ELECTRONIC ENGINEERING

VOL. 28 No. 341

JULY 1956

Commentary

THE National Physical Laboratory Report for the year 1955 has recently been published and, while this gives an interesting overall picture of the year's activity at the Laboratory, including many applications of interest to the electronic engineer, a more specialized report, published concurrently, is of greater topical interest. This is entitled "Wage Accounting by Electronic Computer" and is of especial interest since it is believed to be the first report of its kind in the world. The report is the first product of a study group formed in 1954 and which consists of representatives of the Organization and Methods Division of H.M. Treasury, of the NPL (as computer experts) and the Ministry of Pensions and National Insurance (as potential users of computers).

The report is written to interest both accountants and computer engineers. It deals with the general case of payroll calculations and with a particular case which is analysed in great detail. It gives that type of knowledge and understanding needed by members of both professions in order that they can co-operate successfully in using these machines to the best possible advantage.

The scheme considered uses an existing high speed computer (DEUCE)* with some modifications; it calculates each worker's overtime, finds out how much income tax he should pay and deducts this and his National Insurance and any other deductions from his gross pay. It is also shown how a computer may be used not only for the computation of pay, but for all the marshalling, recording and manipulating of data which make up the work of a wages department.

The report shows that such a scheme can be economic using present-day payroll procedure and it estimates the costs when applied to various payrolls. It shows that the calculations for a staff of 3 400 paid weekly with overtime, but not piecework, can be done by the machine in one hour, but that 43 man-hours are required to feed and extract the information. The wages staff can, however, be reduced from 22 to 11. As in other forms of automation, the staff retained are fairly highly skilled—the machine merely does the routine work, but at a very great speed. This surely is a form of automation that will evoke little protest, for most clerical workers will agree that payroll computation is extremely tedious, while experience has already shown that existing office workers of good average intelligence can be trained readily to computer techniques—and there is, at the same time, a shortage of clerical workers.

The detailed figures estimating the cost of using a computer for this type of work are of considerable interest. The DEUCE, fitted with four magnetic tape units (two for input and two for output) in addition to its present punched card equipment, would be suitable for the weekly wage accounting for the 3 400 staff at the Central Office of the Ministry of Pensions and National Insurance. It would require a total staff of 11, most of them on a part-time basis, against 22 full-time staff used at present. If such a computer was employed on no other work, the annual cost, excluding overheads, of staff, computer and other equipment would be £16 200 against £12 700 by present methods. The computer could, however, also undertake the monthly payroll for 19 000 staff at the Central Office and, if further work could be found to keep the machine fully occupied, the pro-rata annual cost for the weekly wages work, again exclusive of overheads, would be only £2 700. In other words, to be economically justified, a computer, like most other large pieces of machinery, must be kept fully employed.

A computer such as DEUCE could, of course, be used for the weekly payroll as it stands, with punched card input and output only, but, due to the longer time it would take there would be less time available for other work and the cost of the payroll work would consequently be greater. This underlines one of the greatest limitations at present, that is that for this type of work existing input and output equipment is not fast enough to take advantage of the intrinsic speed of the computer. Considerable development of input and output equipment, particularly magnetic tape units, is, however, taking place both in this country and the U.S.A.

In general it would seem that where the number of employees is greater than about five thousand it would be worth installing a computer for pay accounting alone, although most types of organization have a considerable quantity of other work which could conveniently be programmed for a computer. There should also be considerable scope for computer service centres to serve the needs of the smaller firms where an individual installation would be uneconomic. In the case of densely industrialized areas, such as trading estates, this might take the form of a co-operative enterprise on the part of several medium sized firms, for it is clear that, although the development of equipment for this purpose and of techniques of application is only in its infancy, the potentialities of computers for business work are enormous.

* DEUCE (A New Digital Computer). *Electronic Engg. 27*, 179 (1955).

An Electronic Timing Unit

By N. B. Acred* and G. Bishop*

This article describes in some detail the design and construction of an equipment for measuring the time intervals encountered in determining the speed of an aircraft flying over a straight course of known length. The factors which influenced the design are discussed.

THE timing unit to be described was designed in collaboration with the Royal Aircraft Establishment, Farnborough, for the measurement of the time intervals encountered when determining the speed of an aircraft flying over a straight course of known length. The equipment was used in the British attempt on the world air speed record made on March 10 1956 when the Fairey Delta II achieved a speed of 1 132 mile/h. The place of the timing unit in the timing system used for the attempt has been described elsewhere[1,2], but the unit itself is of interest since it represents the application of present-day electronic timing techniques to a traditional method of determining speed.

