooth waveform is arranged to go equally positive and negative, as shown. The modulation voltage is then added to this sawtooth, producing a sawtooth voltage whose mean value corresponds to the modulating voltage. The instant at which this resultant voltage passes through zero varies linearly with the modulating voltage. Hence, by squaring this waveform, pulses are obtained such that the mark-to-space ratio varies linearly with the modulating voltage.

This equipment is satisfactory provided that only moderate accuracy is required. Evidently drift in the squaring amplifier produces a corresponding change in the mark-to-space ratio of the output. A more serious disadvantage is that variation in the pulse repetition frequency gives a variation in the modulated pulses. For example, the sawtooth generator may be a suppressor switched Miller integrator[3]. In this case the circuit will integrate a constant voltage for a time equal to the period of the switching waveform. Thus any change in the switching period gives a corresponding change in the final value of the output voltage.

This simple circuit shows appreciable non-linear behaviour when large variation in the mark-to-space ratio is desired (e.g. variation of 100:1).

These disadvantages are removed if feedback is employed. The principle is to derive a voltage from the output waveform accurately proportional to the mark-to-space ratio, and to compare this signal with the modulating voltage. Any difference between these two voltages is used to change the mark-to-space ratio of the output to its true value. The schematic diagram is shown in Fig. 2.

The mark-to-space ratio of the output is found simply by averaging the output waveform, using a simple *CR* network. The time-constant *CR* is chosen to be much greater than the period of the pulses, but much less than the shortest period present in the modulating signal.

Practical Details

The circuit employed is shown in Fig. 3. This includes a modulator consisting of a triggered multivibrator, chosen to give a large variation in mark-to-space ratio[4]. This is followed by a smoothing unit measuring the mark-to-space ratio of the output and a feedback network in which this signal is compared with the modulating signal. The error is used to vary a control grid voltage in the multivibrator, controlling the resultant mark-to-space ratio. For comparing the behaviour of the unit with and without feedback, the feedback may be removed by opening the switch *S*.

The effect of feedback has been determined experimentally by measuring the mark-to-space ratio of the output waveform as the input is varied, with and without feedback. The linearity of the modulator is determined by applying a steady voltage as the modulating signal, and measuring the output mark-to-space ratio. Without feedback, errors up to 5 per cent can occur, as the mark-to-space ratio varies by 100:1. The error is less than 1 per cent with feedback.

In the application considered for this modulator, pulse-repetition frequency variation must have little effect on the output signal. This requirement has been satisfied, and checked experimentally as shown in Fig. 4. Without feedback, a 10 per cent change in frequency can give a 30 per

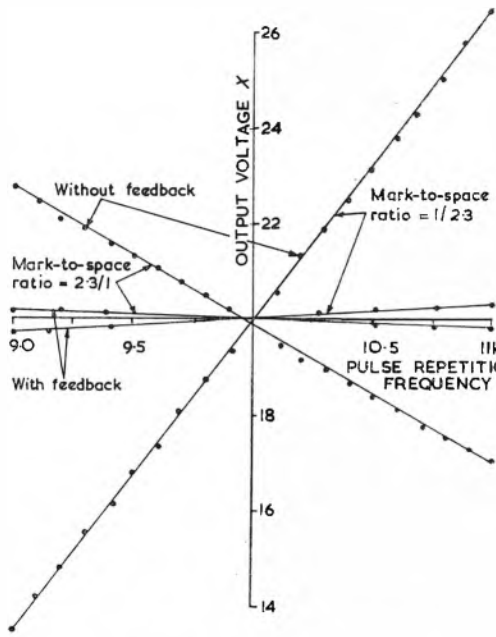


Fig. 4. Effect of repetition frequency variation

cent change in output. With feedback, the percentage change is reduced by a factor of 20.

Conclusions

A simple pulse-width modulator has been described which incorporates feedback to improve the response. Of special interest is the high accuracy of the device for wide variations in the output pulse width, together with freedom from errors due to variation in pulse repetition frequency.

Experimental results verify the advantages of the feedback modulator described.

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Circuit Analysis by Normalization

By J. J. Hupert*, Dipl.Ing., Ph.D., M.I.E.E.

The concept of frequency as a complex number opens a number of interesting possibilities in application of lucid mathematical techniques to the examination of network functions.

This article draws attention to a method of network analysis based on a procedure of normalization similar to that used in the concept of the 'normalized resonance curve.'

ONE of the most commonly known applications of the normalization of frequency variables is the familiar concept of the normalized resonance curve of a resonant circuit[1].

The contention of this article is that similar normalization of frequency variables in the complex frequency plane representing positions of singular points of a transfer function of a network proves a powerful means of network analysis and synthesis.

It has previously been demonstrated that consideration of normalized contributions of critical frequencies to network functions can be of considerable help in the evaluation of non-linear distortion arising in f.m. transmission[2], but it is felt that the popularization of this approach will result in its application to the analysis also of other circuit properties by engineers familiar with lucid mathematical methods made available by the application of the theory of the function of a complex variable to network analysis[3,4,5,6].

In particular, it should be emphasized that the normalized *amplitude* transfer function of a single tuned circuit should be considered as accompanied by a corresponding normalized *phase* transfer curve and the entire family of derivatives of these two curves.

It will be further demonstrated that the transfer amplitude function, transfer phase function, and all their derivatives for any lumped constant network can be represented as a sum of a finite number of appropriate normalized curves drawn each to its own suitably adjusted scale.

In particular the advantages of the method of derivative adjustment were recently pointed out by Lynch,[7] but not in a form recognizing individual contributions of zeros and poles, which approach adds additional insight to the quantitative aspects of the circuit examined. The procedure further described in this article can be followed analytically or graphically. In many practical cases, clarity of the graphical procedure presented is sufficient to find desirable positions of zeros and poles by fast trial-and-error methods and to synthesize networks for prescribed performance when numerical tolerances are given.

Attention is drawn to the fact that the method of adjustment of derivatives of modulus and phase curves leads numerically to more accurate results than adjustments observed only on the curves themselves.

Denoting a network function (transfer or driving-point 'immittance' or ratio) as

$$T(s) = k\frac{(s - s_{z1})(s - s_{z2})\ldots\text{etc.}}{(s - s_{p1})(s - s_{p2})\ldots\text{etc.}} \quad \ldots\ldots\ldots (1)$$

where complex numbers s_{z1}, s_{z2} . . . etc. denote the positions of zeros and s_{p1}, s_{p2} . . . etc. the positions of poles of the function studied[3,4,5,6] s is the independent complex variable $j\omega$ (ω also assumed complex). (Fig. 1).

Every one of the complex numbers s_{z1}, s_{z2} etc., s_{p1}, s_{p2} etc. can be represented as:

$$s_{zk} = b_{rzk} + jb_{izk} \text{ and } s_{pk} = b_{rpk} + jb_{ipk} \quad \ldots\ldots\ldots (2)$$

where b_{rzk}, b_{izk}, b_{rpk}, b_{ipk} are real numbers with the following meaning of subscripts:

r —real part

i—imaginary part

z—of a zero

p—of a pole

k —'running' subscript assuming numbers 1, 2, 3, 4,

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List of Symbols

$T(s)$ = circuit transfer function

For real frequencies $s = j\omega$ or $s = j(\omega/2\pi)$, as applicable. In general

s	—	complex variable in $(\alpha + j\omega)$ or $(\alpha/2\pi + j\omega/2\pi)$ plane
arg	=	angle of a phasor denoting a complex number
k	=	real constant
s_{zk}	=	position of a zero in a complex plane; index z denotes zero; index k running numerical index to be substituted by the applicable number in a case of a specific zero
s_{pk}	=	position of a pole in a complex plane; meaning of indices as above
s_{pk}'	=	conjugate of s_{pk}
ω_0	=	resonant angular frequency of a tuned circuit (or pass-band centre)
$\Delta\omega$	=	increment of frequency variation for a narrow-band case $= \omega - \omega_0$
Q	=	magnification factor of a tuned circuit
g_1, g_2, g_3	=	valve transconductance symbols
x	=	frequency variable, normalized to a position of a pole or zero. Origin at the frequency of this pole or zero
x_k	=	as above-frequency variable normalized to pole or zero (k). Origin at the frequency of pole or zero (k).
b_{rzk}, b_{rpk}, b_{izk}, b_{ipk}		Co-ordinates of poles or zeros in complex plane. Dimensions: frequency or angular frequency. Indices denote: first: real or imaginary component; second: pole or zero; third: running subscript
B_k	=	bandwidth of pole (k) or zero (k).
$\Gamma(s)$	=	natural logarithm of $\lvert T(s)\rvert$ as a function of s
$\Phi(s)$	=	argument of $T(s)$, in radians
Z	=	total number of zeros of a transfer function
P	=	total number of poles of a transfer function
n	=	order of derivative
$A(x)$	=	specific function of x
$\phi(x)$	=	designation of arc tan x. Primes denote first four derivatives of the above two functions

etc. according to the number allocated to the pole or zero in question

It will be noticed that the dimension of terms b above is that of an angular frequency. Therefore, b stands for bandwidth.

Without much difficulty equation (1) above can be rearranged for s to mean $s = j(\omega/2\pi)$ and dimension of b to become that of frequency.

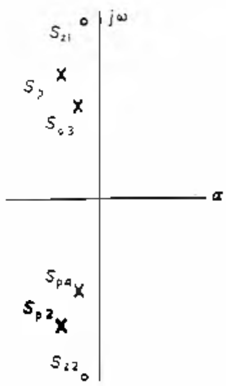

Fig. 1. Pole-zero location in left-half plane for a transfer function

Now, normalizing the frequency variable ω to the bandwidth B_k of a selected pole or zero, for example, to the bandwidth B_{pk} of the pole k, one defines the concept of the 'bandwidth of a pole' as $B_{pk} = 2|b_{rpk}|$.

It will be noticed that B_{pk} would represent the 3dB bandwidth calculated from the universal resonance curve of a tuned circuit derived from the data of pole (k). We can imagine this tuned circuit to be represented by two poles s_{pk} and s_{pk}' and a zero at the origin s_{z1} but with the contributions of s_{pk}' and s_{z1} neglected because of the approximations $|j\omega - s_{pk}'| = 2j\omega_0$ and $\omega_0 = \omega_0 + \Delta\omega$ (or $\Delta\omega \ll \omega_0$).

The above approximations are known as '*narrow-band approximations*'. By defining the concept of 'bandwidth of a pole' one is gaining the advantage of being able to use the concept of the normalized resonance curve even in cases where narrow-band approximation does not apply. In such cases one simply has to add the contributions of all poles, including those which would be neglected in narrow-band approximation.

In order to work out individual contributions of every zero and pole to the amplitude and phase transfer, as well as to their derivatives, equations (2) are substituted for the values of s_{zk} and s_{pk} in equation (1), giving

$$T(s) = k\ \frac{[s - (b_{rz1} + jb_{iz1})]\,[s - (b_{rz2} + jb_{rz2})]\ \ldots \text{etc.}}{[s - (b_{rp1} + jb_{ip1})]} \quad \ldots\ldots\ldots\ldots (3)$$

It is known from the criterion of stability for networks that in the case of a stable network all values of b_{rpk} are negative (poles lie in the negative half-plane). Thus

$$b_{rpk} = -|b_{rpk}|$$

In analogy to the case of a single resonant circuit, the half-bandwidth $B_{pk}/2 = |b_{rpk}|$ of the pole (k) is taken as a reference to which the frequency variable will be normalized. After the above preliminary steps, convenient generalized equations are sought for transfer modulus and phase as well as their respective derivatives as functions of frequency. These equations should be expressed in terms of summation of individual contributions due to all poles and zeros of the complex transfer function.

Instead of modulus function, its logarithm $\Gamma(s) = \ln|T(s)|$ is derived for the sake of the convenience of having additive and not multiplicative contributions of each pole and zero in the result. This step is further justified by the well-known practical convenience of expressing transfer gain or loss of networks in the scale of nepers or decibels.

The desired network functions are

$$\Gamma(s) = \ln|T(s)| = \sum_{k=1}^{k=Z} \ln|s - s_{zk}| - \sum_{k=1}^{k=P} \ln|s - s_{pk}| + \ln k \quad \ldots\ldots\ldots\ldots (4)$$

$$\frac{d^n\Gamma(s)}{d\omega^n} = \sum_{k=1}^{k=Z} (d^n/d\omega^n) \ln|s - s_{zk}| - \sum_{k=1}^{k=P} (d^n/d\omega^n) \ln|s - s_{pk}| \quad \ldots\ldots\ldots\ldots (5)$$

$$\Phi(s) = \arg(T(s)) = \sum_{k=1}^{k=Z} \arg(s - s_{zk}) - \sum_{k=1}^{k=P} \arg(s - s_{pk}) \quad \ldots\ldots\ldots\ldots (6)$$

$$\frac{d^n\Phi(s)}{d\omega^n} = \sum_{k=1}^{k=Z} (d^n/d\omega^n) \arg(s - s_{zk}) - \sum_{k=1}^{k=P} (d^n/d\omega^n) \arg(s - s_{pk}) \quad \ldots\ldots\ldots\ldots (7)$$

Equations (4) to (7) express the contributions due to individual poles and zeros in terms of the original variable. In order to effect the change of variable the contribution of each singular point is expressed in terms of the frequency variable normalized in terms of its own half-bandwidth $B_k/2$.

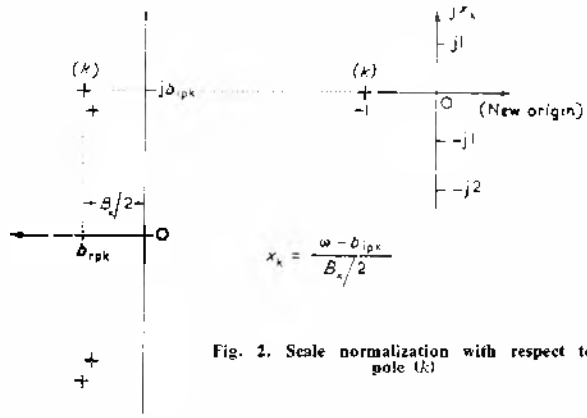

Fig. 2. Scale normalization with respect to pole (k)

The co-ordinates of the complex plane are modified differently for each considered zero or pole (k) in such a manner that the abscissa of (k) is always -1 and its ordinate is zero. Points jx_k are then considered along an imaginary axis, the ordinates of which are in the new system

$$x_k = \frac{\omega - b_{ipk}}{B_k/2} \quad \ldots\ldots\ldots\ldots (8)$$

Equation (8) shows the relationship between original frequency variable ω or $\omega/2\pi$ and ordinates x_k, normalized to the data of pole (k). (Fig. 2).

Accordingly

$$\arg(s - s_{pk}) = \arc\tan x_k \quad \ldots\ldots\ldots (9)$$

$$\ln|s - s_{pk}| = \ln[(B_k/2)(1 + x_k^2)^{\frac{1}{2}}] = \ln(B_k/2) + \ln(1 + x_k^2)^{\frac{1}{2}} \quad \ldots (10)$$

Substituting equation (8) into equations (4), (5), (6) and (7),

$$\frac{d^n\Gamma(s)}{d\omega^n} = \sum_{k=1}^{k=Z} (2/B_k)^n\ \frac{d^n \ln(x_k^2 + 1)^{\frac{1}{2}}}{dx_k^n} - \sum_{k=1}^{k=P} (2/B_k)^n\ \frac{d^n \ln(x_k^2 + 1)^{\frac{1}{2}}}{dx_k^n} \quad (11)$$

and

$$\frac{d^n\phi(s)}{d\omega^n} = \sum_{k=1}^{k=Z}(2/B_k)^n \frac{d^n \arctan x_k}{dx_k^n} - \sum_{k=1}^{k=P}(2/B_k)^n \frac{d^n \arctan x_k}{dx_k^n} \quad (12)$$

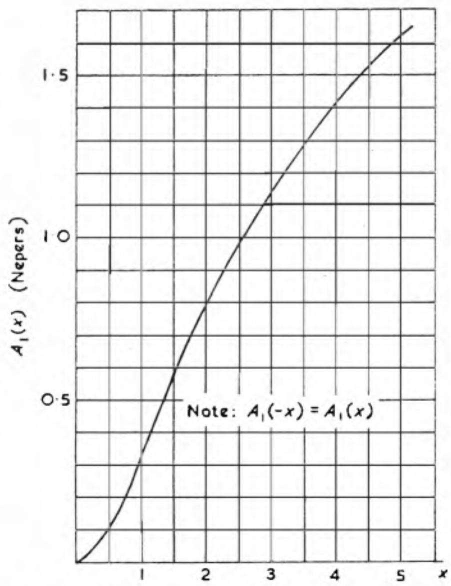

Fig. 3. Normalized contribution of a simple pole to the gain-frequency function

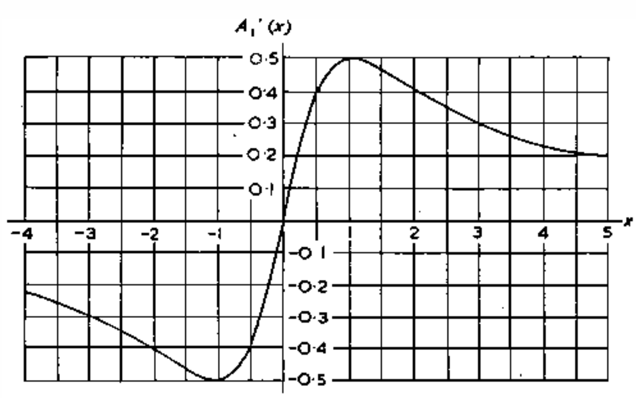

Fig. 4. Normalized contribution of a simple pole to the first derivative of the gain-frequency function

Also

$$\Gamma(s) = \ln k + \sum_{k=1}^{k=Z}\ln(B_k/2) + \sum_{k=1}^{k=Z}\ln(x_k^2+1)^{\frac{1}{2}} - \sum_{k=1}^{k=P}\ln(B_k/2) - \sum_{k=1}^{k=P}\ln(x_k^2+1)^{\frac{1}{2}} \quad \ldots\ldots \quad (13)$$

and

$$\Phi(s) = \sum_{k=1}^{k=Z}\arctan x_k - \sum_{k=1}^{k=P}\arctan x_k \quad \ldots\ldots\ldots \quad (14)$$

From equations (11) to (14) the modulus of the transfer function (in logarithmic scale), its phase and derivatives of desired order of both those functions have been expressed in terms of individual contributions of each pole of zero. The absolute value of each contribution takes the form of modulus, phase, and their derivatives of the universal resonance curve[3] of each pole and each zero $1/(jx_k + 1)$, but normalized to its own independent variable x_k and with the scale of ordinates changed by the factor $(2/B_k)^n$. In the functions $\Gamma(s)$ and $\phi(s)$ no ordinate scale change is involved. Scales of variables x_k *are different for each zero and pole*, the unit of x_k being $B_k/2$. As discussed earlier, contributions of all zeros and poles stem from one of the functions belonging to one of the following *two* families of functions:

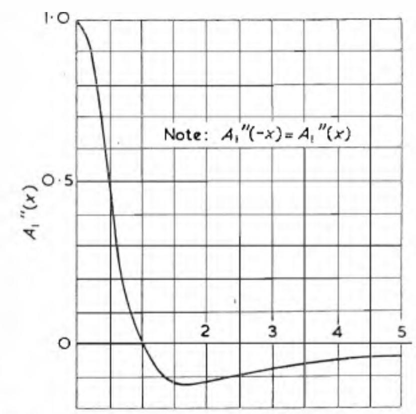

Fig. 5. Normalized contribution of a simple pole to the second derivative of gain-frequency function

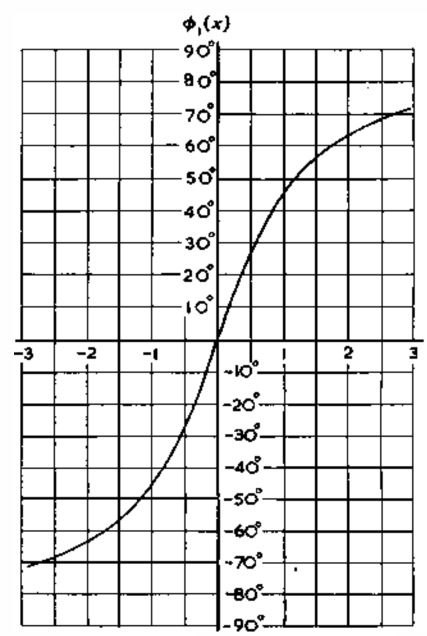

Fig. 6. Normalized contribution of a simple pole to the phase-frequency function

(a) *Modulus and its derivatives:*

$$A(x) = \ln(1 + x^2)^{\frac{1}{2}} \quad \ldots\ldots\ldots\ldots \quad (15)$$

(Fig. 3)

$$A'(x) = (dA/dx) = x(1 + x^2)^{-1} \quad \ldots\ldots \quad (16)$$

(Fig. 4)

$$A''(x) = (d^2A/dx^2) = (1+x^2)^{-1} - 2x^2(1+x^2)^{-2} \quad \ldots \quad (17)$$

(Fig. 5)

$$A'''(x) = (d^3A/dx^3) = -6x(1+x^2)^{-2} + 8x^3(1+x^2)^{-3} \quad \ldots\ldots\ldots \quad (18)$$

$$A''''(x) = (d^4A/dx^4) = -6(1+x^2)^{-2} - 48x^2(1+x^2)^{-3} - 48x^4(1+x^2)^{-4} \quad \ldots\ldots\ldots \quad (19)$$

(b) *Phase and its derivatives:*

$$\phi(x) = \arctan x \quad \ldots\ldots\ldots\ldots \quad (20)$$

(Fig. 6)

$$\phi'(x) = (d\phi/dx) = (1 + x^2)^{-1} \quad \ldots\ldots\ldots \quad (21)$$

(Fig. 7)

$$\phi''(x) = (d^2\phi/dx^2) = -2x(1 + x^2)^{-2} \quad \ldots\ldots\ldots\ldots \quad (22)$$

(Fig. 8)

$$\phi'''(x) = (d^3\phi/dx^3) = 8x^2(1 + x^2)^{-3} - 2(1 + x^2)^{-2}. \quad (23)$$

$$\phi''''(x) = (d^4\phi/dx^4) = -48x^3(1 + x^2)^{-4} + 24x\,(1 + x^2)^{-3} \quad \ldots\ldots\ldots\ldots\ldots \quad (24)$$

By means of the above data, transfer curves and their derivatives can be plotted for any lumped-circuit linear network, emphasis being placed on the usefulness[2,7] of the derivatives of $\Gamma(s)$ and $\Phi(s)$. For example, $\Gamma'(s)$ provides information about the rate of attenuation change outside the pass-band, $\Phi'(s)$ is the delay function, while higher order derivatives[2,9] of the phase transfer are useful in evaluation of f.m. distortion. Analytic equations (15) to (24) have their graphical interpretation which, in addition to providing a useful analysis method, may serve as a basis for some suggestions for placement of the zeros and poles for the purposes of network synthesis. Operation on derivative curves instead of $\Gamma(s)$ or $\Phi(s)$ is in many cases preferable since it enhances the accuracy of synthesis.