Design Requirements

The primary requirements was that the timing unit should measure and display time intervals between 1 and 99·999sec with an accuracy of 1 part in 10^4 or 1msec, whichever is the greater. The time interval would be defined by the closing of contacts at the ends of two lines. One pair of contacts would provide a signal (the start signal) to initiate operation of the unit, the other pair would terminate the operation by a similar signal (the stop signal).

Owing to the length of the course and the disposition of the timing equipment one line would be much longer than the other. To eliminate the timing error caused by this, the unit incorporated a network which could be included in the shorter line to equate the delay characteristics.

Further requirements affected both the electrical and mechanical design of the equipment which would be used in the field, possibly under widely varying climatic conditions.

Electrical Requirements

The equipment had to be reliable, operate satisfactorily in ambient temperatures up to 55°C, operate satisfactorily after being subjected to ambient temperatures of the order of −30°C, and be simple to set up in the field. It also had to provide pulses to operate a Neostron tube used for aircraft camera shutter calibration purposes, and provide, at the end of each time interval measurement, a power pulse for operating a timing camera which would photographically record the displayed time interval.

Mechanical Requirements

Since the equipment might have to be transported a considerable distance, it had to be sufficiently robust to withstand the vibration and minor accidents liable to occur during such an operation, and at the same time be of a convenient size and weight for handling. It also had to be dustproof, capable of withstanding a wide range of ambient temperatures, and able to dissipate sufficient heat to minimize interval temperature rise.

Fig. 1. The complete timing unit

Design Principles

The timing unit employs the well-known principle of determining time by counting the number of cycles of a known frequency occurring during the interval defined by the start and stop signals. It was decided to use two interconnected chassis: one, the gate unit, carries the time-base generator and various switching, metering and testing facilities; the other, the display unit, carries the Dekatrons and other items connected with the displaying and photographic recording of the measured time interval.

Mechanical Design

The two chassis are fitted with standard 19in rack panels and they are mounted one above the other in a steel rack cabinet of robust construction (Fig. 1). Rubber sealing gaskets are fitted between the panels and the cabinet to prevent the ingress of dust. The cabinet has a large radiating surface and it was estimated that the internal temperature rise would be between 10 and 15°C. Forced draught cooling was therefore considered to be unnecessary and subsequent tests justified this conclusion. The low internal temperature rise can be ascribed mainly to the use of cold-cathode tubes and low current techniques. These factors, coupled with the use of components working well within their maximum ratings, also contribute to the reliability of the unit. The problem of vibration was overcome by care in the mounting of components on the chassis and by the use of supports for the chassis in addition to the normal panel securing screws.

Electrical Design

The methods adopted to meet the circuit design requirements will be apparent from the circuit description below.

Gate Unit

This is illustrated by the block diagram Fig. 2 and the circuit diagram Fig. 3 (a and b).

Crystal Oscillator

It was decided that the oscillator frequency should be 20kc/s since this is convenient for dividing to the time-base frequency and a quartz crystal of small physical dimen-

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sions can be used. In the interests of frequency stability, which is of great importance in an application of this nature, the oscillator drive voltage was amplitude stabilized and the crystal, which is in an evacuated glass envelope, was maintained at a constant temperature by a thermostatically controlled oven.

Since the timing unit may be subject to considerable vibration during transport to a site, there was a chance that the crystal envelope might be fractured with consequent destruction of the vacuum and an undesirable change in operating frequency. As it is totally enclosed in the oven the crystal cannot readily be inspected and other methods of determining the state of the vacuum had to be devised. The final design of the oscillator incorporates a testing facility which enables the state of the vacuum to be ascertained.

frequency lies well within the specified limits of accuracy. The actual operating temperature is approximately 75°C and is maintained within $\pm$ 0·5° by the thermostat. The long-term frequency stability of the oscillator is better than 20 parts in 10^6 and the short-term (8 hours) stability is better than 3 parts in 10^6.