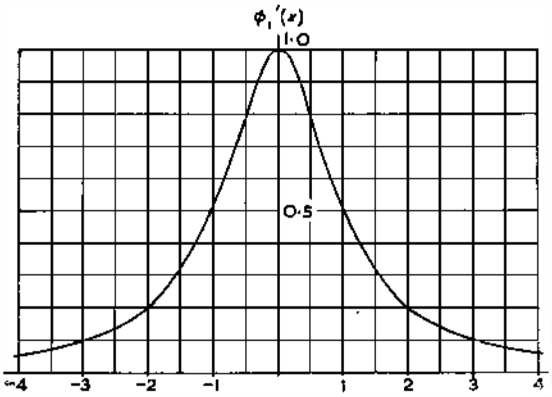

Fig. 7. Normalized contribution of a simple pole to the delay-frequency function

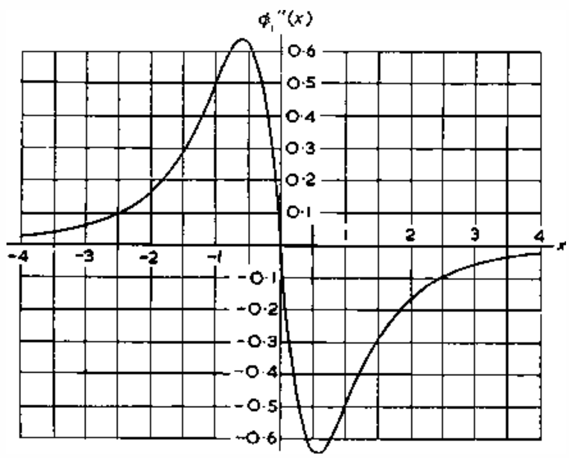

Fig. 8. Normalized contribution of a simple pole to the second derivative of the phase-frequency function

From equations (11) to (14) the following set of rules is derived for graphical procedure of computation, assuming that a set of poles and zeros is given:

(a) The values of $(B_k/2)$ is read for each zero and pole.

(b) The scales of x_k ($B_k/2$ is the length of one unit) are drawn in. The scale is different for each zero and pole, the origin lying at jb_{ipk} or jb_{izk} respectively.

(c) Individual contributions (as functions) of each zero and pole are drawn in, derived from Figs. 3 to 8 or calculated from equations (15) to (24), as applicable, by modifying the scale of ordinates by the factor $(2/B_k)^n$ for applicable n and reversing the sign of contributions in the case of poles.

(d) The sum of the individual contributions represents the desired characteristic curve of the network.

Example

As an example, the delay-versus-frequency characteristic is computed for a maximally-flat, staggered triple. The respective pole positions can be established from data available in the literature[10].

Considering three poles in 'narrow-band' representation, the following numerical data is assumed:

Band centre: $\omega_0/4\pi = 4\text{Mc/s}$

Q-factors of the three circuits:

$$Q_1 = 40 \;\therefore\; Q_2 = 20 \;\therefore\; Q_3 = 40$$

Corresponding values of $B_{pk}/2$ are equal to:

$$\frac{B_{p1}/2}{2\pi} = 50\text{kc/s} \;\therefore\; \frac{B_{p2}/2}{2\pi} = 100\text{kc/s} \;\therefore\; \frac{B_{p3}/2}{2\pi} = 50\text{kc/s}$$

The transfer function of the circuit shown in Fig. 9 is

$$T(s) = (V_o/E_1) = \frac{-g_1g_2g_3}{C_1C_2C_3}\cdot\frac{s^3}{(s-s_{p1})(s-s_{p2})(s-s_{p3})(s-s_{p1}')(s-s_{p2}')(s-s_{p3}')}$$

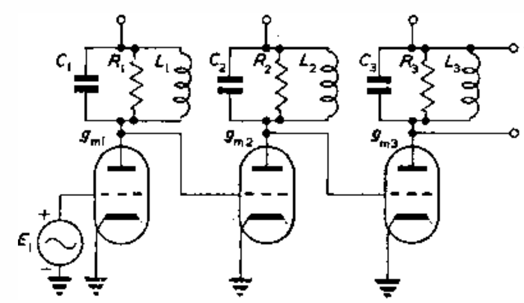

Fig. 9. Circuit diagram for numerical example

Since narrow band representation is being used, it can be considered that all three circuits are tuned to the vicinity of $\omega_0/2\pi$ and that in the numerator $s^3 = (j\omega_0)^3$ and in the denominator $s - s_{p1}' \simeq s - s_{p2}' \simeq s - s_{p3}' \simeq 2j\omega_0$, so that the transfer function becomes:

$$T(s) = -\frac{g_1g_2g_3}{8\,.\,C_1C_2C_3}\cdot\frac{1}{(s - s_{p1})(s - s_{p2})(s - s_{p3})}$$

where

$$s_{p1} = -(\omega_0/2Q_1) + j\omega_0\left(1 + \frac{1}{2Q_1\tan 30^\circ}\right)$$

$$s_{p2} = (\omega_0/2Q_2) + j\omega_0$$

$$s_{p3} = (\omega_0/2Q_3) + j\omega_0\left(1 + \frac{1}{2Q_3\tan 30^\circ}\right)$$

or

$$s_{p1}/2\pi = -50 + j\,(4\,000 + 86{\cdot}6)\text{kc/s}$$

$$s_{p2}/2\pi = -100 + j\,4\,000\text{kc/s}$$

$$s_{p3}/2\pi = -50 + j\,(4\,000 - 86{\cdot}6)\text{kc/s}$$

In accordance with the rules previously outlined it can be established that

unit of x_1 = unit of x_3 = 50kc/s

unit of x_2 = 100kc/s

Scales are drawn corresponding to: (a) absolute scale, (b) scales of x_1, x_2 and x_3. Origins of those last two scales lie opposite ordinates of poles in the absolute scale (Fig. 10).

As the next step the component delay curves are computed with the aid of Fig. 7.

$-(2/B_1)\,\phi'(x_1)$ (curve 1)

$-(2/B_2)\,\phi'(x_2)$ (curve 2)

$-(2/B_3)\,\phi'(x_3)$ (curve 3)

The described method lends itself well to the study of conditions of maximal flatness of delay.

The study of numerical tolerances from normalized delay derivatives is helpful in calculational and experimental evaluation of f.m. distortion of networks, an application where close numerical control of delay derivatives is of utmost importance.

Similar critical control of delay is necessary in application to colour television circuits and, to a lesser degree, in black-and-white television and in general in circuits where good reproduction of transient phenomena is of essence.

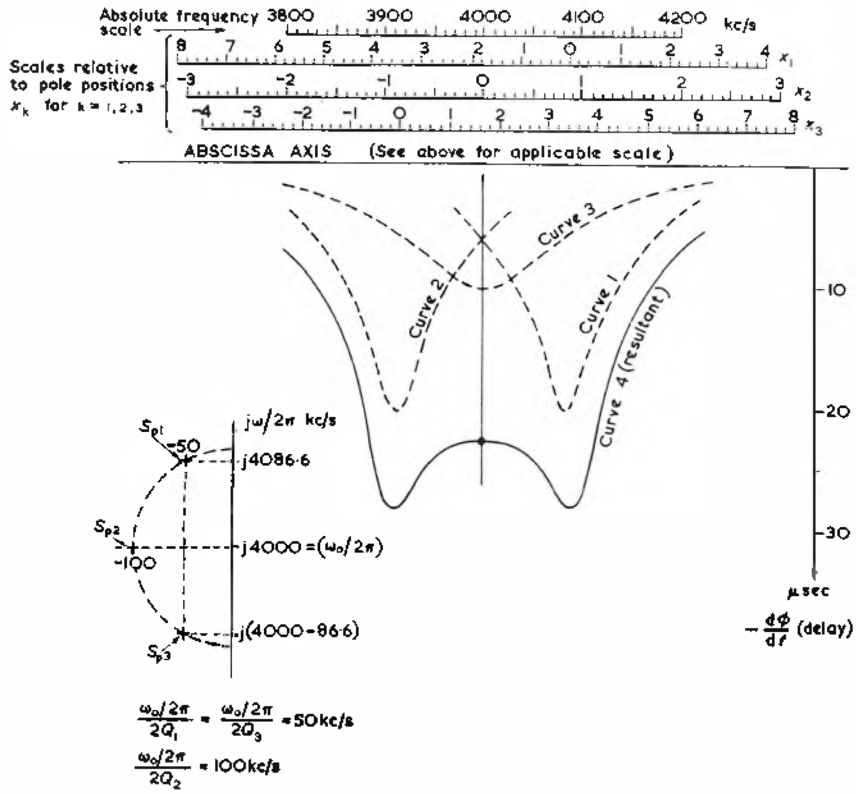

Fig. 10. Delay-frequency function of a triple, maximally flat in amplitude

The resultant sum of the three curves is shown as curve 4. The ordinates of curves 1, 2 and 3 are expressed in seconds if $B_1/2\pi$, $B_2/2\pi$ and $B_3/2\pi$ are in cycles per second. From the shape of curve 4 it is apparent that the maximum amplitude flatness does not correspond to the maximum delay flatness.

From inspection of the shape of the delay curve shown in Fig. 10, one can derive some useful suggestions as to how relative positions of poles should be adjusted in order to decrease the extent of delay variation, noticeable in the pass-band (amounting to about 5μsec). Thus, for example, in order to reduce two pronounced peaks in the delay curve, caused by contributions of poles p_1 and p_3, we should move these poles somewhat farther away from the imaginary axis. The direction of the desired change is evident from the graphical interpretation before the exact position of poles for maximum flatness of delay is worked out. Using other curves from the two families shown in Figs. 3 to 5 and Figs. 6 to 8, similar suggestions for network synthesis may be obtained concerning other properties of networks. Separate consideration of normalized contributions of all natural modes yields particularly profitable results when the nature of the network is such that the step from the required position of poles to the construction of the circuit configuration is relatively easy. This applies to a large category of circuits composed of cascaded valve interstages, each containing single-, double-, or triple-tuned circuits.

If the method of trial and error is applied to the adjustment of pole location the resulting favourable disposition of poles is equally applicable to the case of a low-pass network (in this case $\omega_0 = 0$), as to the case of a narrow-band band-pass network. The method discussed obviously does not pertain to the problem of realizing the desirable pole distribution by suitable circuit element configuration.

As is usual in network synthesis problems, the evaluation of a desirable transfer function forms a separate step from its realization by a detailed circuit of the designer's choice.

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An Optical Scanning and Recording System for a Photo-Electric Optical Bench

By T. N. J. Archard*, and D. H. Rumsey*

On an optical bench a method of lens assessment is to scan a narrow slit of light passed by the lens under test, and to measure the light flux photo-electrically; a graph is then plotted from the results obtained. The apparatus described performs these functions automatically, using impulse motors, a d.c. amplifier and a recording milliammeter.

THE assessment of the performance of a lens is a task of some difficulty and various methods have from time to time been proposed. When the use of the lens involves a photographic emulsion, it is the practice to evaluate the lens plus film combination. When the lens is used with a television camera, the light energy is converted directly into photo-electrons at the photo-cathode, and this process is a linear one over a considerable range of light intensities. Thus a knowledge of the intensity distribution in the image plane of a lens is the starting point of an analysis of the television process. If the test object is a regular pattern of black and white bars, the image will consist of a somewhat similar set of bars, except that the contrast ratio will in general be lower than that of the test object. The way in which the contrast ratio depends on the spacing of the bars forms one method of expressing the performance of a lens. Alternatively the test pattern can consist of a single fine slit and a determination of the intensity distribution in the image plane gives an indication of the image quality.

These methods of lens assessment involve the use of an optical bench on which it is necessary to scan the image produced by the lens under test, in either a sagittal or tangential direction. The method as shown in Fig. 1 uses the test object (a) consisting of a grating pattern or fine slit. Light from this object is passed via the collimating lens (b) to the lens under test (c). To measure the intensity distribution of the light forming the image, the image is magnified by a suitable microscope objective (d), and the thus magnified image is focused on to a thin slit (e). The light passing through the slit falls on to the cathode of a photo-multiplier cell (f), and the electrical signals generated are the measure of the light flux through the aperture.

The scanning of the image from the lens under test was originally carried out by hand-operated micrometer screws. In order to plot the response curve of the lens, it was necessary for the operator to adjust the screw and then note the changes of current (i.e., light flux) on the microammeter. As may be imagined, this process was time-consuming and involved considerable tedium on the operator. It was for these reasons that automation of the scanning and recording system was decided upon.

In order to keep the weight and cost of the scanning system to a minimum, simple impulse motor drives are used to give the necessary sagittal and tangential scans. These motors are shown as modified in Fig. 2 and are easily recognizable as 'Government surplus' units such as were used in the RAF transmitter/Receiver T1196, etc.

The motors are provided with pulses from an electromechanical pulse generator as shown in Fig. 3. It was decided that one pulse should equal one micron of movement in the scanning head. As the ratchet wheel in the impulse motor has 104 teeth, a gear train is employed to drive the existing $\frac{1}{2}$mm pitch micrometer screws to give this movement in increments of 1μ. The pulse generator also

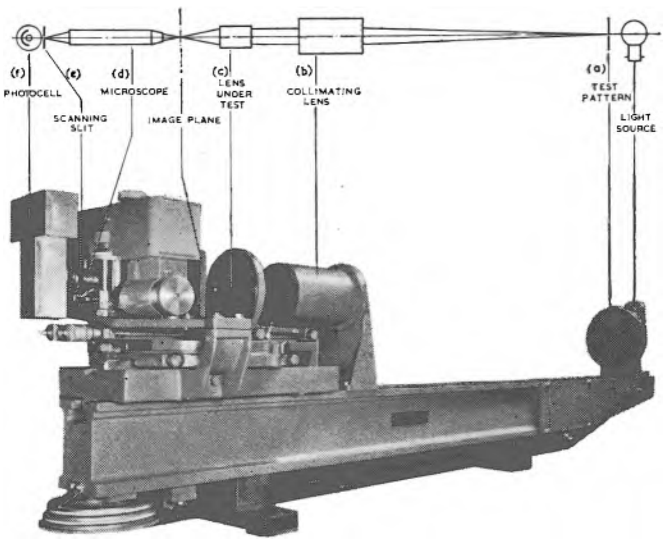
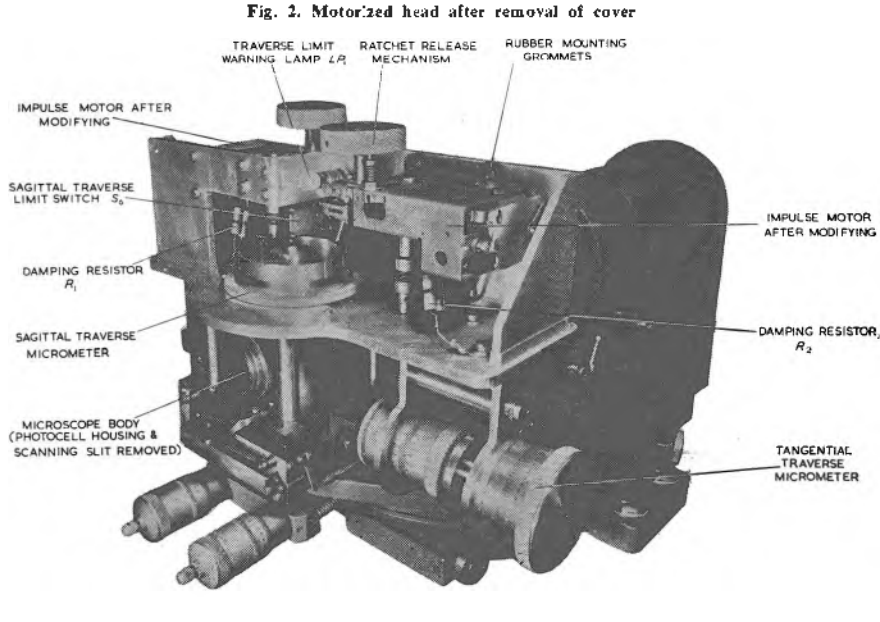