Start and Stop Trigger Tubes

It has been mentioned previously that the functions of the timing unit are initiated and terminated by the operation of contacts at the ends of two lines. Both lines are normally charged to approximately -150V and electrical similarity is ensured by the inclusion of a network in the shorter line, suitable component values being chosen by the setting of the line length switch, S_1. The direction switch, S_2, determines the initiating station and in Fig. 3 the start signal is

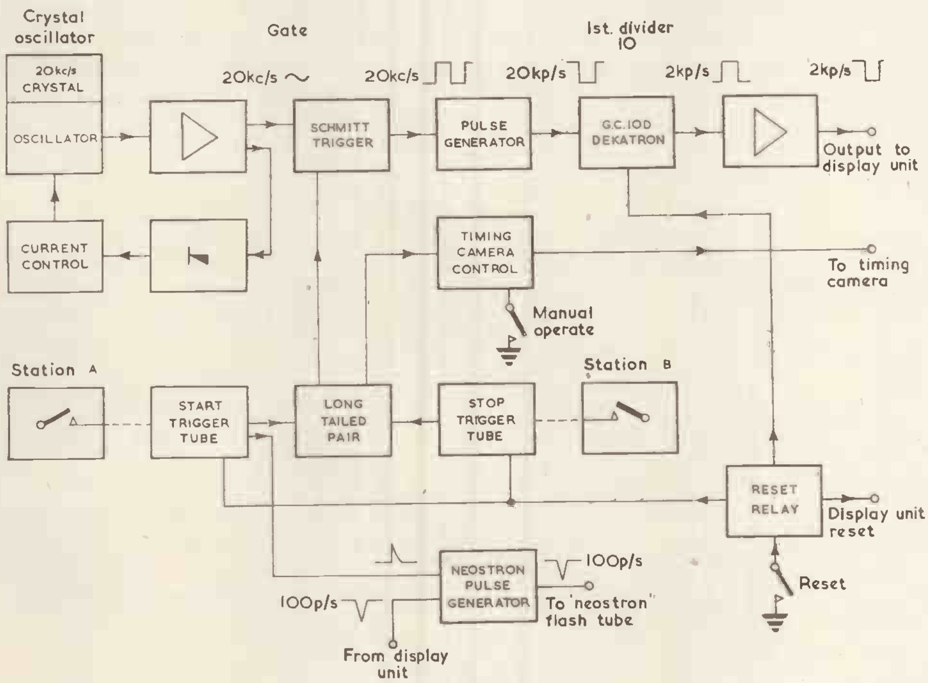


Fig. 2. Arrangement of gate unit

The crystal, XL_1, is maintained in oscillation in its series resonant mode by the cathode coupled amplifier V_1, and the 20kc/s output is fed via C_8 to the amplifier V_{3a}. The anode load of this valve is tapped to provide a voltage which is applied via C_9 to the diode-connected right-hand half of V_2, and after rectification, to the left-hand grid of V_2. This half of the valve controls the current through V_1 and any variation of output from V_{3a} will cause a variation of the current through V_1 and hence a variation of input to V_{3a}. The feedback loop is such that the output of V_{3a} tends to remain constant and amplitude stabilization of the crystal drive voltage is thus achieved. Changes in amplifier gain and variations of h.t. voltage have negligible effect on the oscillator.

The operating mutual conductance of V_1 is related to the effective series resistance of the crystal at resonance and since the current taken by V_1 gives an indication of the operating mutual conductance it will also indicate the value of the effective series resistance. This value increases if the crystal envelope vacuum is destroyed and the change is indicated by an increase in the current through V_1. This can be monitored on the meter, M_1.

During the initial setting up of the oscillator the crystal oven temperature is adjusted to ensure that the oscillator

obtained from station B via the distant line. This is connected via C_{15} and S_{2b} to the trigger electrode of the normally de-ionized start trigger tube V_4. Operation of the contacts at station B discharges the line and C_{41}, causing C_{15} to discharge into the trigger circuit via S_{2b}. The resultant voltage step, of approximately 150V, on the trigger electrode of V_4 causes the tube to ionize. The cathode potential then increases sufficiently to cause ionization of the panel mounted neon lamp, LP_1, which indicates that a start signal has been received. The voltage step, of approximately 100V, at the junction $R_{30}R_{31}$ is applied to the right-hand grid of the long-tailed pair, V_6.

The operation of the stop trigger tube, V_7, is similar to that of V_4, but in this case the neon lamp LP_2 ionizes to indicate receipt of a stop signal and the 150V step produced in the cathode circuit of the tube is applied to the left-hand grid of V_6.

Long Tailed Pair

This is formed by V_6 and its associated components. Under quiescent conditions the valve is near cut-off and only a very small current flows.

(a) *Start Pulse Effect*

The increase in grid potential caused by the ionization of

V_4 permits a current flow of approximately 2mA in the right-hand half of V_6. This current is applied to the Schmitt Trigger, V_5.