Fig. 1. General arrangement of optical test bench

Fig. 2. Motorized head after removal of cover



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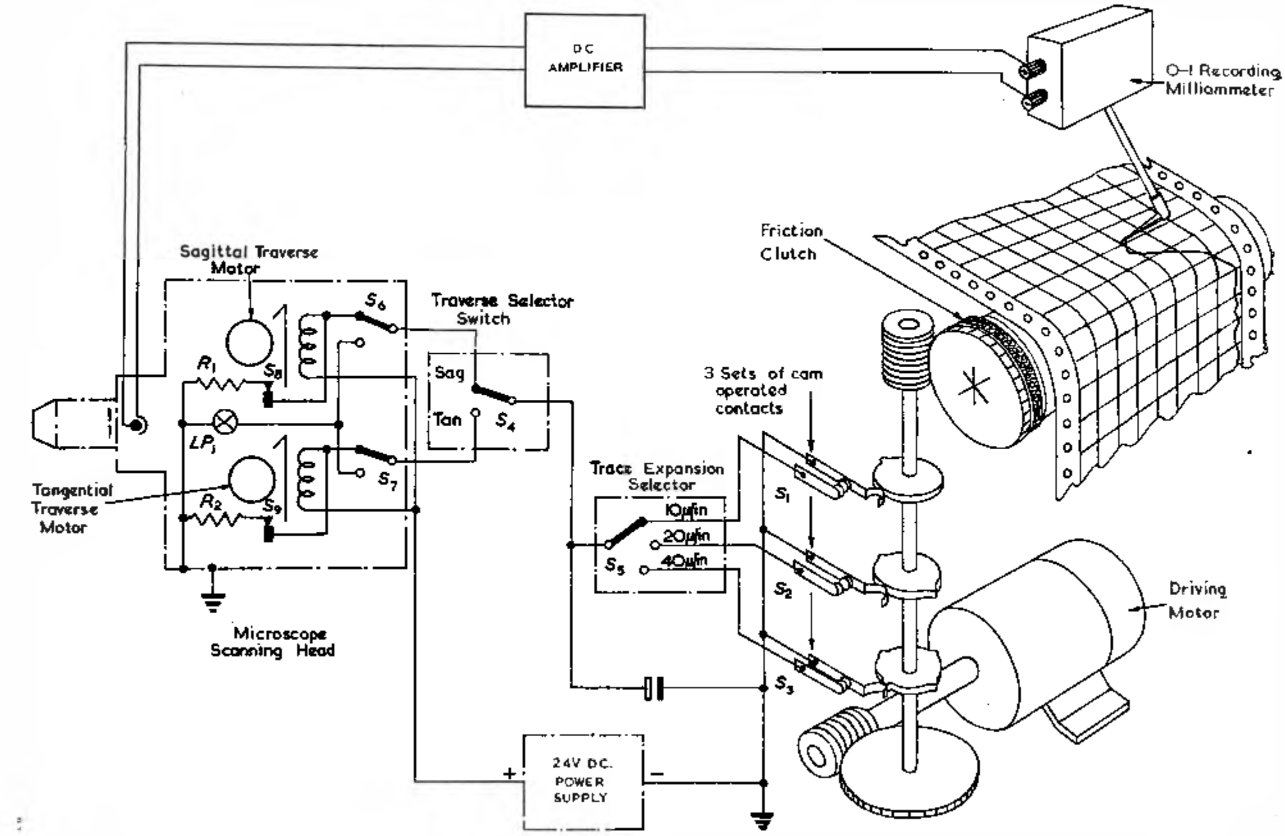
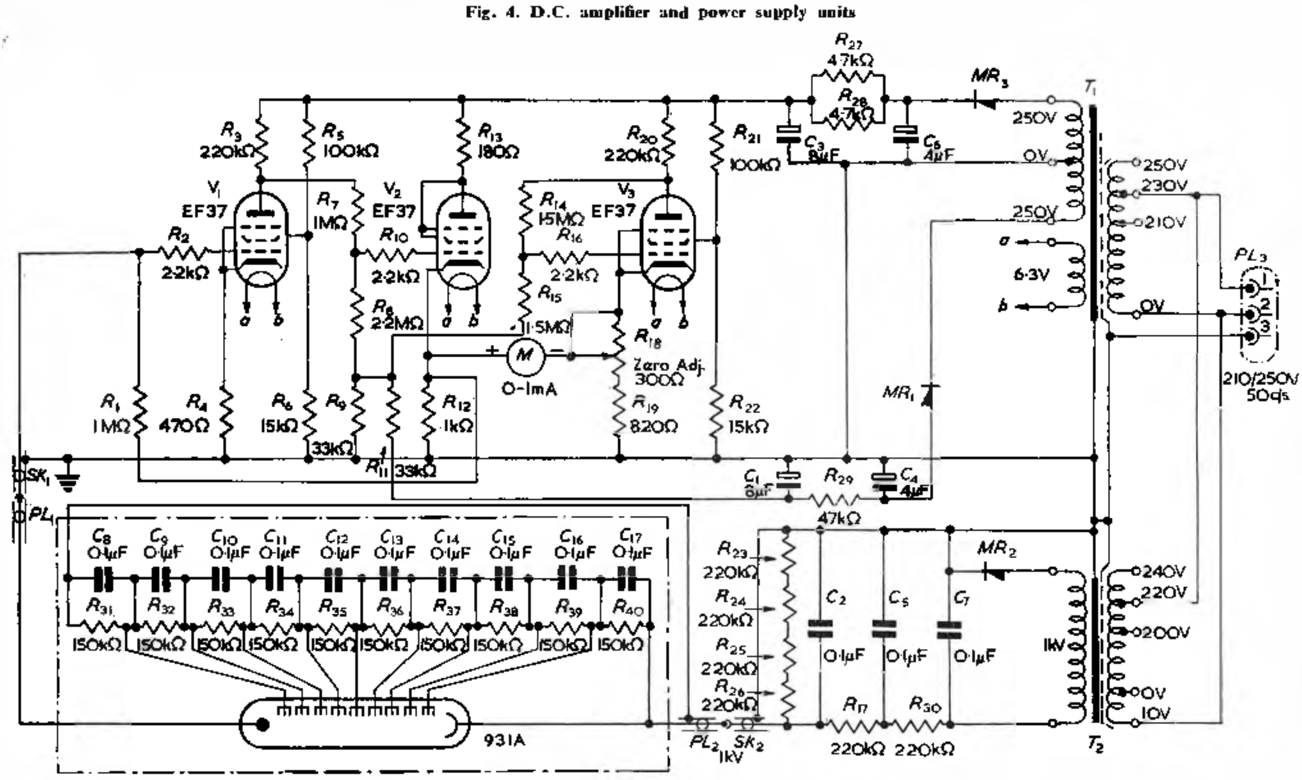


Fig. 3. Basic arrangement of scanning and recording system

Fig. 4. D.C. amplifier and power supply units



drives through a worm and wheel the paper feed mechanism of a recording 0 to 1 milliammeter. The meter itself is deflected according to the signal derived from the output of the photocell, as amplified by the d.c. amplifier. Using different cam operated contacts ($S_{1,2,3}$) in the pulse generator, various ratios of trace expansion may be obtained. In this case, the ratios are 10, 20 and 40μ (or pulses) per inch of chart movement. The ratios are selected by switch S_5. Either sagittal or tangential scanning is selected by switch S_4. To minimize vibration due to the intermittent motion of the impulse motors, they are mounted on rubber grommets and an electro-mechanical damping arrangement is employed. Using the existing armature-operated contacts (S_8S_9) on the motors, a voltage is applied to either coil via resistors R_1 or R_2 before the return of the armature is completed. This voltage, of approximately half the working amplitude, has the effect of reducing the velocity of the armature so that the remainder of the stroke is retarded. In order that the initial setting up of the micrometer heads can be performed by hand, it is necessary to make a release mechanism for the ratchet drive on the impulse motors, this is shown in Fig. 2. Over-run of the heads is prevented by the operation of micro-switches S_6 and S_7, Figs. 2 and 3. Their operation causes the current to the impulse motors to be switched to the lamp LP_1, giving a visual indication to the operator that the limit of travel has been reached.

The d.c. amplifier is shown in Fig. 4 and the purpose of this amplifier is to provide sufficient power to operate the 0 to 1mA meter from the photocell output. V_1 and V_2 form a two-stage d.c.-connected current amplifier with feedback taken from the cathode of V_2 to the bottom of the photocell load resistor R_1. The voltage gain is therefore approximately unity, but the current gain is approximately 1 000 times (i.e. R_1/R_{12}).

To make the meter independent of supply variations and the effects of temperature change, the negative terminal of the meter is connected to the cathode of a third valve V_3 whose anode and grid circuits are made similar to those between V_1 and V_2. The grid of V_3 is therefore (except for signal voltages) held always at the same potential as that of V_2.

The meter then reads the unbalance in the bridge formed by $V_{2,3}$, R_{12} and $R_{18,19}$, whenever a signal voltage appears at the grid of V_2.

Reverting to the mechanical arrangement, it may be argued that the drive system is not fool-proof, in as much as that if the impulse motors fail to engage a tooth on the ratchet wheel, an inaccurate reading would result; this is true, but in practice, the motors have proved very reliable, and no trouble has been experienced. Copies of the traces that have been provided by the apparatus are reproduced in Fig. 5. The Fourier transforms of these curves give the modulation/frequency responses of the lens under test.

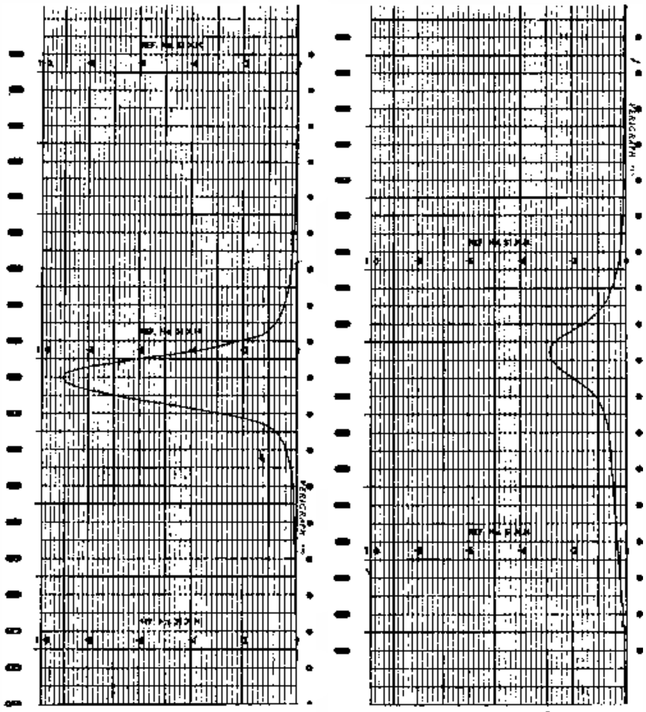


Fig. 5. Typical traces showing response of lens to a fine slit pattern

It is of interest to note that the time taken to produce a curve has been cut from 20min to approximately 2min (including setting up) and the tedium from the operator's point of view has been completely removed.

Acknowledgments

The authors are indebted to the Chief Engineer, BBC, for the permission to publish this article, and to colleagues of the Research Department, BBC, for their assistance and co-operation.

Industrial Television in Czechoslovakia

A photo-conductive type television camera tube for industrial uses has recently been developed in Czechoslovakia and is known as the 'Quanticon'.

The 'Quanticon' has been developed on the basis of Vidicon. A cathode system, with an electron gun, is inserted at the end of a glass tube. The image of the scanned object or scene is thrown by means of a lens on to a signal plate mounted in the frontal tube wall. The signal plate consists of a small transparent plate provided with a photo-sensitive layer on the surface facing the cathode; and it is supplied with a positive electric charge of about 20V. An electron beam is projected on to the conductive layer in which it causes a negative charge. This beam is directed by means of a deflecting system so that it scans each point on the layer 25 times per second. Thus a potential difference arises between the signal plate and the photo-conductive layer. As the resistance of the layer diminishes in respect of the more clearly illuminated parts of a picture, the charge created in the layer diminishes also. Illuminated spots have, consequently, a higher potential than those less well or not illuminated.

The changes of potential are registered by the signal plate, which transfers them to a connected resistance between tubes. In this way the image is formed.

Compared to most other camera tubes, 'Quanticon' is of simple design and, as a result, of small dimensions. The tube is about 4in long and about 1in in diameter. It functions reliably under an illumination of as little as 35 luxes; and it has been tested in 390-line transmissions.

The tube satisfies the requirements of the industrial use of television very well. It has made possible the construction of a camera which weighs 7lb and is about $8\frac{1}{4}$in long by $5\frac{3}{4}$in deep. The camera is fitted with a remote-control attachment and linked by cable with an impulse-emitting and amplifying unit and a receiver (screen approximately 9 by 12in). Distance of transmission is up to 2 000ft, but with a modulator and use of a standard receiver this may be more than doubled.

Uses to which it is intended to put industrial television in Czechoslovakia are: in the production of electricity, to keep under observation the functioning of boilers, inaccessible equipment, and parts of transmission systems; in atomic power plant, at various key points; in mining, to assist in control of transport and inspecting automatic equipment; in foundries, to watch processes in blast furnaces, etc.; in civil engineering, to afford 'links' between various sectors of a big site; and in advanced technical and medical instruction.

The 'Quanticon' tube has been developed at the Electronics Research Institute by technicians working under Ing. Bohumil Holy, and work on industrial television cameras has been conducted by another group, headed by Ing. Eduard Severin.

TRANSISTOR RELAXATION OSCILLATIONS

By F. J. Hyde*, and R. W. Smith*

The continuous relaxation oscillations occurring in a point-contact transistor circuit including a parallel CR combination in shunt with the emitter input have been studied as functions of emitter bias and ambient temperature.

The rapid and substantially linear increase in frequency for small values of emitter bias lends the circuit to telemetry applications. The dependence of the frequency on temperature may be made small over a restricted frequency range if a low resistance thermistor comprises part of the external circuit resistance.

SOME aspects of the operation of the point-contact transistor oscillator of Fig. 1 have been considered by other authors[1-5].

The purpose of the present investigation has been to study the circuit operation as a function of temperature and of various circuit parameters.

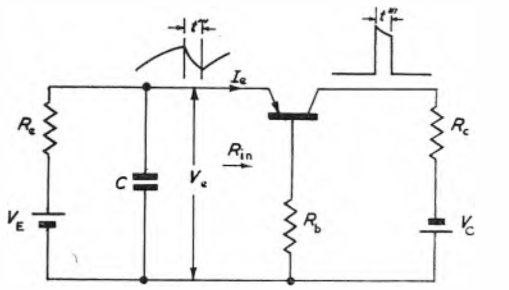


Fig. 1. Circuit of transistor relaxation oscillator

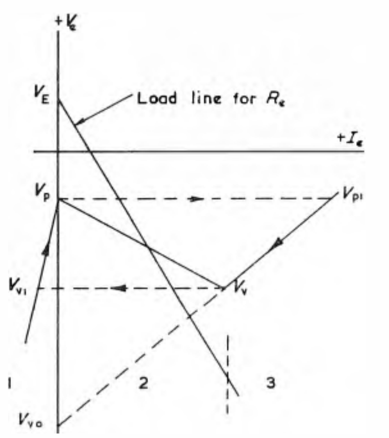


Fig. 2. Linearized emitter input characteristic

Circuit Operation

This is usually described in terms of the static emitter input characteristic with fixed collector supply voltage which is assumed to have the form shown in Fig. 2. Necessary conditions for oscillation may readily be derived[2] if in the negative-resistance region 2 it is assumed that the transistor may be represented by the usual low-frequency equivalent-T circuit, in which the internal current amplification factor is constrained to vary with frequency as $\alpha = \alpha_0 / (1 + j\omega\tau)$. These conditions are: (a) that the emitter circuit load line should intersect the negative-resistance region of the characteristic, (b) $-\omega_c > (R_e + R_\infty'')/CR_eR_o''$, where $\omega_c = 1/\tau$ is the angular α cut-off frequency and R_o'' and R_∞'' are respectively the zero frequency and infinite frequency input resistances of the transistor under the assumptions made. Single, double and treble primes on parameters denote values appropriate to regions 1, 2 and 3 respectively of Fig. 2. These conditions have been experimentally verified by Farley[2].

Provided that C is large, a graphical description of the circuit operation is that the operating point moves cyclically round the trapezium $V_p \rightarrow V_{p1} \rightarrow V_v \rightarrow V_{v1}$. The speed of movement from V_{v1} to V_p and from V_{p1} to V_v is determined by the time-constant of C in parallel with both the emitter load and appropriate input resistance, multiplied by logarithmic terms involving these resistances together with the values of the voltages V_E, V_p, V_v and V_{vo}, as was demonstrated by McDuffie[5]. The speed of movement from V_p to V_{p1} and from V_v to V_{v1} is dominated by the internal parameters of the transistor[2]. These latter transitions are very rapid for most transistors ($< 1\mu$sec) so that for sufficiently low recurrence frequencies there is negligible error in assuming that the transitions in the regions 1 and 3 completely determine the period of the oscillation The recurrence frequency is then given by

$$f = (t' + t''')^{-1} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where

$$t' = \frac{CR_eR_o'}{R_e + R_o'} \ln \left\{ \frac{(V_v - V_p)R_e + (V_v - V_E)R_o'}{(V_p - V_E)R_o'} \right\} \quad . (2)$$

$$t''' = \frac{CR_eR_o'''}{R_e + R_o'''} \ln \left\{ \frac{(V_p - V_{vo})R_e + (V_p - V_E)R_o'''}{(V_v - V_{vo})R_e + (V_v - V_E)R_o'''} \right\} \quad \ldots\ldots\ldots\ldots (3)$$

At low frequencies (small values of V_E) $t''' < t'$ so that the collector waveform is a positive-going pulse of duration t''' with rapid rise and fall times. The appropriate emitter and collector voltage waveforms are shown in Fig. 1.

Actual emitter input characteristics encountered approach the idealized form of Fig. 2 to varying degrees of approximation. For the present investigation a type EW51 transistor was used with the following operating conditions:

$T = 20°$, $R_e = 27\text{k}\Omega$, $R_b = 470\Omega$, $R_c = 2{\cdot}2\text{k}\Omega$,
$C = 0{\cdot}209\mu\text{F}$, $V_c = -18\text{V}$.

In region 1 the characteristic is linear with a slope $R_o' = 52\text{k}\Omega$, the peak point V_p is sharp (occurring at $I_e = +7\mu\text{A}$); in region 3 the characteristic is substantially linear with mean slope $R_o''' = 400\Omega$ and with slight rounding towards the valley point V_v. The slope in the negative-resistance region 2 is variable, but since transitions in this region are assumed to be of negligibly short duration, this is unimportant.

Dependence of Pulse Recurrence Frequency on Emitter Voltage

In Fig. 3 the experimentally-observed and theoretical dependence of p.r.f. on emitter bias voltage V_E are compared for the above operating conditions. The theoretical curve is calculated using equations (1) to (3) in which the experimentally-derived values of the individual parameters

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have been inserted. The experimental measurements were made by comparing the p.r.f. and frequency of a calibrated sine wave generator.

An interesting feature is the initial substantially linear and rapid increase of p.r.f. with emitter bias. This lends the circuit to telemetry applications in which a rectified radio frequency signal (d.c.) of variable amplitude can be used to generate a monotonically-varying audio frequency, provided that the emitter bias is limited to values less than that for which the p.r.f. is a maximum.

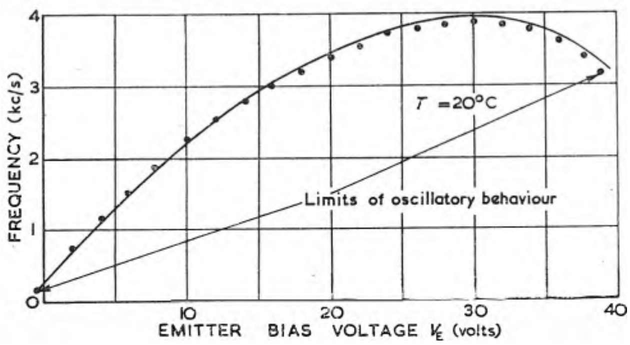

Fig. 3. Comparison of experimental and theoretical dependence of frequency on emitter bias

Solid line is theoretical, dots are experimental

Transistor type EW51, R_e = 27kΩ, R_b = 470Ω, R_c = 2·2kΩ, V_c = −18V, C = 0·209μF

Circuit Stability

Temperature

The usefulness of the circuit is to a large extent governed by its sensitivity to temperature change. Accordingly, first the emitter input characteristic and secondly the circuit operation were studied as a function of temperature for various combinations of circuit parameters. For convenience, only the transistor was subjected to the full temperature changes, the possibility of changes in the values of the other circuit components being ignored. The emitter input characteristic was found to vary little in shape between −30°C and +40°C. The dependence of the relevant parameters on temperature is listed in Table 1.

TABLE 1

Variation of Emitter Input Circuit Parameters with Temperature

T°C	V_p (V)	V_v (V)	V_{vo} (V)	R_o'(kΩ)	R_o''' (Ω)
−30	−1·09	−1·80	−2·5	78	370
−20	−1·10	−1·81	−2·5	73	385
−10	−1·10	−1·83	−2·55	68	385
0	−1·10	−1·85	−2·55	66	385
10	−1·11	−1·88	−2·55	56	390
20	−1·13	−1·91	−2·6	52	400
30	−1·15	−1·91	−2·6	50	405
40	−1·17	−1·95	−2·6	48	410

It will be noted that the voltage parameters vary slowly with temperature, the most marked effect being the decrease in the input resistance of region 1 with increasing temperature.

The theoretical effect of these parameter changes on the p.r.f. may be studied using equations (1) to (3).

(a) t' (dead-time). Accompanying the decrease of R_o' with increasing temperature there is a decrease of the coefficient of equation (2). The logarithmic part, however, will increase at the same time.

(b) t''' (pulse width). The increase of R_o''' with increasing temperature will cause increases in both the coefficient and the logarithmic part of equation (3). There will result an increase in pulse width with increasing temperature.

The overall effect on the p.r.f. will be a function of the absolute values of the parameters governing the above trends, so that it is possible under differing circumstances to expect either an increase or a decrease of p.r.f. with changing temperature.

For the working conditions for which Table 1 applies it was found that for low values of external emitter bias there was in fact a slight increase of p.r.f. between −30°C and +40°C (e.g. 3c/s in 347c/s), while at higher values of bias there was a decrease of p.r.f. over the same temperature range (e.g. −82c/s in 1 030c/s). The results obtained at these two extremes of temperature are recorded in Table 2 for the circuit conditions already described and also for different combinations of emitter and base resistances. For these different combinations concomitant changes of C were made so that the frequency of operation was kept in the low audio range.