(b) *Stop Pulse Effect*

The voltage step produced by the ionization of V_7 causes a transfer of current to the left-hand half of V_6 owing to cathode-follower action and the right-hand half restores to the quiescent condition. Approximately 3mA flows in the left-hand anode circuit and this is sufficient to operate relay *A* via C_{12}. After operation of the relay the anode current flows via R_{35} and the relay restores after a delay which depends on the time-constant of relay *A* and C_{12}. The contact A_1 operates the timing camera via the slave relay *C*. When V_4 and V_7 are de-ionized by operation of the reset relay contact B_1, the +150V step is removed from the left-hand half of V_6 and both halves of the valve are restored to the quiescent condition. The tendency of relay *A* to operate as C_{12} recharges is counteracted by the shunting effect of the metal rectifier MR_1.

Schmitt Trigger

V_5 and its associated components form the Schmitt trigger. It is a bistable circuit which can only operate when the 2mA anode current is flowing in the right-hand half of V_6; it has two functions:

(a) To permit or prevent the application of the 20kc/s output from the oscillator to subsequent stages (V_{14}, V_8).

(b) When operating, to square up the sinusoidal output from V_{3a} before it is applied to V_{14}, V_8.

The left-hand grid of V_5 has a potential which varies sinusoidally owing to the superimposing of the output of V_{3a} on the d.c. grid potential. The right-hand grid has a d.c. potential derived from R_{42}, R_{43}, R_{46}, R_{54}, and the excursions on the right-hand grid are arranged to be within the potential limits of the left-hand grid. Thus when anode current flows in the right-hand half of V_6 it will flow in the right- and left-hand halves of V_5 alternately. The square wave output is taken from the right-hand anode circuit via C_{17}. The capacitor C_{16} is included to decrease the switching time of V_5 by discharging the stray capacitance associated with