TABLE 2

Dependence of p.r.f. on Temperature for Different Combinations of R_e and R_b with V_c held constant

V_E (V)	R_b(Ω)	R_e (kΩ)	C(μF)	FREQUENCY (c/s) T= −30°C	T= +40°C	Δf (c/s)
(*a*)						
0	470	10	0·5	335	323	− 12
		27	0·2	347	350	+ 3
		68	0·1	365	372	+ 7
	1000	10	0·5	330	320	− 10
		27	0·2	357	360	+ 3
		68	0·1	295	308	+ 13
	2200	10	0·5	325	302	− 23
		27	0·2	420	402	− 18
		68	0·1	347	350	+ 3
(*b*)						
3	470	10	0·5	1030	948	− 82
		27	0·2	1020	955	− 65
		68	0·1	1040	990	− 50
	1000	10	0·5	760	700	− 60
		27	0·2	1010	970	− 40
		68	0·1	1010	1000	− 10
	2200	27	0·2	880	800	− 80
		68	0·1	1060	1000	− 60
(*c*)						
8	470	27	0·2	2100	1900	−200
		68	0·1	2050	1900	−150
	1000	27	0·2	1900	1730	−170
		68	0·1	2080	2000	− 80
	2200	68	0·1	1810	1650	−160
(*d*)						
16	470	27	0·2	3200	2850	−350
		68	0·1	2970	2640	−330
	1000	68	0·1	3200	3000	−200

The maximum oscillation frequency decreases with increasing values of R_b and decreasing values of R_e.

It is seen that for low values of emitter bias either an increase or a decrease is possible but as the bias is increased there is a general tendency for the p.r.f. to fall with increasing temperature.

From the data it appears that if an oscillator is required to work at a fixed low frequency independent of the temperature then this can be achieved by a suitable choice of components. The choice of zero external emitter bias as in Table 2(a) leads to circuit simplicity. On the other hand in order that telemetry of a varying signal level may be carried out, a range of oscillation frequency is required. It will then be necessary to compensate for the natural fall in frequency with increasing temperature at the higher frequencies. A suitable way of doing this is to incorporate a thermistor or parallel combination of thermistor and resistor as part of the emitter load resistance. This will have the effect of decreasing R_e as the temperature increases. In so far as f' is concerned, the decrease of the coefficient will be augmented, while the increase of the logarithmic term will be abated. For f''' the increase of the coefficient will be abated, while the increase of the logarithmic term will be augmented. The overall effect to be expected is an increase in p.r.f. over that obtained with no thermistor present. In view of the fact that there is only a small change of p.r.f. with respect to temperature for low emitter bias in the uncompensated circuit and that this may be a positive change, it is evident that in the compensated circuit any improvement produced in temperature-stability at higher frequencies will to some extent be offset by a worsening at the low frequencies. Provided that a limited frequency range is chosen, however, the compensated circuit will give considerable improvement.

In Fig. 4 are shown the percentage change of the frequency at $-30°C$ in going from $-30°C$ to $+40°C$ for the uncompensated circuit in which $R_c = 27k\Omega$, $R_b = 470\Omega$, $R_e = 2{\cdot}2k\Omega$, $V_c = 18V$, $C = 0{\cdot}209\mu F$ and that for a compensated circuit in which R_e comprises a $27k\Omega$ resistor in series with a parallel combination of a Type A1311/100 thermistor and $4{\cdot}7k\Omega$ resistor. For the uncompensated circuit the change of frequency with temperature was monotonic. For the compensated circuit, however, this was the case only at low frequencies (<1kc/s). At higher frequencies there was first a rise and then a fall in frequency with increasing temperature, the maximum occurring somewhat below 0°C. Such maxima were not pronounced.

Time

For the compensated circuit as above with zero emitter bias the p.r.f. was measured as a function of time. Over a period of 250 hours the frequency was found to be invariant within measuring accuracy, which was $\pm 0{\cdot}2$ per cent.

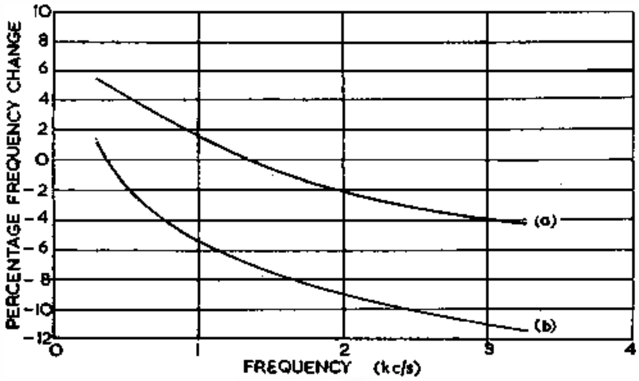

Fig. 4. Percentage change of the frequency at —30°C for a temperature change from —30°C to +40°C (a) compensated (b) uncompensated

Conclusions

It has been found that there is in general a tendency for the p.r.f. of this relaxation oscillator to fall with increasing temperature, but some improvement can be achieved by including a thermistor in the emitter circuit. For such a compensated circuit the temperature coefficient of the p.r.f. is of the order of 0·03 per cent/°C over a range of frequencies although better stability than this can be achieved at single frequencies.

Under suitable operating conditions it is possible to produce a change of p.r.f. of ten to one by varying the emitter bias.

Acknowledgments

The work described above was carried out as part of the programme of the Radio Research Board and this article is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

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The Precise Measurement of Distance

A new system of microwave measurement of distance has recently been introduced to this country by Cooke, Troughton & Simms in conjunction with Tellurometer (Pty) Ltd. of Cape Town. The invention of Mr. T. L. Wadley, of the South African Telecommunication Research Laboratory, the new Tellurometer operates in the 10cm wavelength region and measures the transit of radio waves over the length to be determined, with an accuracy of a fraction of a millimicrosecond.

The Tellurometer has a range of at least twenty miles and its introduction will greatly assist surveying, civil engineering and construction work of a major nature.

To measure a single line, one Master station and one Remote station are necessary. Measurements are made from the Master station and an operator is required at the Remote station. A built-in duplex radio telephone circuit is included in the system and is used in the measurement procedure. If a number of Remote stations are employed, a number of lines may be measured from the Master station in the course of one setting up.

The instruments at Master and Remote stations are physically similar, but are not interchangeable. Both the Master and Remote instruments are self-contained units with built-in aerial systems.

The accuracy of the instrument is such that the distance between London and Reading could be measured with an error not likely to exceed six or seven inches or perhaps less. In other words, the probable error is one part in 300 000 + 2 inches. The instrument has a range of from about 500 feet to 100 miles, but for practical purposes the maximum working range is about 35 miles.

Royal Society International Geophysical Year Base Established

The advance party of the Royal Society Antarctic Expedition, consisting of ten men led by Surgeon Lieutenant-Commander David Dalgliesh, R.N., returned to London on March 13 on board the M.S. Magga Dan.

They sailed in the M.V. Tottan from Southampton on November 22, 1955, and landed on January 6, 1956, at 75°31'S, 26°36'W, on the shores of a bay, where a gentle slope lying between two ice headlands gave easy access to the hinterland. This bay has since been named Halley Bay after Edmond Halley, the eminent geophysicist and sometime Secretary of the Royal Society, who was born in 1656.

This advance party proceeded to erect a hut 130ft in length, 2 miles inland from Halley Bay, in preparation for the arrival of the main party of 21 men in January 1957. This hut and the surrounding huts which have been built to house other scientific equipment, have been named Royal Society Base and it forms the principal United Kingdom contribution to the Antarctic programme of the International Geophysical Year (July 1957-December 1958). Other contributions in this region are being made by the Commonwealth Trans-Antarctic Expedition at Shackleton, and by many of the bases of the Falkland Islands Dependency Survey (FIDS). In all, 12 nations will man 58 stations on the Antarctic continent and the Sub-Antarctic islands during the International Geophysical Year (IGY).

Tolerance Limit of Resistors in Binary Scaling Units

By B. M. Banerjee* and (Miss) Sneha Choudhury*

The operating ranges of Higinbotham type binary scaling units, for various sets of values resistors and for the valve types 6SN7, 6SL7, E90CC and 12AT7 have been found experimentally. The variation of the valve electrode voltages have also been studied. The tolerance limits have been found by introducing known amounts of dissymmetries. Effects of supply voltage variation are also presented.

Contrary to the usual belief, it has been found that the tolerance limits are wider than ±20 per cent, when the values of the common cathode resistor and the output resistor of the previous binary, are optimized. A circuit with a much greater dissymmetry could be made to work when these two resistors are given suitable values. The 'range' increases when the output resistance of the previous binary decreases.

HIGINBOTHAM[1] type binary units are commonly used in scaling circuits required for nuclear physics experiments. The idea is prevalent[1,2] that the resistors used in such units must be matched carefully and held to close tolerances for proper functioning of these units. An analytical approach to the problem has been made by Pressman[2]. Experimentally it is found, that actual tolerances are much wider than predicted by Pressman. This is because the limiting conditions assumed by Pressman in his analysis are idealized and arbitrary. Estimation of these limits in practical cases is almost impossible as they really depend on too many factors. Actual data is best obtained experimentally and these are being presented in this article.

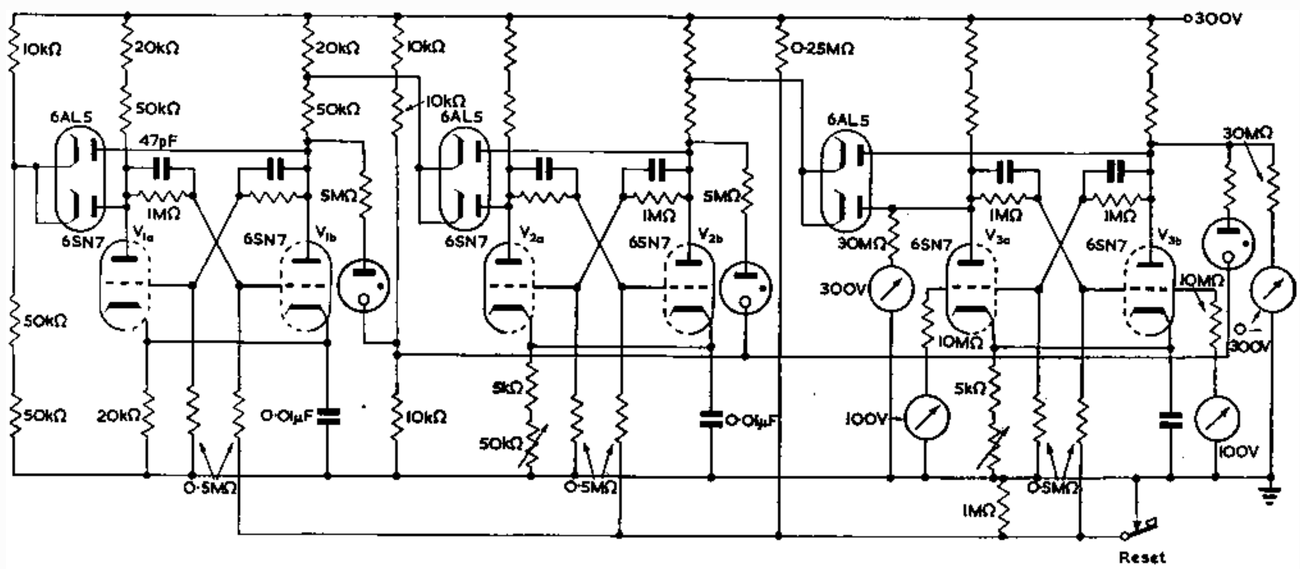


Fig. 1. Experimental arrangement

Experiments

Investigations were carried on with the circuit arrangement shown in Fig. 1. Three binary units have been connected in cascade. The components (e.g. anode resistors), effects of whose variation are to be studied, are simultaneously changed in all the stages. The first two stages are for application of switching pulses of suitable size and form as are delivered by scaling units, to the third stage, the stage under observation. Accordingly the components of the first two stages were adjusted to the optimum values. With a given set of suitable resistor values, the cathode resistor R_k is continuously altered till the circuit ceases to function or functions imperfectly. The ranges of cathode resistance over which the circuit operates are shown by curves similar to Fig. 2.

Experimental Findings

It is clear from Figs. 2, 3, 4, 5, 6 and 7 that the circuit operates over a wide range of cathode resistances. It works over a smaller range of cathode voltages. The 'range' may be conveniently defined as the ratio of the maximum operating value to the minimum operating value. It is a maximum for matched values of feedback (anode to grid) and grid (grid to negative line) resistance pairs (Fig. 6). It is also affected by the inequality of anode resistances, but not as seriously. It shrinks as the anode resistance is diminished below the valve internal resistance value. It also depends on the output resistance of the preceding binary.

Below a certain cathode resistance or cathode voltage, the circuit generates multivibrator type oscillations. With matched values of anode, grid and feedback resistances, the binary unit scales correctly as soon as the cathode resistance (or voltage) is large enough to stop multi-

* *Institute of Nuclear Physics, Calcutta, India.*

* *About five to ten times the anode resistance value.*

vibrator oscillations. The circuit shows two stable states up to a very large value* of cathode resistance (and a critical upper limit of cathode voltage beyond which current is shut off in both the valves). It continues to be triggered by the preceding unit up to a maximum cathode resistance value usually several times the minimum value†. With accurately matched values of feedback and grid resistances, the operating range of cathode resistance increases as the output resistance of the preceding binary decreases. As the triggering pulse from the preceding binary is received, the lowered potential of the output point (of the preceding binary) will catch up and hold the non-conducting anode potential to this lower value. If the conducting

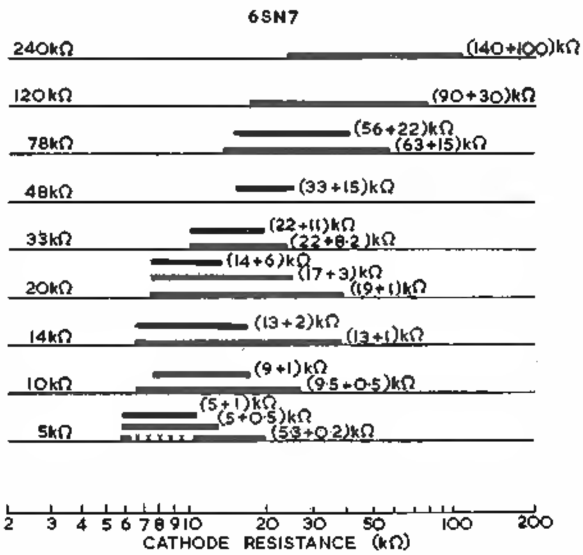


Fig. 2. Operating ranges for a binary using 6SN7
(for explanation see paragraph entitled "Data Presented")

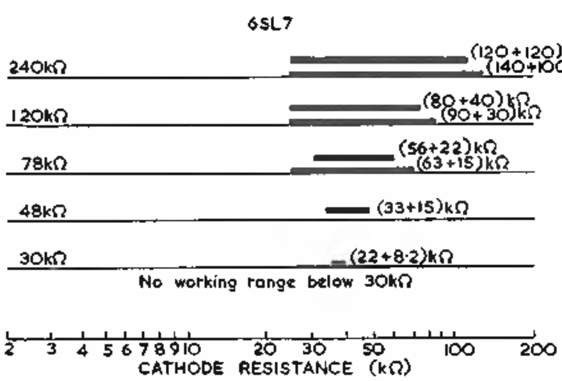


Fig. 3. Operating ranges for a binary using 6SL7

anode potential is not sufficiently below this value, the switch over to the other state will not take place. When the cathode resistance is large, the conducting anode potential remains high. Reduction in the value of output resistance therefore extends the operating range to greater values of cathode resistance, because the lowering of potential of the output point below the h.t. supply is smaller. This increase in range continues till the output pulse fails to overcome the reverse voltage at the switching diodes*, when the circuit ceases to be triggered. When the feedback

to grid resistance ratio in the two chains are unequal, the range of cathode resistance for proper circuit operation is smaller, and decreases as the output coupling resistance is diminished below some value that is dependent on the degree of unequality between the feedback chains, (Fig. 6).

The current drain in the interpolator neon produces an unbalance in the circuit and reduces the operating range. This begins to be apparent when anode resistor values exceed 100kΩ. A reduction in the current drain achieved

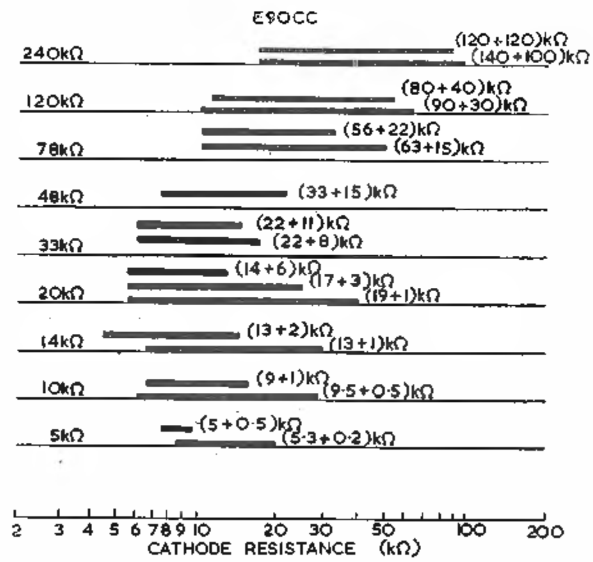


Fig. 4. Operating ranges for a binary using E90CC

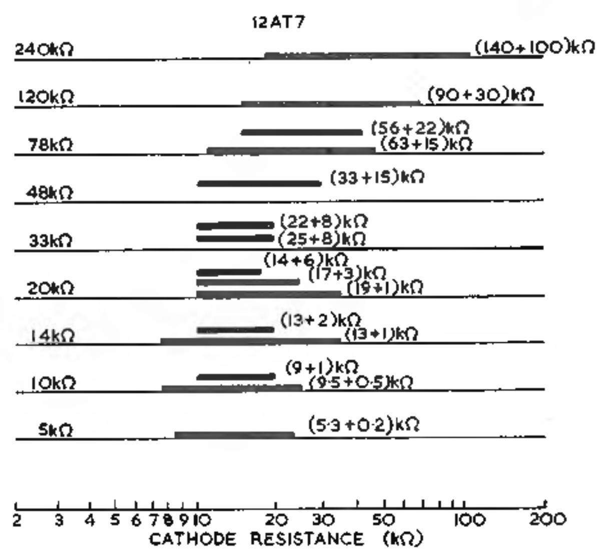


Fig. 5. Operating ranges for a binary using 12AT7

by increasing the neon series resistor from 1MΩ to 5MΩ, produces appreciable increase in the operating range in a number of cases.

Figs. 7, 8, 9, and 10 show the variation of conducting valve and non-conducting valve electrode voltages as the common cathode resistance (R_k) is varied. It is seen that the anode voltage of the conducting valve increases continuously with cathode resistance. The conducting valve grid voltage increases almost linearly with R_k and as it approaches the limit,

$$(V_B - R_a I_L)\,\frac{R_g}{R_g + R_a + R_f}$$

† *The minimum value of cathode resistance is a little greater than that at which multivibrator oscillations stop.*

* *A reverse voltage is developed between the cathode anode of the switching diode due to the flow of feedback chain current and the interpolator neon current, through the anode resistor. This is small when the feedback chain resistances are large compared to the anode resistances.*

where V_B = H.T. supply voltage
I_L = interpolation neon current
R_a = anode resistance
R_g = grid resistance
R_f = feedback resistance

its rate of increase gets drastically reduced, and then it attains a more or less constant value. The non-conducting grid voltage goes on increasing with cathode resistance. The common cathode voltage follows the conducting valve grid voltage. It is somewhat higher than the grid beyond the knee. The non-conducting anode voltage has two discrete values—the higher value when the output anode of the previous binary is non-conducting, and the lower one when it is conducting.

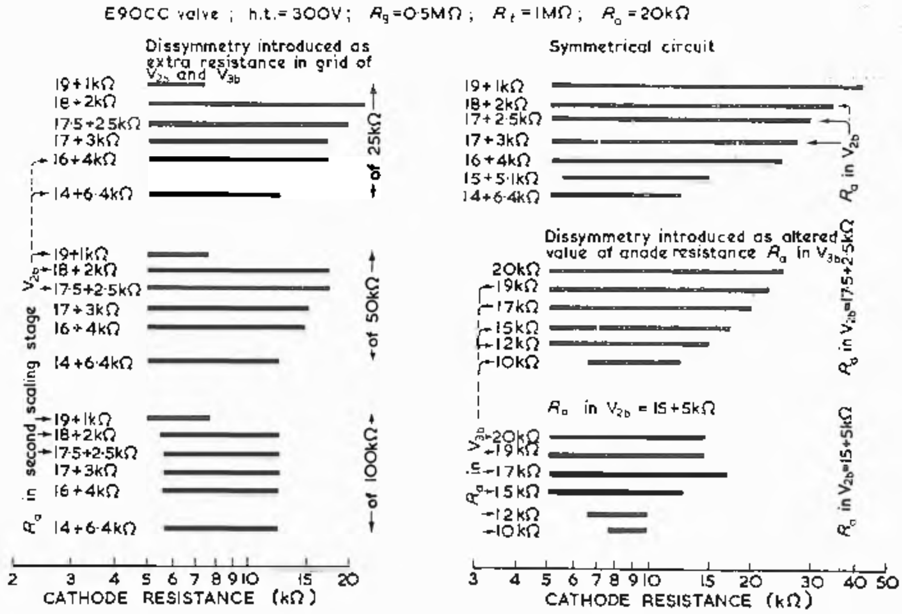


Fig. 6. Effect on operating ranges of dissymmetries

Fig. 11 shows that the effect of h.t. supply variations on the operating range is small. Also these binary units operate quite satisfactorily over a range of feedback to grid resistance ratios from 1:1 to 3:1. Further, the circuit would work over a small range of R_k values, even with very large dissymmetries in grid resistors.

Data Presented

Figs. 2, 3, 4 and 5 present the operating ranges of a circuit (Fig. 1) that uses accurately matched values of resistors using 6SN7, 6SL7, E90CC and 12AT7 valves. Fig. 6 presents the same information for a circuit into which dissymmetry in the form of altered resistance in one of the grids, or one of the anodes, has been introduced deliberately. Fig. 7 shows how the electrode voltages vary in an unsymmetrical circuit. The valves 6SN7, 6SL7, E90CC and 12AT7, represent the four basic types that have distinctly different characteristics. They are also commonly used in this application.

Operating ranges are shown as patches that extend from the lowest cathode resistance at which the circuit starts scaling correctly to the highest cathode resistance at which the circuit ceases to be triggered. Operating ranges of different values of total anode resistance (indicated at the left) are grouped together. The preceding binary anode resistances are indicated at the right. Thus one finds three patches indicating the operating ranges for a total anode resistor of 20kΩ using E90CC tube in Fig. 4. The lowest one coincides with the base line of 20kΩ and has (19 + 1)kΩ written at its right. The output generating resistance of the preceding binary for this case is 1kΩ and the patch shows that the range extends from 5·5kΩ to 42kΩ. Besides, as the patch has been drawn coincident with the base line of 20kΩ, the output resistance of 1kΩ is the minimum value that will trigger the following binary. Similarly, for 48kΩ total resistance, data for an ouput resistance of 15kΩ only, which is not the minimum resistance, is presented. The cathode resistance range in this case extends from 7·5kΩ to 22·5kΩ. And for 240kΩ, the minimum value for output generating resistance is 100kΩ, which gives a working range from 18kΩ to 100kΩ.

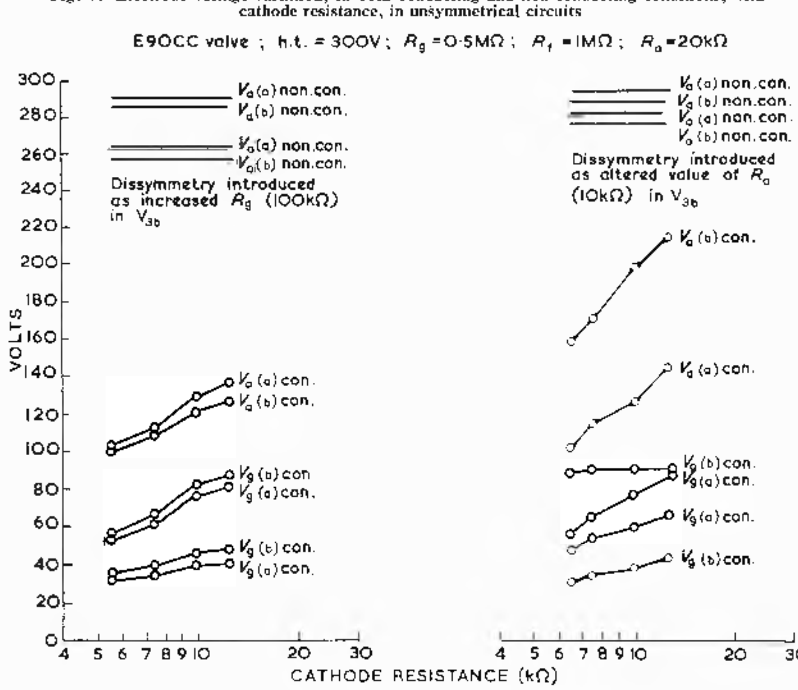
Conclusion

The data presented (e.g. Fig. 6, 7 and 11) makes it clear that a variation of even 20 per cent of any of the resistors is unlikely to throw the binary out of operation if these values are optimized.

The procedure that may be followed in the design of such binary units is first of all to settle the values of anode, feedback and grid resistors that one desires to use. One can then utilize the data presented in Figs. 2, 3, 4 and 5 for selection of the cathode resistance and output resistance.

Valve type E90CC (or its near equivalent 6J6) appears to be the most desirable in this application. Its low anode

Fig. 7. Electrode voltage variation, in both conducting and non-conducting conditions, with cathode resistance, in unsymmetrical circuits



resistance and high transconductance guarantees wide operating ranges and fast transition.

The optimum value of cathode resistor is somewhat lower than that corresponding to the knee of the conducting valve grid voltage. (c.f. Figs. 8, 9 and 10). This usually has an appreciably smaller value than the midpoint of the operating ranges presented in Figs. 2, 3, 4 and 5.

The resolving time[3] of these units depends, evidently, on the decay time* of the RC combination at the grids. Lower values of feedback and grid resistance are thus naturally needed for smaller resolving times. This increases the current drain through the feedback chain. The reverse voltage at the switching diodes increase in their turn as this is produced by the current through the anode resistor.

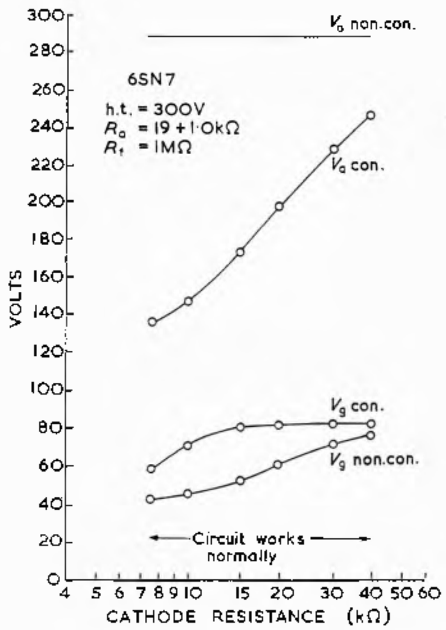


Fig. 8. Variation of anode and grid voltages, in both the conducting and non-conducting conditions, with cathode resistance in a binary using 6SN7

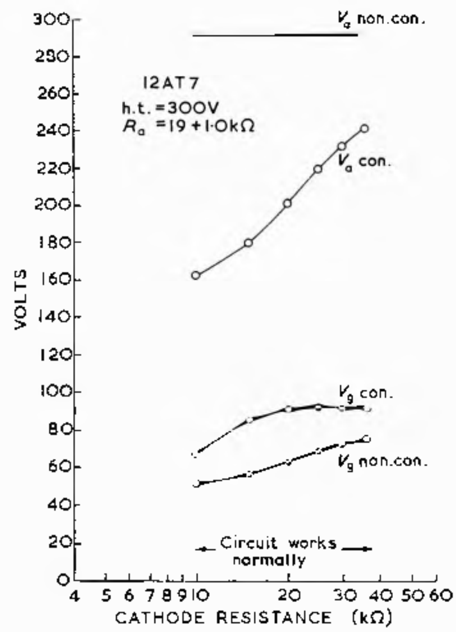


Fig. 10. Variation of the anode and grid voltages with cathode resistance in a binary using 12AT7

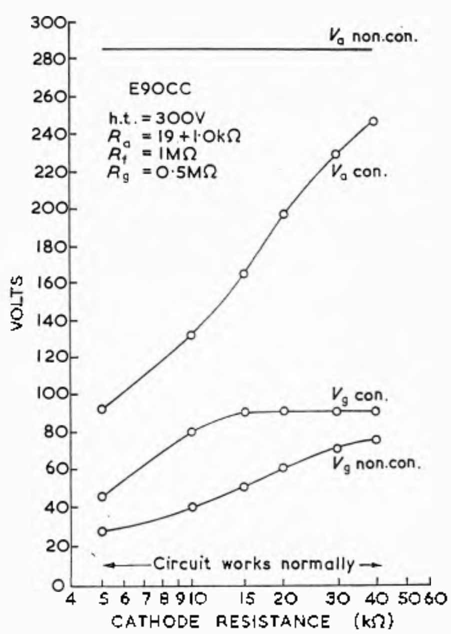


Fig. 9. Variation of the anode and grid voltages with cathode resistance in a binary using E90CC

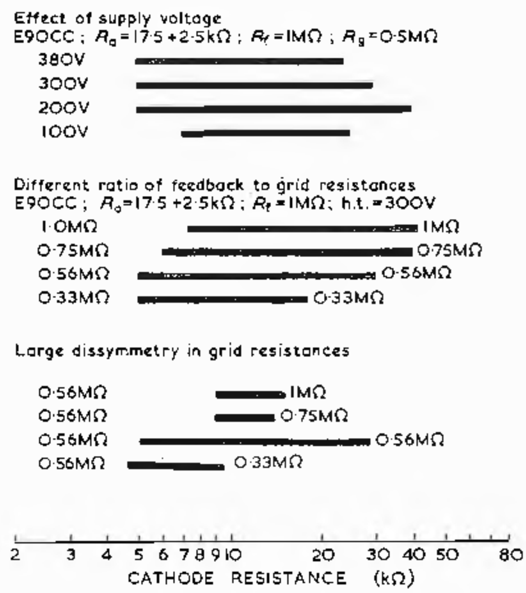


Fig. 11. Effect of h.t. on operating ranges (upper); with different ratios of feedback to grid resistors (middle); with large dissymmetries in the grid resistances; in the figure, left valve grid resistances are indicated at the left and right valve grid resistances at the right

It is wise to keep the value of the output resistance somewhat greater (by more than 20 per cent) than the minimum at which the units operate.

The data presented in Fig. 11 indicates that the ratio of feedback to grid resistance should have a value between 2 : 1 to 1 : 1. A ratio of perhaps 3 : 2 is more desirable than the more commonly used 2 : 1.

Better resolving time thus requires smaller values of anode resistance.

The range data presented in this article are for 1MΩ and 0·5MΩ values of feedback and grid resistance. For lower values in these positions, the minimum output

* *The switching times (c.f. Banerjee[4]) must be still smaller. The smallest switching times are obtained with anode resistance values of between 2 to 10kΩ.*

generator resistance needed will have greater magnitudes. The span of operating ranges obtainable is, however, not greatly affected.

Although the span of the operating ranges with anode resistance values of the order of 5kΩ, is quite large, units nevertheless become critical. The magnitude of the pulse injected by the previous unit, has to be critically adjusted (by adjustment of output generator resistance or the cathode resistance). Unless the need for the very best resolving times dictate such small values, these are to be avoided.

Finally, it may be pointed out that the circuit configuration utilized in Fig. 1 is the more desirable. The other configuration in which cathodes are grounded and the grid resistors returned to a negative supply voltage, possess the disadvantage that all units in a multi-unit scaler must utilize the same ratio of feedback to grid resistor values. Besides the span of the grid-cathode voltage for proper circuit operation is much smaller compared to the span of cathode resistors, and thus has to be held within a closer tolerance.

Acknowledgments

The authors are indebted to Prof. M. N. Saha, D.Sc., F.R.S., for his kind interest.

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CROWD—A Performance Recorder

Most factories find it necessary to keep continuous records of the operation of machinery, especially where the same machines are used for manufacturing a variety of products and where large numbers of different orders are dealt with in the course of a year.

In order to allocate costs, plan production programmes, and increase the time for which the machinery is available for useful work it is necessary to have such information as the time during which a machine is in operation, the time for which it is idle, the reasons for stoppages, and, sometimes, the number and type of articles made.

Some of this information can be obtained by using pen recorders, but the analysis of many yards of chart is tedious and time consuming.

If the chart is replaced by paper tape and an automatic perforator is used to punch holes representing the information, the record can be made in a form suitable for automatic analysis by an electronic computer or by a simple relay analyser. This technique eliminates the tedium of analysis but, because the tape is running all the time that the machine is in operation, long tapes have to be fed into the analyser and this takes time.

A punched tape recording system, originally developed for a study of crane movement, has recently been further developed by the British Iron and Steel Research Association and applied to similar time studies of rolling mills.

Designated CROWD (*C*entral *R*ecording *O*f *W*orks *D*ata) it summarizes the operation of the machine and produces a condensed record for analysis. The way in which it works is best described by referring to equipment designed for the specific purpose of recording the operation of a rolling mill. Three operating conditions had to be recorded:

(1) Running time (strip being rolled);
(2) Idle time (mill stationary during production, threading new coils or changing screwdown and guide settings);
(3) Stoppage time (mill out of production owing to roll change, failure of plant, or absence of operator).

Each order rolled by the mill is designated by an order number, against which running and idle time are allocated. Running and idle times are referred to as chargeable time. The duration of each stoppage is allocated to one of the following causes: roll-changing; mechanical fault; electrical fault; operational defect; meal break; late start; early finish.

The performance recorder has two acumulators, one of which records running time, the other idle time.

A clock incorporated in the equipment is arrange to provide electrical pulses at intervals of 3·6sec. These pulses are automatically sent either to the idle time accumulator or to the running time accumulator according to the position of a relay on the mill.

The operator has under his control a set of push-buttons for recording on the tape the code number of each order or the stoppage code number. He also has an 'incident button', and when this is pressed the data stored in the accumulators and on the push-buttons are punched in the tape and each accumulator and push-button station is reset to zero.

The sequence of events during the rolling of an order is as follows:

At the end of the previous order the operator will have pressed the 'incident' button, thus emptying the contents of the accumulators on to the tape. Idle time will then have begun to accumulate in the idle time accumulator. When the new order is received, the operator sets up the order number on the code push-buttons. As soon as rolling begins, the time pulses are switched automatically from the idle time accumulator to the running time accumulator.

When one complete coil has been rolled, the mill is stopped while the next is threaded. During this period the clock pulses are automatically switched back to the idle time accumulator. When the mill starts rolling again the time pulses are added to those already stored in the running time accumulator.

This sequence is repeated for each coil until the order is completed. The operator then presses the 'incident' button and both the total time for which the mill was actually rolling and the total idle time are removed from the accumulators and punched in the tape, together with the code number representing the order. The code number can be set up or changed at any time before pressing the 'incident' button.

If a fault should occur during the rolling of an order the operator presses the 'incident' button, thus emptying the contents of the two accumulators on to the tape. When the cause of the stoppage is discovered he sets up on the push-buttons the code number representing this particular fault. Once the fault is remedied he presses the 'incident' button again, whereupon the duration of the stoppage and its cause are then punched in the tape. When rolling continues, it is only necessary for the operator to set up the order number once more on the push-buttons.

The tapes can be analysed in several ways. As there are only ten different characters on the tape it can easily be read by eye if the equipment for automatically reading-out or printing-out the results is not available.

The data on the tapes can, however, be transcribed on to punched cards either automatically, by means of a tape-to-card convertor, or manually. The analysis can then be performed by punched card equipment. Alternatively, the data on the tape can be printed out by a standard teleprint page printer.

CROWD is now being manufactured commercially and an installation to record the operation of eighteen rolling mills is being built for a steelworks. This installation incorporates several additional features. It has a queueing device that makes it possible to record the operation of up to twenty-five machines with a single tape perforator. If the 'incident' buttons of more than one of the machines are pressed simultaneously, this queueing device causes the recorders concerned to read out in sequence. Another feature of this installation is a time switch which causes all the accumulated totals to be read out at the end of each shift, leaving the setting of the order number of fault number undisturbed. The time switch also automatically changes the shift number which is included in the recorded data.

THE I.E.A. EXHIBITION

A description, compiled from information supplied by the manufacturers, of selected exhibits to be shown at the Instruments, Electronics and Automation Exhibition at Olympia, from 7 to 17 May.

(Figures in parentheses refer to stand numbers)

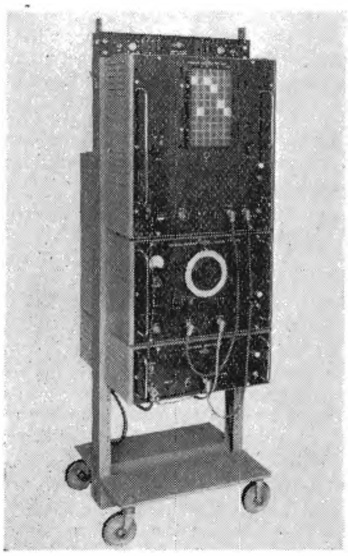
AIRMEC LTD (408)

High Wycombe, Buckinghamshire

TIME AND FREQUENCY MEASURING EQUIPMENT *(Illustrated below)*

The Airmec time and frequency measuring equipment will measure and monitor time intervals from 1μsec to 10 000sec and frequencies from 0 to 31 Mc/s with an extremely high degree of accuracy.

Time measurements are made by counting the number of standard microsecond signals that occur during the interval being measured, while frequency measurements are normally made by counting the



number of cycles of the input frequency that occur during a standard period of time. In both cases the results are recorded on an electronic counter possessing a clear and brilliant display.

The overall accuracy of the equipment is normally only limited by the accuracy of the frequency standard employed. For instance, if the oscillator type 213 is employed accuracies approaching 1 part in 10^7 can be obtained while even greater accuracy could be obtained with a better standard such as the oscillator type 218. Being of unit construction the equipment is flexible in operation and can be purchased in the most economical form for a particular application. For example, where a suitable source of standard frequencies is already available the standard oscillator unit may be omitted, or when frequencies up to 1·1Mc/s only require to be measured the standard frequency changer unit may be dispensed with.

All units are suitable for forward mounting on a standard 19in rack and can be supplied ready mounted on a trolley rack, as shown in the illustration.

ASSOCIATED AUTOMATION LTD (510)

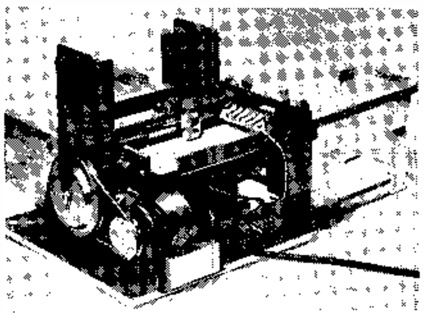
70, Dudden Hill Lane, Willesden, London, N.W.10

CARD READER *(Illustrated below)*

The high speed card reader has been designed to read punched cards column by column, photo-electrically, so that the information in the cards can be fed directly to an electronic computer.

The speed of the machine is such that ten complete cards may be read every second and the information entered into a computer at that speed.

In many applications it is required that a time interval be provided between the reading of one card and that of the next. For this purpose a special one-turn clutch is employed, actuated by a magnet under control of the computer itself, whereby the machine operation can be arrested



after the passage of each card and restarted on a signal from the computer.

When the starting signal is held on, the machine runs continuously, feeding cards at 600 a minute.

The cards are fed from a familiar type of magazine holding up to 650 cards and are discharged to a convenient receiver.

A knock-off is provided which stops the reader and signals the computer when the magazine runs out of cards.

AUTOMATIC COIL WINDER & ELECTRICAL EQUIPMENT CO LTD (942A)

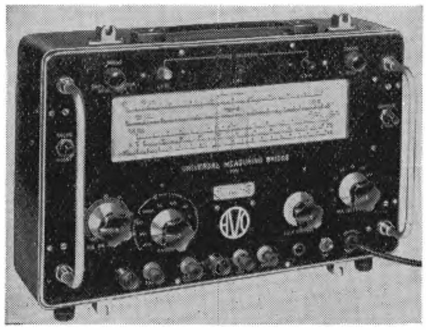
Avoset House, 92-96, Vauxhall Bridge Road, S.W.1

UNIVERSAL MEASURING BRIDGE *(Illustrated next column)*

This instrument is a self-contained mains driven model incorporating 24 calibrated ranges for the measurement of resistance, capacitance and inductance over an extremely wide range and to a high degree of discrimination. The value of the unknown is directly indicated on a clearly marked scale together with a suitable multiplier, the total effective scale length being approximately 50in for each type of measurement.