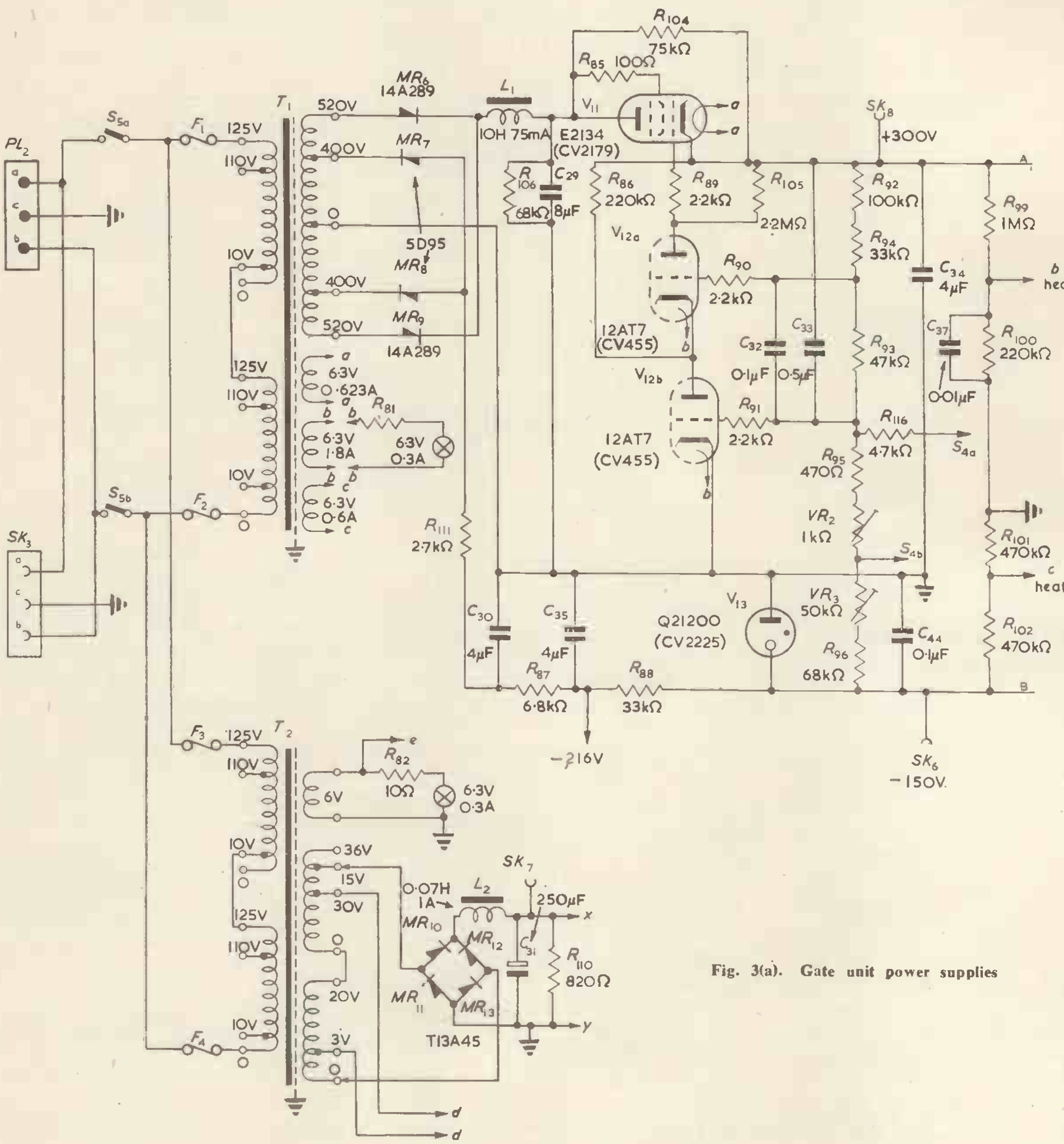


Fig. 3(a). Gate unit power supplies

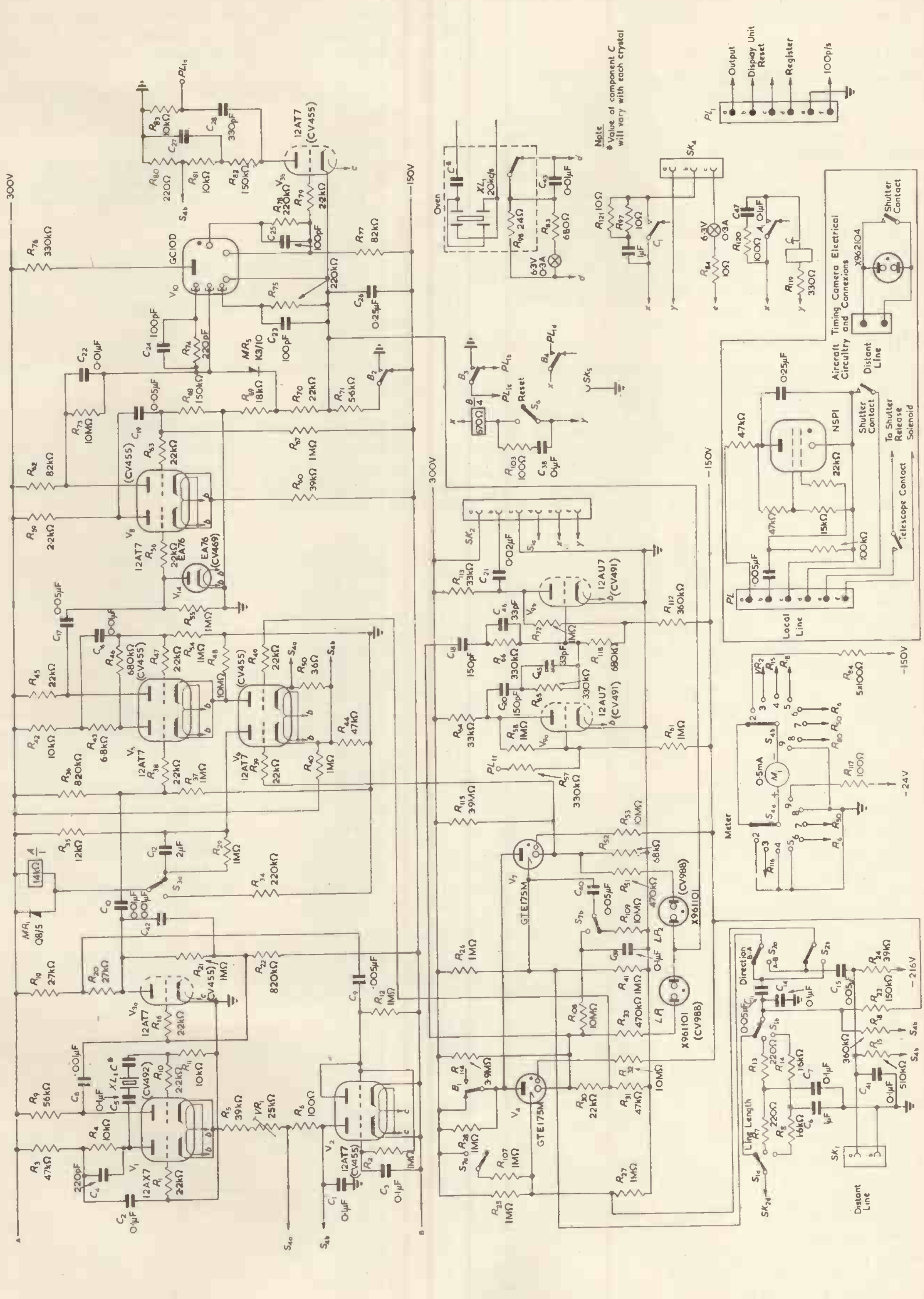


Fig. 3(b). The gate unit

the right-hand grid circuit through the low impedance anode load, thereby ensuring that the coupling to the grid circuit is of wide bandwidth and low impedance.

Pulse Generator

Although designated a pulse generator, the circuit of V_{14} and V_8 actually functions as a pulse shaper or slicer.

The input pulses are d.c. restored by V_{14} and applied to the cathode coupled amplifier V_8. This valve has positive feedback applied to the right-hand grid via C_{19} and the potential of this grid is such that, normally, no current flows in the right-hand anode circuit. The excursions of the left-hand grid potential are such that the right-hand half of the valve is pulsed into conduction by the diversion of the cathode current. The time taken for this diversion of current is reduced to a minimum by the positive feedback.

Output pulses are taken from the right-hand anode and applied to the next stage via C_{22}.

First Divider

The single pulse Dekatron, V_{10}, forms the divider and the operation of this type of tube is described elsewhere[3]. The input pulses are d.c. restored by MR_5 to the guide bias potential, a function which is assisted by a small positive bias applied via R_{73}. The output pulses from the Dekatron have a repetition rate of 2kp/s and they are fed to the switching valve V_{3b}. Pulses derived from the anode circuit of V_{3b} are differentiated by C_{28}, R_{83}, and fed to the display unit via PL_{1a}.

The meter, M_1, can be connected to the anode circuit of V_{3b} via S_{4a} and the integrating circuit R_{81}, C_{27}. The reading obtained gives an indication of the correct functioning of the divider, and unless a fault condition exists, one of the following readings will be obtained:

(a) Zero; when the Dekatron glow is stationary on a cathode other than the output cathode;

(b) 80 per cent of f.s.d.; when the Dekatron glow is stationary on the output cathode;

(c) 11 per cent of f.s.d.; when the Dekatron is dividing.

Neostron Pulse Generator