The instrument incorporates its own l.f. generator operating at 1kc/s and this feeds the bridge circuit for the measurement of inductance and capacitance, but resistance measurements are made by feeding d.c. to the bridge network. Balance indication is clearly shown by a panel meter operating in conjunction with a valve voltmeter circuit, this method of presentation being much more satisfactory than other visual or aural methods.

Provision has been made for measuring the leakage of capacitors and pre-



senting the leakage current in microamperes throughout a range of test voltages from 5 to 450V d.c.

Resistance Ranges. 6 calibrated ranges covering 0·1Ω-1 000MΩ (Accuracy ±1 per cent at mid-scale).

Capacitance. 6 calibrated ranges covering a nominal 1pF-1 000μF. (Accuracy ± 1 per cent at mid-scale).

Inductance. 6 calibrated ranges covering 1mH-1 000H. (Accuracy ±2 per cent at mid scale.)

ELECTRONIC ENGINEERING
and
ELECTRICAL ENERGY
will occupy Stand No. 206 where visitors will be welcome.

BRITISH THOMSON-HOUSTON CO. LTD (309)

Rugby, Warwickshire

Included in the B.T.H. exhibits will be several instruments for vibration measurement and analysis, these are as follows:

ROTOBALANCE ELECTRONIC BALANCING EQUIPMENT

This is for vibration analysis and the final balancing of all sizes of machines with the minimum of test running. Phase, amplitude, and velocity of vibration data is produced in readily-plotted form, whereby size and location of the balance weights required are clearly indicated. Produced in three forms for shaft speeds from 100 to 30 000 rev/min.

PORTABLE VIBRATION MEASURING EQUIPMENT

This provides vibration measurements

over a wide range of frequency and magnitude. It weighs only 32lb, and can be used on any standard a.c. supply. Measurements of vibration with indications of size and frequency of the principal components appear as meter readings and, at the same time, the vibration is displayed on an oscilloscope.

PORTABLE OPTICAL VIBRATION METER

This is for investigating the characteristics of vibration of surfaces and bodies under test. Light from an integral lamp reflected from the vibrating surface affects a photo-transistor, the output of which drives a transistor amplifier connected to headphones or an oscilloscope. The fundamental component frequency can be measured by the BTH frequency meter.

CATHODEON LTD (501)

Church Street, Cambridge

TELEVISION CAMERA TUBES (*Illustrated below*)

In addition to their range of conventional camera tubes Cathodeon will be showing the photo-conductive 'Staticon' which is widely used in industrial applica-

tions. Also exhibited will be a developmental model of an even smaller tube, the 'Minicon', which has a diameter of ½in and is about 3in long.

A. E. CAWKELL (716)

6-8 Victory Arcade, Southall, Middlesex

DOUBLE THREE-PHASE POWER SUPPLY

The SP50 double three-phase 400c/s power supply unit consists of two entirely separate three-phase channels operating from a common power supply. The equipment is operated from a three-phase precision 400c/s source and is capable of generating 150VA per phase to either a delta or star connected load. The output voltage will remain constant to within 4 per cent when the load is changed from zero to maximum. Numerous monitoring and safety devices are incorporated, and the equipment is designed to operate unattended continuously.

TIME CALIBRATOR (*Illustrated above*)

This instrument (type PE31CU) generates time markers in the range 0·5μsec to 1msec which may be free running, gated from an external gate, or gated from an internal gate having a continuously variable repetition rate of 1μsec to 10msec, and a continuously variable length of 2μsec to 100msec. A crystal controlled checker is fitted. The markers are available individually from sockets controlled by switches, or combined to give a 'timing comb' type of presentation, variable in height, and either positive or negative going.

The instrument may be used wherever a precision time calibrating source is required, and under certain conditions can be used to multiply or divide input pulses.

CROMPTON PARKINSON LTD (408)

Crompton House, Aldwych, London, W.C.2

MOTOR SPEED CONTROL

This electronic speed control equipment has been produced in collaboration with Airmec Ltd (who are exhibiting on the adjacent stand), and it will be arranged with a tachometer-generator and speed indicator to demonstrate continuous speed variation of a ½ h.p. motor, from the maximum speed of 2 800 rev/min down to standstill.

The control equipment is of the type that employs thyratron rectifiers and a hand-operated bias resistor to vary the thyratrons' outputs and, thereby, the motor armature voltage and the motor speed. It enables variable speed motors to be operated from single-phase a.c. mains and, in fact, as a result of the smoothing arrangements which have been incorporated in the rectifier circuit, standard d.c. motors can be used just as if they were operating from a d.c. supply.

The equipment makes provision for overload and no-volt protection with dynamic braking and reversing facilities. Among other features, the rectifier system is compactly designed and housed, and the latest construction techniques have been employed, such as the use of printed circuits and plug-in assembly for units containing small components.

DAWE INSTRUMENTS LTD (939)

99, Uxbridge Road, Ealing, London, W.5

ULTRASONIC CLEANING EQUIPMENT (*Illustrated above*)

Type 413 ultrasonic cleaning equipment embodies several new features, a more rugged construction and an improved cabinet design.

The pulsed operation at a frequency of about 40kc/s has been proved suitable for most applications and retained. The generator produces an output power of 125W mean and 500W peak.

Cylindrical tank transducers are available with barium titanate elements built into the base. The two sizes available are 5¼in diameter by 6in deep and 6¼in diameter by 7¾in deep. These have capacities of approximately half and one gallon respectively. Alternatively, immersible units permit the easy introduction of ultrasonic cleaning into existing baths. The immersible transducers measure 2¼in by 2⅝in by 9in overall. All assemblies are of stainless steel. Highly corrosive liquids may be used by placing them within an inner plastic or glass container. The output may be switched to either of two sockets enabling washing and swilling baths to be used alternately.

Cleaning occurs as a result of the rapid formation and collapse of vapour bubbles within the liquid This cavitation cleanses many assemblies difficult or impossible to treat by other means and in almost all cases ultrasonic cleaning affords a saving in plant or time over most methods.

VIBRATION INDICATOR (*Illustrated above*)

The type 1414 vibration indicator is a scaled down version of the type 1402 vibration meter, sufficiently small to be easily held in the hand.

The instrument consists of a Rochelle salt piezo-electric crystal accelerometer and a high gain amplifier which employs subminiature valves and components. The complete instrument measures only 8¾in by 3⅜in by 2⅞in and weighs approximately 2½lb.

Suitable integrating networks in the input of the amplifier enable measurements to be made directly in terms of the three vibration characteristics: displacement, velocity and acceleration.

Six full-scale meter ranges are provided for each function as follows:

Displacement: 0-0·001 to 0-0·3in zero to peak.

Velocity: 0-0·03 to 0-10in/sec peak.

Acceleration: 0-10 to 0-3 000in/sec^2 peak.

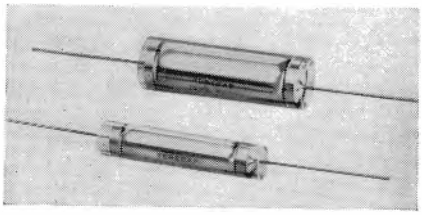
The gain of the amplifier is adjusted to match the sensitivity of the accelerometer supplied. Alternatively, a miniature barium titanate accelerometer can be used with a suitable correction factor to allow for the much lower pick-up sensitivity. An external cathode-follower unit is normally required for barium titanate accelerometers.

DUBILIER CONDENSER CO. (1925) LTD (908)

Ducon Works, Victoria Road, Acton, London, W.3.

TERYLENE FILM CAPACITORS (*Illustrated next column*)

A comprehensive range of capacitors and resistors will be shown including the



newly developed 'Terecap'; a metal cased tubular capacitor having a Terylene film dielectric which is particularly suitable where a high insulation resistance at high ambient temperatures is required.

ELCONTROL LTD (918)

10, Wyndham Place, London, W.1

PHOTO-SWITCHES

By making use of a recently introduced miniature photocell, Elcontrol are now producing a completely new range of photo-switches. Their most important characteristic is the small size of the photo-head, which is housed in a cylindrical aluminium fitting only 3in long by ¾in diameter. The associated relay unit is also considerably reduced in size, being approximately 7½in by 5¾in by 5in.

BATCH COUNTERS

Also on show will be a new fully developed batch counter making use of Dekatrons. The standard unit incorporates up to four counting stages and can be provided with a variety of input stages for picking up photoelectric signals, signals from external high speed switches, externally generated pulses, etc.

E.M.I. ELECTRONICS LTD (404)

Hayes, Middlesex

AUTOMATIC WEIGHING EQUIPMENT

An example will be featured on the stand in a working demonstration showing the application of these techniques to an automatic weighing system developed in co-operation with Western Manufacturing (Reading) Ltd.

The exhibit comprises a conveyor, a hopper and feed system for dry materials and a 20lb scale which is fitted with a graduated disk and photocells. These operate by means of a digital transducer and control automatically the filling of the containers so that the exact pre-determined weight of material is filled into each one. The actual weight of every filled container is recorded automatically by an electric typewriter.

An important feature of the system is that, while the tare weight of the container can also be recorded, it does not influence the weighing of the contents. The required net weight is determined solely by the setting of the dials and the control unit, and it is the actual weight delivered into the container that is measured.

The equipment is arranged to allow for two different feed rates, so that the initial filling is done quickly and the final stage slowly, giving a fine control of the approach to the point of equilibrium, especially with difficult materials.

Further developments of the system will allow a series of operations to be programmed on punched tape—for instance, for automatic mixing plants in the rubber and food industries.

The E.M.I. equipment is based on the use of digital transducers of various kinds to control the operation and electronic counters to record the quantities to any degree of accuracy within the limits of the basic equipment—for example, the scale of the weighing machine in this demonstration.

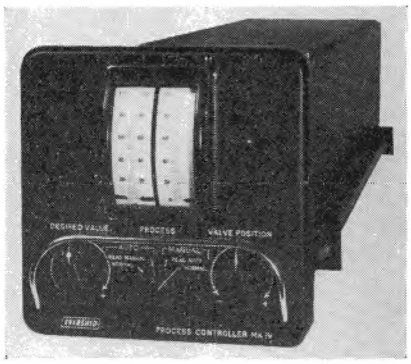
It follows, therefore, that this system is a general one for controlling or recording the position of a shaft. It need not use an optical digital transducer, nor is it restricted to weighing operations.

EVERSHED & VIGNOLES LTD (405)

Acton Lane Works, Chiswick, London, W.4

PROCESS CONTROLLER *(Illustrated below)*

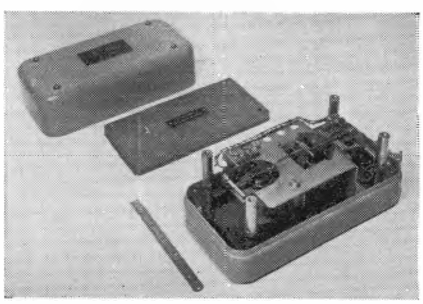
The process controller mark 4 offers considerable advances over the mark 3 version. It is completely electronic in nature and in the actual controller no moving parts are employed. The three control terms, proportional, integral and derivative are generated separately, and they are therefore free from interaction.



The new controller has simple d.c. circuits and the advantage offered by built-in test facilities. This new process controller operates in conjunction with a wide range of Evershed measurement transducers covering almost any physical parameter, but in principle any transducer providing a suitable electrical signal may be employed.

SIMPLE ANALOGUE COMPUTER *(Illustrated below)*

This new computer is a simple analogue computing device which will provide a continuous and instantaneous solution of algebraic equations occurring in the day to day operation of process controllers, boiler control schemes, etc.



The input information to the computer is an electrical signal which requires no programming or processing, and the output of suitable transducers, measuring devices or manually controlled signal sources may be introduced directly as variables in the input to the computer. Similarly the computer's output (i.e. the mathematical evaluation according to a predetermined formula of the variables and constants) can be employed directly as an operating signal for electro-mechanical control mechanisms, etc. It can also be displayed on indicating and recording instruments if required.

Thus the simple computer may be completely integrated into a process control scheme, evaluating continuously the variables and constants entering into the controlled process, and its output used to maintain the process in a desired manner. Alternatively, it may be used purely for the display of the solution of algebraic equations of dynamic problems.

The Evershed simple computer will be demonstrated in a typical application to a flowmetering problem.

GENERAL RADIOLOGICAL LTD (703)

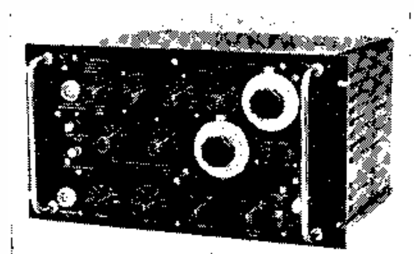
18, New Cavendish Street, London, W.1

WAVEFORM GENERATOR *(Illustrated below)*

This waveform generator is a precision instrument for laboratory or industrial use, intended for use in setting up pulse analysers, for use in the design of amplifiers and discriminators, and as a precision research tool.

The generator is built into a single unit having all the necessary controls easily accessible on the front panel. A trigger output pulse is provided, and monitor check points for the various d.c. levels are available on the front panel. It is also possible to monitor the currents which determine the output waveform amplitude.

This equipment provides rectangular pulses and 1/1 square waveforms of high amplitude accuracy for overall stability



and linearity measurements on pulse amplifier systems. The unit may also be used for direct tests on pulse amplitude analysers and as a source of absolute pulse amplitudes for general calibration work. The square waveform having a fast negative edge and a slow positive edge (or vice versa) when injected into the input of an amplifier via a small capacitance allows the charge signals from various types of particle detector to be simulated. For the appropriate injection capacitor this instrument reads directly in ion pairs. The provision of a 1μsec pre-pulse allows time-bases of a cathode-ray oscillograph to be triggered in advance of the leading edge of the output waveform so facilitating the checking of amplifier transient response.

GEORGE KENT LTD (403)

Luton, Bedfordshire

SYNCHRONOUS CONVERTOR (*Illustrated above*)

The Kent synchronous convertor is basically a single-pole double-throw switch that operates continuously at mains frequency. It is designed primarily for converting small direct currents into proportional alternating signals for easy drift-free amplification. It can also be used for the phase-sensitive rectification of small alternating currents, or for both purposes simultaneously (as, for example, in chopper-type d.c. amplifiers).

Action of the convertor is as follows. A soft-iron armature, carried on a beryllium-copper reed, is vibrated at mains frequency by the combined action of a permanent magnet and an energizing coil. Contacts attached to the reed engage with two others mounted on light flexible blades and pre-loaded inwards against adjustable stops. This arrangement eliminates contact bounce and enables the 'make' periods to be pre-adjusted independently to any setting. Four standard settings are available and the convertor is offered for operation on mains frequencies of 40, 50 and 60c/s, the construction differing only in the thickness of the central reed.

LAURENCE, SCOTT & ELECTROMOTORS LTD (104)

Gothic Works, Norwich

INSTRUMENT TYPE MACHINES

A range of tacho-generators, magslip servo motors, split-field and other instrument type machines will be shown. The tacho-generators are available in both industrial and instrument types, the latter having particularly low output ripple content and high output voltage per unit of speed.

AMPLIFIERS

Examples exhibited will include, a two channel amplifier for controlling a resolver producing sine and cosine voltages of very high accuracy.

A three-channel feedback amplifier for impedance changing and adding. Both this and the former unit may be supplied in hermetically sealed form with wiring at the front of the units, for use by the Armed Services.

An electronic servo amplifier controlling either a split-field motor or an exciter for industrial control applications. This amplifier has been designed for the highest reliability combined with ease of access. Since it is generally used in conjunction with large switchgear, its physical size is not of great importance.

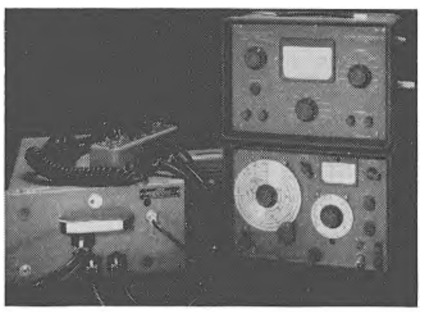
MARCONI INSTRUMENTS LTD (504)

St. Albans, Hertfordshire

V.H.F. MOBILE RADIO TEST EQUIPMENT (*Illustrated below*)

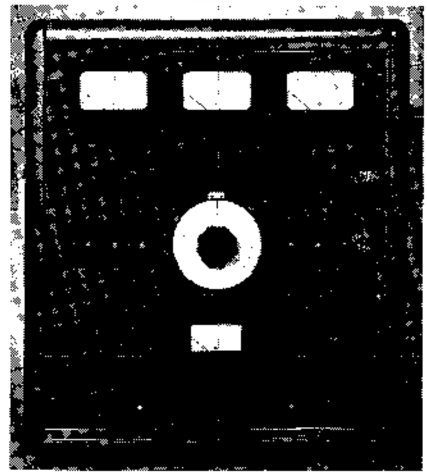
New instruments providing full facilities for testing v.h.f. mobile radio will be shown in public for the first time. The equipment comprises the v.h.f. signal generator TF1064 and transmitter and receiver output test set TF1065. Together these two instruments provide full facilities for testing f.m. and a.m. mobile transmitter/receiver equipments; their compactness and portability make them particularly suitable for use in the field. The signal generator covers 68 to 174Mc/s and 450 to 470Mc/s. An interesting feature of the test set is the inclusion of a transistor amplifier powered by internal mercury cells; this enables the instrument to preserve its complete independence from external power supplies.

The illustration shows the TF1064 and the TF1065 in use with the Marconi v.h.f., a.m., telephone equipment.



VALVE-VOLTMETER (*Illustrated above*)

A number of instruments for voltage measurement will be on view. The sensitive valve-voltmeter TF1100, which recently joined the range, is an amplifier-rectifier instrument for the measurement of a.c. voltages from 100μV to 300V in the frequency range 10c/s to 10Mc/s. It may also be used as a wide-band video amplifier.



MICROCELL LTD (707)

Imperial Buildings, 56, Kingsway, London, W.C.2

TRANSISTOR MEASURING SET (*Illustrated above*)

The transistor measuring set type 107 has been developed as a laboratory instrument capable of giving full information on the characteristics of pnp and npn transistors. Collector and base direct currents are separately and continuously variable over a wide range. Small signal measurements are made at 1kc/s and include h.t. network, open and closed parameters. Circuit configurations are either common emitter or common base.

Measurements are relatively simple, most being direct readings on a single calibrated dial.

MILLETT, LEVENS (ENGRAVERS) LTD (921)

Stirling Corner, Barnet By-Pass, Borehamwood, Hertfordshire

SPECIAL LAMINATIONS

The frequent requirements of those engaged in the production of engineering and electronic products is the supply of small or medium quantity shapes, where to date, it has been uneconomical due to the high tooling cost involved or great expenditure of labour in hand shaping. A technique has now been developed whereby transducer and transformer laminations together with special shaped articles can be produced economically without involving the production of press tools, by the 'Plasmet' technique.

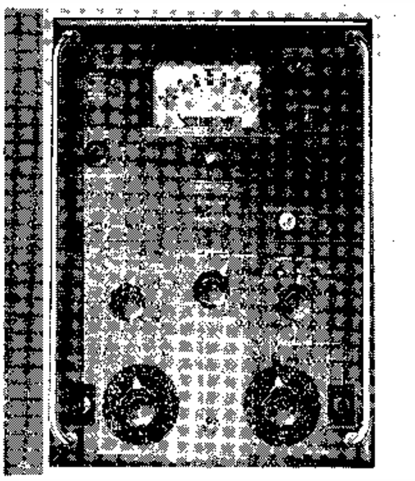
Examples of this technique are illustrated showing a transducer lamination produced from 0·005in thick nickel silver, a digitizor code disc in 0·005in stainless steel, air compressor valves in 0·005in stainless steel, and a condensor fan blade in 26 gauge phosphor bronze.

MUIRHEAD & CO LTD (901)

Beckenham, Kent

SYNCHRO TEST EQUIPMENT

This equipment has been specially developed for testing synchros to specification MIL-S-16892 on a quantity basis but to an accuracy usually associated only with laboratory measurements. It has facilities for measuring the electrical zero, electrical error (to ±0·5 minute), residual voltage, transformation ratio,



open-circuit power and current, and friction torque of CX, CT and CDX elements, and electrical accuracy and residual voltage or torque synchros.

PHASEMETER (Low Frequency) (*Illustrated above*)

This instrument (type D-729) gives direct indication of the phase angle (0-360°) and amplitude difference (up to 70dB) between two sinusoidal voltages in the frequency range 0·5c/s-10kc/s. Expansion of the degree scale to 20° for f.s.d. is provided. For measurements with severely distorted waveforms, provision is made for connecting a filter in the meter circuit (the D-788 low frequency analyser is ideal for this purpose) under these conditions, scale expansion to 2° for f.s.d. can be achieved. Uses include measurement of phase/amplitude characteristics of servo systems and active or passive networks, voltage and Q measurements.

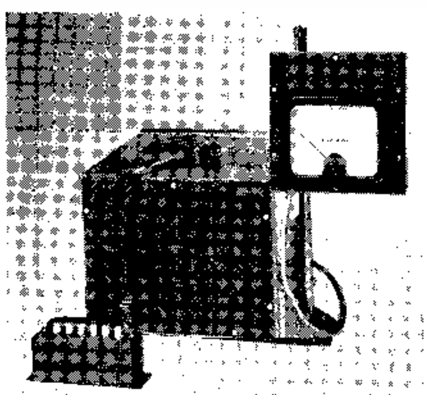
MURPHY RADIO LTD (703)

Welwyn Garden City, Hertfordshire

OHMMETER (*Illustrated below*)

This instrument has been designed and developed to facilitate rapid circuit testing on the production line.

It provides continuous coverage from 0 to 10MΩ in six ranges. The two low ranges from 0 to 100Ω and 0 to 1 000Ω are supplied from a low impedance voltage source of 0·25V, and have a common Set Zero control. The remaining four ranges, 0 to 10kΩ, 0 to 100kΩ, 0 to 1MΩ, and 0 to 10MΩ, are supplied by a full wave valve rectifier, the supply being 250V on the highest range. Individual



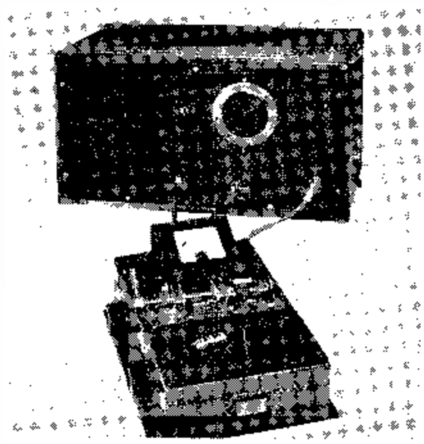
Set Zero controls are provided on these four ranges.

Range selection is carried out by means of a remotely controlled electro-magnetic switch unit; two types of switch control being available. The Ohmmeter "G," type "A" (No. 6125) has a six-way push-button unit contained in a metal box, connected to the ohmmeter by means of a multicore cable and six-way plug. This enables the control unit to be located in the most convenient position for the operator. The Type "B" (No. 6143) version operates through a bidirectional electro-magnetic switch controlled by push-buttons located in the test-prods. Connexion is again made through a six-way plug.

The indicator unit comprises a 5in by 6in rectangular moving-coil meter having a 4½in scale. It is secured to a chromium plated pillar by two clamps, which enables the height and angle of the meter to be adjusted between the limits of 6in and 17in coloured indicating lamps mounted above the meter show the range in use.

TURRET COIL TESTER (*Illustrated below*)

This equipment is designed for the production testing and adjustment of the



inductance of coils used in television turret tuners. Two similar types of instrument are available for testing either Band 1 or Band 3 coils.

The equipment comprises two units: a jig on which the coil to be adjusted is placed, and a main unit containing the major part of the valve circuits including the power supplies. The channel number and winding of the coil to be adjusted are selected by rotary switches on the side of the jig unit, and the percentage error between the winding under test and a standard winding, to which the jig has previously been adjusted, is displayed on a meter on the jig.

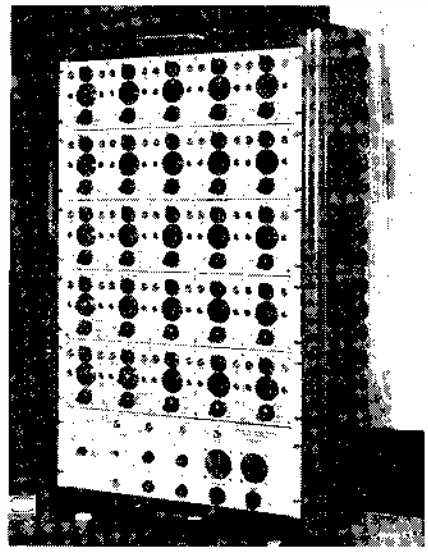
The coil is rigidly supported on the jig so that spacing between the turns of the winding under test may be adjusted to bring the inductance to within the required percentage error as indicated on the meter.

NAGARD LTD (216)

18, Avenue Road, Belmont, Surrey

PRECISION PROCESS TIMER (*Illustrated next column*)

This was designed to provide a means



of automatically and precisely time controlling a series of operations from a compact, accurate and reliable unit capable of adaptation to the widest possible conditions of industrial and scientific process control.

Basically, a master unit provides power supply and an electronic trigger control device which can be connected to any number of time interval controlling stages.

Started by an internal push-button, or remote contacts, these control units will then operate for pre-set time intervals in automatic sequence, or can be stopped at any desired stage to cycle back to the start, repeating the cycle automatically until stopped by the master switch.

Visual and audible signals can be set up to monitor the whole process, which, in the event of an external fault at any one stage, can be stopped and restarted either from the stopping place or from the beginning of the sequence.

Reliability and stable accuracy of a high degree has been achieved, giving an accuracy within 1 per cent for both timing and repetition functions, independent of supply mains variations between 200V and 250V.

A wide range of timing intervals can be provided, while the number of stages can be multiplied *ad infinitum*.

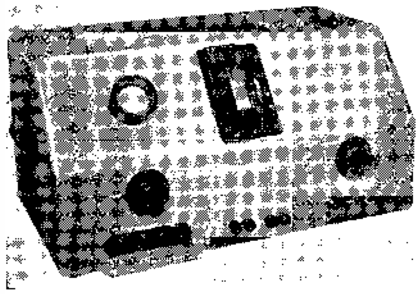
The example illustrated was designed for the control of a colour film developing process, and provides 25 timing stages for sequential or cyclic operation; 21 stages being variable up to 10 minutes each, and 4 stages covering 1 minute intervals.

W. G. PYE & CO LTD (501)

Granta Works, Cambridge

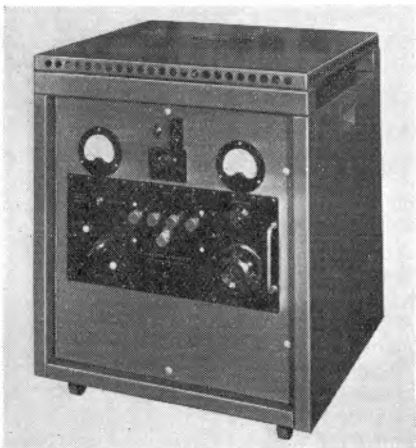
CONDUCTANCE BRIDGE (*Illustrated next column*)

Designed for general laboratory work, including analytical and quality control determinations, this bridge has a high accuracy and novel fool-proof method of balance indication by cathode-ray oscilloscope making the usual phase and amplitude balances largely independent. A built-in oscillator provides a signal at



a frequency appropriate to the range in use, the frequency being automatically selected by the range-changing switch. The use of the 'sealed second-cell' technique enables temperature compensation to be effected. The bridge can be used for direct conductivity determinations as well as for the comparisons of known and unknown samples.

A total range of 0·1 micromhos to 10 micromhos (i.e. 10MΩ resistance to 0·1Ω) is covered in four switched ranges. The bridge circuit comprises a slide wire and series arm. The accuracy is to ±0·2 per cent of reading. Discrimination is similarly to ±0·2 per cent of reading. These figures are for the worst conditions, i.e. at each end of the scale. Higher accuracy is obtained in the centre of the scale.



SERVOMEX CONTROLS LTD (900)

Crowborough Hill, Jarvis Brook, Sussex

D.C. Stabilizers (*Illustrated above*)

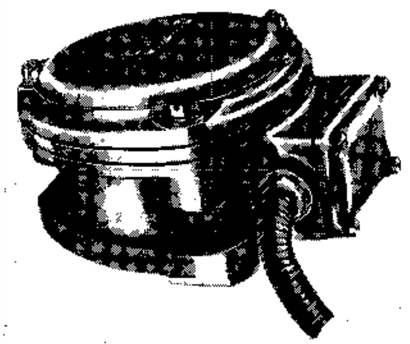
A new range of 1·2kW d.c. voltage stabilizers will be shown, illustrated is the DC 56. This instrument is for general laboratory use to take the place of accumulators with the advantage of continuously variable voltage and better long term stability. The response time is 100msec which although relatively slow is considerably faster than is usually available in units of this rating and this is combined with excellent ripple filtering.

SMITHS INDUSTRIAL INSTRUMENTS LTD (929)

Chronos Works, North Circular Road, London, N.W.2

Tachometer Generator (*Illustrated next column*)

The series 'M' generator is an all purpose heavy duty tachometer generator for remote indication of rotational speeds. It has been specially developed to meet the demand for a robust, waterproof, long life unit having physical dimensions for a very wide range of speeds and duties. This generator with its generous design and heavier construction is suitable for the arduous conditions encountered in such applications as rail traction, marine and heavy industry. It also gives a higher voltage output and can operate up to three heavy indicators if desired. Generators of this class are available for ranges varying between 0 to 500rev/min and 0 to 6 000rev/min.



Calibrated Relay

This calibrated relay has a very high accuracy, as much as ± 1 per cent under static conditions. It has three other good features, which are a very low power input, a high contact pressure, 15 to 60 grams, and finally contact wear does not affect accuracy of setting.

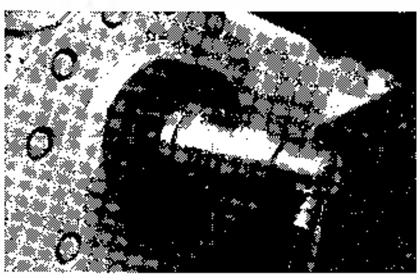
SOUTHERN INSTRUMENTS LTD (100)

Frimley Road, Camberley, Surrey

Torsional Vibration Pick-up (*Illustrated below*)

This is a seismic type of pick-up which can be used to measure the amplitude of torsional oscillation or cyclic variation at the free end of a shaft. It is particularly applicable to internal combustion engine crankshafts and camshafts, machine tool drives and gearboxes.

The pick-up contains a variable-inductance sensing element which detects the angular displacement between the fixing flange of the pick-up and a seismic inertia



mounted within the unit. The system contains neither fluid nor springs, a permanent magnet being used to supply both the damping and restoring forces. No rubbing sliprings are employed to couple the unit to its stationary connecting cable and it can be used on shafts which are merely oscillating as well as on those which are rotating continuously. The pick-up can be calibrated when mounted in its test position. Accurately known sustained angular displacements of the seismic inertia can be made by means of the micrometer fixture provided. This enables an overall static calibration of the recording system to be achieved.

The pick-up is designed for use with the Southern Instruments F.M. system. At its maximum amplitude of ± 2°, an output of approximately 10V peak is obtained from the MR220 F.M. pre-amplifier. The useful frequency range extends from approximately 5c/s upwards. The output of the pre-amplifier is directly proportional to amplitude of oscillation and may thus be applied to an oscillograph, analyser or valve-voltmeter to enable accurate measurements to be made throughout the frequency range, without the need for integrating stages or calibration curves.

STANDARD TELEPHONES & CABLES LTD (307, 308)

Connaught House, Aldwych, London, W.C.2

Digital Computer

The Stantec Zebra Electronic Digital Computer is a general purpose computer of medium size and speed. It is suitable for carrying out a large variety of the information processing problems required to be solved in research establishments and in industrial and commercial organizations generally.

The version of Stantec Zebra to be shown (including input and output equipment) is primarily a mathematical machine, and as such, the direct fields of application are likely to be found in laboratories where mechanical, electrical, structural, and aeronautical design and development work is carried out. The machine is also of great use in analytical work and in the compilation of mathematical and navigational tables.

In other versions Stantec Zebra forms the main computing component of a system of information processing devices designed to deal with problems of accountancy and finance, stock-keeping and production control, management and administration organization.

Magnetic Tape Storage Machine

This machine provides an infinitely extensible method of magnetic storage of information with a low random access time to any of the information.

Magnetic tape in an endless loop passes between a rotor and a stator and under a magnetic head assembly.

Stator and rotor are provided with compressed air and vacuum under the control of a special high-speed valve. Motion of the tape is started and stopped at high speed by moving it between stator and rotor, each using air pressure and suction alternately. Tape loops, up to 300ft in length, are carried in narrow flat containers which form a jack-in unit; all the capstan drive and magnetic head electrical and air connexions being on jack-in plugs and sockets.

Ten such units can be accommodated in a single frame mounting.

BOOK REVIEWS

Electrons, Waves and Messages

By John R. Pierce. 318 pp. 70 figs. Demy 8vo. Doubleday & Co., Inc., Publishers. 1956. Price 36s.

ONE cannot but regret that in the attempt to write a book of interest to both the intelligent laymen and to the man with a scientific background, the author has produced a book which is satisfactory to neither. It may be unfortunate, but it is nevertheless the fact, that modern developments in science are generally so highly specialized that their detailed intricacies are incomprehensible, except in the most general terms, to anyone who lacks at least some basic knowledge or experience in the particular field. To attempt, as the author has done, to explain to the laymen the significance of Maxwell's equations, the *raison d'être* of communication theory, the operation of a reflex klystron or that of a travelling wave tube is no easy task, yet in trying simultaneously to instruct the specialist, the author has failed to satisfy either.

With the object, one must assume, of interesting the widest possible circle of readers, the author has devoted much of Chapter I to elementary mathematics but, surely, even a 12-year old schoolboy does not need to be reminded that

"Division is always written as a fraction. Thus the statement that h is A divided by W is written

$$h = \frac{A}{W}$$

We would read this "h equals A over W" ...." (page 24). Chapter II deals with Newton's Laws of Motion in a manner appropriate to an intelligent student of 14 while Chapters III and IV attempt a popular explanation of electric and magnetic fields.

At the outset, it is true, the specialist reader is warned that "he will have to judge for himself what he should read and what he should skip"—but how does one know what to skip unless one reads it, and does any sensible reader enjoy paying a fair price for a book, at least half of which has to be skipped?

In Chapter VI we have a useful essay entitled "Maxwell's Wonderful Equations". This is one of the best explanations of the derivation and meaning of Maxwell's equations which your reviewer has seen and few radio engineers could read it without gaining from its contents. The non-technical reader will now, however, be doing the skipping for if he needed to be told in Chapter I that the area of a rectangle of height H and width w is written HW, he is fairly certain to boggle pretty seriously at the statement that

$$\phi\, H_t\, dl = \frac{\partial \phi}{\partial t} + I_c \text{ (page 108)}$$

Chapters VII and VIII deal in an elementary but informative manner with various forms of thermionic amplifiers and it is certain that the schoolboy who is "interested in radio" could read them with advantage. In Chapters IX and X we have essays on Noise and Radiation which cannot fail to be of value to the great majority of radio engineers and the pity is that the rest of the book is not written at the same level. Succeeding chapters deal with such subjects as "Picking up Television" "Communication Theory" and "Relativity", but while every chapter contains matter of interest, the informed reader continually experiences the sort of feeling which a student of the differential calculus would have if his lecturer interjected every few sentences with a reminder that a multiplied by b is written ab.

Interesting though much of this book most certainly is, it is very regrettable that the author attempted to write for too wide a circle. The science of electronics now covers so wide a field and each little corner is so highly specialized that one cannot hope to describe its broad expanse, even in a superficial manner, in language which is intelligible to the simple layman and, at the same time, of interest to the mathematician or the atomic physicist. By all means let books on science be written for the multitude, let rather more advanced books be written for the technical student but let us not make the mistake again of attempting to include within the covers of a single book matter which is appropriate almost evenly to readers of all ages from twelve to twenty-four and all classes from the plain layman to the qualified engineer or scientist.

G. R. M. GARRATT

Frequency Modulation Receivers

By J. D. Jones. 111 pp. 30 figs. Demy 8vo. Geo. Newnes Ltd. 1956. Price 17s. 6d.

THE principles of frequency modulation and its advantages are outlined and explained in the introduction and first two chapters of the book, together with a description of how an f.m. transmission may be produced. The succeeding chapters are devoted to a stage by stage discussion of the design procedure of a typical commercial broadcast receiver suitable for mass production, and examples of the numerical calculations involved are worked through. The methods used in the design laboratory for measuring and testing each stage, together with many practical hints are also given. A comprehensive survey of possible r.f. amplifier stages is made in Chapter III and the next two chapters are similarly devoted to the frequency changing stage, including oscillator frequency stability, and to the i.f. amplifier. Detectors for f.m. reception are explained in Chapter VI, and the design of a ratio detector circuit worked out. A typical complete receiver design is discussed in the final chapter. It is a pity that the text is marred in a few places by ambiguities of phrasing, which might cause confusion to the student, and there are a few misprints. However, these should cause little difficulty to the professional design engineer embarking on the development of an f.m. receiver, to whom this work is most likely to appeal, and information on current practice in this field is conveniently brought together in this volume.

H. N. GANT

V.H.F. Television Tuners

By D. H. Fisher. 136 pp. 50 figs. Demy 8vo. Heywood & Co. Ltd. 1956. Price 21s.

THIS is an excellent book for the television engineers who wish to know the theory, practice and particular problems of television tuners.

The author is obviously familiar with the subject and writes clearly and interestingly about it.

The book begins with the historical development of tuners starting in the U.S.A. and explains the basic requirements and performance criteria followed by a useful chapter on signal frequency amplification. In this chapter the problem of obtaining amplification at various frequencies is outlined, sources of noise are described and noise factor and factors determining it are covered. It is a very useful chapter with practical examples included.

Frequency changers or convertors are next described. This chapter contains a number of curves showing the performance under varying conditions of modern triode and pentode valves. As in the previous chapter the author discusses the very important aspect of noise and its contribution to the overall noise factor.

Various oscillators are next described and the components and conditions contributing to frequency drift are mentioned and tabled.

Chapter VIII is concerned with switch tuners and their practical form. Two complete circuits of such tuners are given and photographs of component parts are included. There are also photographs of turret tuners although these are the subject of the next chapter, where the advantages and disadvantages of the two types of tuner are also covered.

The book ends with useful chapters on tuner measurements, testing and servicing. The reviewer enjoyed reading this well illustrated book and benefited from it.

C. H. BANTHORPE

Circuit Theory and Design

By John L. Stewart. 480 pp. 100 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd., London. 1956. Price 76s.

THIS is, in the reviewer's opinion, the best book on circuit theory and design since the classical works of Guillemin, Shea and Bode. It presents, in a most stimulating and lucid manner, the analysis and synthesis of networks by relatively simple concepts of poles and zeros on a surface representing a function of a complex variable in the so-called p-plane. The principle adopted is as follows: any required frequency response of a transmission system can be described approximately by a polynomial expression, over a limited range of frequency. The response can be approximated even more closely by an expression which is the ratio of two polynomials and this more sophisticated expression is valid over a wider frequency range. Now each polynomial is uniquely defined by its roots, therefore the network or transmission system is uniquely defined by the poles and zeros corresponding to the roots of the denominator and numerator, respectively, of the response equation. The independent variable is, of course, the frequency "p" which can be a complex number, corresponding to transient conditions or purely imaginary, corresponding to steady state sinusoidal conditions. All of this is clearly explained in the book.

It then follows that many descriptive pole-zero diagrams can be written down by inspection for various transmission elements, including electronic amplifiers and servo-mechanisms and, what is most important, one can see how certain pole or zero positions must be altered to achieve stability or some other characteristic. This requirement can then be interpreted directly in terms of circuit or transmission element changes, the technique being the same whether purely electrical or electromechanical feed-back loops are involved.

The very clear and straightforward theoretical treatment is supplemented by many worked examples and numerous exercises and some important chapters have sections devoted to "Practical Matters" which ably demonstrates that the author speaks with authority based on considerable industrial experience.

It must be said, however, that the one or two chapters not dealing with the pole and zero method, are below the standard reached in the rest of the book, consequently the reader should not be discouraged by Chapter I, which is particularly disappointing with its cumbersome notation and rather hurried if not careless definitions. It is clear that the author was impatient to get on with the real and exciting task of p-z analysis, which he does so admirably.

W. R. HINTON

Advances in Electronics and Electron Physics.—Volume VIII

Edited by L. Morton. 562 pp. 40 figs. Demy 8vo. Academic Press Inc., New York. Academic Books Ltd., London. 1956. Price 105s.

THE contents of this, the eighth survey of its kind, range widely over the various fields termed "electronic", healthily ignoring current attempts to redefine "electronics" so as to exclude circuitry. Of the nine papers within its 562 pages, four concern instrumentation in atomic research, three deal with fundamental electron physics, and two with problems in tube development, though, of course, these categories overlap.

Two of the first group deal specifically with circuit techniques, E. Baldinger and W. Franzen reviewing methods of Amplitude and Time Measurement in Nuclear Physics, and J. L. W. Churchill and S. C. Curran discussing Pulse Amplitude Analysis. Interesting developments in the theory and applications of Field Emission are surveyed by W. P. Dyke and W. W. Dolan, and valuable reviews offered by Charles Süsskind on Electron Guns and Focusing for High-Density Electron Beams, and by E. C. Okress on Magnetron Mode Transitions. M. Knoll and B. Kazan discuss Viewing Storage Tubes, and G. H. Metson summarizes results of research on the Electrical Life of an Oxide-Cathode Receiving Tube—information gathered mainly by G.P.O. staff in the development of submerged repeaters.

The standard of the volume is high, though somewhat uneven, particularly in respect of the English style of its authors. (Incidentally, has American usage now turned "nickel" into "nickle"? (page 57, twice).)

The value is greatly enhanced by indexes of both authors and subjects, so often omitted by hard-pressed editors of reviews; but some ingenuity is still necessary in tracing information, for example, on transit-time spreads, which is not explicitly listed.

Among minor criticisms one might register objection to the practice of writing pulse rates as (e.g.) 10^4sec (page 326) rather than 10^4 per sec. on page 354 a resistor seems to be missing from the first bias circuit of Fig. 23. More generally, it seems regrettable that none of the various aspects of solid-state electronics could find a place this year. Progress here is so rapid that even if a major review has been recently offered, an "annual supplement" at least might be justified.

But this perhaps is asking too much. The present authors have well earned the gratitude of physicists and engineers for the pains they have taken to reduce our labour in keeping up to date.

D. M. MACKAY.

Foundations of Wireless

By M. G. Scroggie. 349 pp. 249 figs. Demy 8vo. Iliffe and Sons Ltd., 1957. Price 12s. 6d.

THIS book was first published in 1936. The basic theory of radio is covered, starting with elementary principles. No previous technical knowledge is assumed.

Apart from the fundamental laws of electricity and radio, the theory of valves, transmitters and all types of modern receivers are described. There is also an introduction to the techniques of television and radar. Aerials, power supplies and transmission lines are also dealt with.

Because the previous edition was entirely rewritten and illustrated with over 200 new diagrams, only minor modifications have been required to bring the present edition up to date, except for the addition of an entirely new chapter dealing with semi-conductors.

Transistors in Radio and Television

By M. S. Kiver. 324 pp. 80 figs. Demy 8vo. Chapman & Hall Ltd. 1956. Price 36s.

THIS book starts with a short chapter on "Modern Electron Theory." For those not very familiar with this very basis of solid state physics, it provides an easily assimilated simple introduction to the basic transistor action.

In the next two chapters the point contact and junction transistor are introduced and their different properties compared.

Chapter IV describes the principles underlying the operation of basic audio amplifiers while Chapters V and VI explain the operation of oscillators and typical radio receiver circuits. Chapter VII suggests possible applications of transistors in television sets but since this aspect of the use of transistors is still very fluid inevitably it only describes general problems and their general solution rather than detailed circuits.

Chapter VIII covers the newer types of semi-conductor device and briefly touches on photo diodes, pnip and tetrode transistors, diffusion techniques and the field effect transistor. All these new devices show promise in one way or another but until they are in general commercial use a lack of detailed information and experience on their characteristics prevents a more detailed approach.

Those readers who have had much experience in the servicing of valve circuits will particularly appreciate Chapter IX on the precautions necessary when servicing transistor circuits. The precautions listed are very necessary and the author goes into considerable detail. Chapter X lists a range of simple experiments that can be performed to familiarize oneself with transistor techniques in practice. English types of transistor are of course available over here to replace the U.S.A. types described in this chapter. The book concludes with several pages of bibliography and a table giving some details of a range of American transistors.

This book might be criticized on a number of minor grounds but three in particular will be noted here. Firstly, the circuits are drawn with power supplies placed in many different positions some with positive polarity at the top and some with negative, some with the battery horizontal. Some circuits even have two batteries in the same circuit with opposite polarities at the top. This indiscriminate positioning of batteries makes it much more difficult to trace the action of any particular electrode in a circuit and is most unfortunate in a book intended for students and newcomers to the subject. Secondly, there are one or two mistakes in the text, in particular the circuit on page 96 Fig. 3 shows the collector resistance $R2$ to be 47kΩ—surely this should be 4·7kΩ since with 15kΩ in the base the collector would otherwise be firmly "bottomed." Even this 15kΩ would need to be larger with modern high gain transistors.

Thirdly, probably because the book is intended for the student rather than for use in the design of actual circuits, practical details such as d.c. stabilization are often omitted.

Similarly, in Fig. 28 on page 115 no means is shown for adjusting the quiescent current of the output stage to minimise cross over distortion.

This book can be recommended for its claimed purpose of covering transistor applications for radio service men, radio amateurs and students. It is not of particular value to those who already keep at the forefront of the transistor art and look for new information or critical survey of existing literature.

Although it is perhaps a little too non-mathematical science masters may find it gives an easy way to introduce transistors into their curriculum and certainly anyone who is familiar with radio valves and has never yet met the transistor could do worse than start the new subject by reading this book.

G. GRIMSDELL.

Frequency Modulated Radio

By K. R. Sturley. 120 pp. 105 figs. Demy 8vo. Geo. Newnes Ltd. 1956. Price 15s.

THE general principles of f.m. and its advantages are discussed in the first chapter, the explanations being mostly carried out with the aid of vectors, diagrams and numerical examples, so that no fundamental knowledge is called for. Nevertheless, the fundamental concepts are adequately explained and should be readily understood by the practical man using and maintaining such equipment or by the amateur constructor, to which classes of reader this work is primarily addressed. A knowledge of the principles of an ordinary a.m. broadcast receiver is assumed. The second chapter discusses the production of f.m. signals and particularly the common modulation methods, with examples from current B.B.C. practice for broadcast transmitters. The third and largest chapter is devoted to the receiver for f.m. signals. Aerials, r.f. amplifiers and frequency changers for v.h.f. are discussed in detail. The function and design of limiters and discriminators are discussed, as is the ratio detector, the operation of which is fully explained and for which a suitable transformer design is given. In the next chapter the standard factory tests on a receiver are outlined, and clear instructions are given to enable the amateur to set up and align a receiver with the aid of only very limited equipment. A final chapter is devoted to a discussion of a typical commercial combined broadcast receiver for long and medium wave a.m. or v.h.f. f.m. reception. Like the famous ELECTRONIC ENGINEERING monograph on the subject by the same author of some years ago, of which one might consider this an enlarged and up-to-date version, the accent throughout the book is on the use of f.m. for v.h.f. broadcasting and very little mention is made of other applications. Only one self evident misprint was noted.

H. N. GANT

Computers—Their Operation and Application

By F. C. Berkeley and L. Wainwright. 366 pp. 49 figs. Demy 8vo. Chapman & Hall Ltd. 1956. Price 64s.

THIS book has been written by specialists familiar with the "art" of computional techniques, for engineers who are on the fringe of both the analogue and digital computer field.

It presents general information on the principles underlying different types of computers and specific information about computers marketed by Remington Rand Inc and IBM. There seems to be some unbalance here because 170 pages are devoted to covering the principles of both digital and analogue computers, but only 20 pages devoted to the specific machines UNIVAC and IBM 701/4.

Nevertheless these 170 pages make very interesting reading because the mathematics, so essential to the descriptions of machines that are in fact intended to handle mathematical problems, has been kept to a minimum and liberally backed-up by written explanations describing what the formula means.

The authors have attempted to coax the reader into the complexity of big machine logic and programme, by describing in very great detail a small simulator called SIMON.

In the reviewer's opinion this does not help the book or the reader to understand "Computer Logic" because so many tricks have been used to keep the equipment to a minimum (1·25cu.ft), that its logic only resembles its full-size brother and must surely confuse those struggling to grasp new concepts. The "Home Constructor", on the other hand, would certainly benefit once he had constructed the machine from the detailed diagrams provided for this purpose.

Although the reviewer submits a mild protest against some of the machines listed on page 161 under a heading "Computers Which are Neither Digital nor Analogue", and must finish up by paying the authors a compliment in saying that the descriptions of analogue principles are so lucid that even he, a confirmed digit-addict, could understand them.

W. WOODS-HILL

Television Explained

By W. E. Miller (Revised by E. A. W. Spreadbury), 184 pp. 107 figs. Demy 8vo. 6th Edition. Iliffe & Sons, Ltd. 1957. Price 12s. 6d.

OWING to the changes that have taken place in the design of television receivers since the publication of the fifth edition, mainly due to the introduction of alternative television programmes, it has been necessary, in the present edition, to re-write several of the chapters completely, thoroughly revise the remainder, and introduce a number of additional chapters in order to cover such new features as switched tuning, automatic gain control, the increased use of flywheel synchronization, and dual aerial systems.

LETTERS TO THE EDITOR

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

A Transistor Multivibrator Circuit

DEAR SIR,—Recently multivibrators using junction transistors have been described[1]. These circuits have usually resembled quite closely the vacuum valve multivibrators. The writer has found that, in some cases, a slight modification of this basic circuit can give a more nearly rectangular waveform, which is often desired.

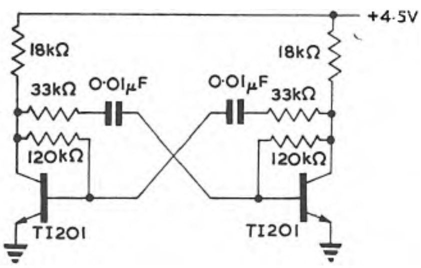


Fig. 1. The modified multivibrator circuit

Fig. 1 shows the circuit of this modified multivibrator. It differs from the other designs chiefly in having the 33kΩ resistors in series with the coupling capacitors. Similar series resistors have sometimes been added to valve multivibrators, again to give a more nearly rectangular waveform[2], but with the transistor circuit the improvement seems to be even more pronounced.

The 120kΩ resistors between collector and base carry the bias current. This biasing arrangement has been used frequently in amplifiers, since it provides some stabilization against changes in transistor parameters due to replacements, or to changes in temperature.

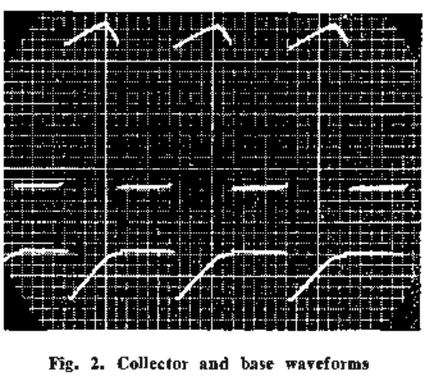


Fig. 2. Collector and base waveforms

Fig. 2 shows the collector and base waveforms of one transistor of the circuit of Fig. 1. Fig. 3 shows the waveforms when the 33kΩ resistors were shorted out. In both figures the frequency is about 1kc/s. The improvement in the waveform of Fig. 2 is apparent.

The reason why the series resistor improves the waveform seems to be as follows. If there is none, when one transistor is cut off, the voltage at its collector can rise only as quickly as the coupling capacitor can charge through the load resistor, the large charging current flowing through the low resistance at the base of the opposite transistor. This base current is much greater than is needed to bottom the transistor, and the excess does no particular good.

Now let the series resistors be added, so as just to pass enough current to bottom the transistor. Then, when one transistor is cut off, its collector will rise

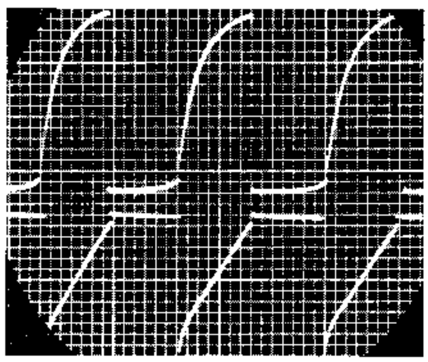


Fig. 3. Waveform with resistors short-circuited

immediately to a voltage determined by the ratio of the series resistor and the load resistor, and then will rise quite slowly, since the current charging the coupling gives a more nearly rectangular waveform, as was desired.

The above argument, of course assumes a frequency low enough that the input capacitance of the transistors is unimportant. At higher frequencies, the series resistors would give less improvement.

Yours faithfully,

H. L. ARMSTRONG,
National Research Council,
Ottawa.

REFERENCES

1. JACKET, A. E. Multivibrator Circuits using Junction Transistors. *Electronic Eng.* 28, 184, (1956).
2. TERMAN, F. E. *Radio Engineering*, third edition, p. 601. McGraw-Hill, 1947.

Applying Digital Computers

DEAR SIR,—In your January 'Commentary' you said: "Yet it does seem that the technique of producing these machines is perhaps outstripping the research into their economic application and that intensive research into the best use of the computer in industry and commerce would pay handsome dividends." As proof that I agree with you, may I say that some work in this direction is already contemplated in this University? From 15 to 19 April there was a short course on The Industrial Applications of Electronic Computers, which was organized jointly by the Institute for Engineering Production and the Department of Electrical Engineering of the University of Birmingham.

The difficulty about carrying out a 'research' in the engineering sense is the shortage of experimental data, since so few firms have made serious use of computers in industry. I am, however, keenly interested in the economic aspects of automation, and should be very glad to receive information on the economics of computer installation from either manufacturers or users of computers, with a view to assessing the profitability of applications. At present it seems that the field in which most evidence is likely to be available is the use of computers for clerical work.

Yours faithfully,

D. A. BELL,
Electrical Engineering Department,
University of Birmingham.

Erratum. The date of Reference 1 in the letter from W. V. Richings which appeared in the April issue (page 197) should read "(Dec. 22, 1944)" and not "(Dec. 22, 1957)".

Addendum. The last part of the letter from Mr. C. S. Evans and Mr. J. L. Carroll which appeared in the March issue (page 143), should be amended to read as follows:

"A comparison of the three references may be of value.

For a $4\frac{1}{2}$V reference the slope impedances are all about the same value (10-15Ω) and the temperature coefficients are:

(a) Zener diodes −·04 to +·01/°C[1]

(b) Batteries ·001/°C

(c) Accumulateur étanche −·002/°C[2]

Although batteries have the smallest temperature coefficient the accumulateur étanches are only slightly inferior in this respect have a much longer life and the variation of voltage with time under operating conditions is much smaller."

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1. SMITH, D. H. The Suitability of the Silicon Alloy Junction Diode as a Reference Standard in Regulated Metallic Rectifier Circuits. *Trans. Amer. Inst. Elect. Engrs.* 73, Pt. 1, 645 (1954).
2. SHEA, R. F. Principles of Transistor Circuits. (Chapman & Hall Ltd., 1953).

Short News Items

A Conference will form a part of the Instruments, Electronics and Automation Exhibition at Olympia. It will be held from 8-16 May within the exhibition building, and the field has been subdivided into a number of sections, a complete day being devoted to each subject. The programme includes an automation day, a nuclear day, a medical day, an industrial day, a computer day, and a communications and navigation day. The morning session in each case will be devoted to a general review of the subject, while more specific techniques and applications will be dealt with in the afternoon. The lectures and discussions will range from the control of nuclear reactors to the use of the artificial heart in surgery, and from the part which electronics will play in the office of tomorrow, to the use of automation in the cooking and tinning of foodstuffs. Further details may be obtained from Industrial Exhibitions Ltd, 9 Argyll Street, London, W.1.

The British Electrical and Allied Manufacturers' Association Education Committee is proposing to hold a one-day conference on education at the Royal Festival Hall, London, on 8 May. It will be for representatives of BEAMA member firms only, and will aim at assisting those members who have not had much experience of comprehensive schemes of training, but who are nevertheless concerned about ensuring that they are adequately staffed in the future.

Decca Radar Ltd have announced that, since October last when the True Motion Radar was introduced, their demonstration yacht 'Navigator' has been running continuously on the River Thames to meet the demands for 'live' demonstrations of this new marine radar. She has been visited by over 1 000 shipowners, marine superintendents, technical experts and Naval officers from the leading shipping nations in Europe, as well as the U.S.A., Canada and Japan. Some 200 orders have already been received, of which a large proportion are from overseas countries. The first three ships in the world to be fitted with production models of the Decca True Motion Radar are now at sea. They are the British m.v. 'Dartwood', the Norwegian m.v. 'Astrea' and the French t.s.s. 'Antilles'.

A National Conference on Aeronautical Electronics is to be held at the Dayton Biltmore Hotel, Dayton, Ohio, from 13-15 May.

Selig Electro-Magnetics Ltd of 296 Kilburn Lane, London, W.10, have been appointed sole agents in the United Kingdom for Eichhoff-Werke G.m.b.H. of Ludenscheid, Western Germany, whose works are equipped for all types of metal work, plastic moulding, winding and plating. They offer a service to the electrical, communication and electronic industries in Western Europe and North America by the production of custom-built components in short or long runs on a contract basis. Quotations are prepared from samples, drawings or customers' specifications.

The Metal Gravure Co Ltd of Babington Works, Knollys Road, Streatham, London, S.W.16, have announced the introduction of 'Adezograph'. It is the name given to an adhesive-backed name-plate or label which can be attached to plant, equipment or component where it is inconvenient to apply by normal methods using screws or rivets. Permanent adhesion has been achieved on all kinds of metals and paint finishes as well as on glass and vitreous enamels.

Transfer of the General Headquarters of The Institution of Electronics from London to Manchester. At a recent meeting of the General Council of the Institution of Electronics it was decided that the address of the general headquarters of the Institution of Electronics shall be 78 Shaw Road, Rochdale, Lancs. The former general headquarters at 20 Buckingham Street, Strand, London, W.C.2, is to be retained as the headquarters of the London Division of the Institution.

G. A. Stanley Palmer Ltd have announced that DEAC (Deutsche Edison-Akkumulatoren Co) Perma-Seal Hermetically Sealed Nickel Cadmium Accumulators are now available in two new battery types, namely the 5/900D and the 5/D1·5.

Mr. J. A. Smale has retired from the post of Engineer-in-Chief of Cable and Wireless Ltd and has become Technical Consultant in Telecommunications Engineering to Marconi's.

The Plessey Co Ltd have announced that Dr. Denis Taylor, head of the Electronics and Instrument Division at the Atomic Energy Research Establishment, Harwell, has been appointed a Director and General Manager of Plessey Nucleonics Ltd. In addition, he will become Research Executive of the Aircraft and Electronics Groups of the Plessey Co Ltd, and will be responsible for the co-ordination of the company's nuclear programme. Dr. Taylor will take up his new appointment on 1 May.

Mr. R. Arbib of Multicore Solders Ltd has been elected Chairman of the Radio and Electronic Component Manufacturers' Federation. Mr. K. G. Smith has been elected Vice-Chairman. Sir Robert Renwick, President of the Federation for ten years, has resigned. His successor is Major L. H. Peter, Chief Development Engineer, Westinghouse Brake & Signal Co Ltd.

Dr. Paul Eisler has been made an officer of the French "Order of Merit for Research and Invention" for his invention and pioneering work on printed circuits. The investiture took place recently in Paris.

The Northern Polytechnic, Department of Telecommunications, are organizing a course of ten lectures on High Quality Sound Reproduction, commencing on Monday, 13 May, at 6.30 p.m. The lectures will take place each Monday and Thursday evening for five weeks, with the exception of Whitsun week. The fee for the course is 21s. Further information may be obtained from the Head of the Department of Telecommunications, The Northern Polytechnic, Holloway, London, N.7.

An Electronic Computer Exhibition, to include data handling equipment of all kinds, is to be held at Olympia from 28 November to 4 December 1958. The exhibition will be the first of its kind to be held in Great Britain. It is being sponsored, at the suggestion of the National Research Development Corporation, by a joint committee of the Radio Communication and Electronic Engineering Association and the Office Appliance and Business Equipment Trades Association under the chairmanship of Mr. J. A. Cumming, chairman of the Exhibition Committee of the O.A.B.E.T.A. Concurrently with the exhibition there will be a symposium at which papers dealing with the applications of computers to problems in business, industry and science will be read and discussed. Mrs. S. S. Elliott, Exhibition Director of O.A.B.E.T.A., has been appointed organizer and Mr. J. F. Richardson, Technical Secretary of R.C.E.E.A., is secretary of the Committee. Enquiries should be addressed to him at the Radio Communication and Electronic Engineering Association, 11 Green Street, London W.1.

Marconi Instruments announce that an electronic demonstration van, "The Blue Goose", is being used on the U.S.A. West Coast to boost the sales of telecommunications test equipments made by Marconi Instruments Ltd. The van has just completed a two-month tour of Northern California and is now visiting Los Angeles electronic firms.