When V_4 is ionized by the start signal, the potential increase in its cathode circuit results in the application of a positive pulse to the grid of the partially conducting phase inverting amplifier V_{9b}. This produces in the anode circuit a negative going pulse which is applied to the Neostron tube, NSP_1, via C_{21} and SK_{2b}. After 10msec, and thereafter at 10msec intervals, a negative-going pulse is received from the display unit via PL_{1f}. These pulses are phase inverted by V_{9a} and applied to the grid of V_{9b} via the differentiating circuit C_{20}, R_{65}. Thus, during the timed interval, pulses are fed to the Neostron tube every 10msec, the first pulse being coincident with the start signal.

A typical self-quench Neostron tube circuit is shown. The tube is mounted in the local aircraft camera and its flashes are recorded on the photographic plate in the form of a series of small squares which can be used for shutter calibration purposes.

Timing Camera

The operation of relay A and its slave relay C has been mentioned previously. The power pulse necessary to operate the timing camera is derived from the rectified output of T_2 applied to the camera solenoid via the contacts C_1 and SK_{4b}.

Should more than one photograph of the displayed time interval be required, the timing camera can be actuated by manual operation of the switch S_3. This operates relay A via R_{34}.

The indicator lamp LP_4 glows when there is no film in the camera.

Test Circuit

The switch S_7 is provided to enable the functions of the timing unit to be checked. Two operations of the switch are necessary to simulate, respectively, the start and stop signals.

First Operation

S_{7a} puts R_{107} in parallel with R_{25}, thereby raising the trigger potential of V_4 to a value high enough to cause i nization. The sequence of events previously described then ensues.

S_{7b} has no effect, but it should be noted that the increase in cathode potential when V_4 is ionized charges C_{39} via R_{108}.

Second Operation

S_{7a} has no effect since V_4 is already ionized.

S_{7b} connects C_{40} to C_{39} which then discharges. The consequent pulse on the trigger electrode of V_7 causes ionization of the tube and the sequence of events previously described then ensues.

Reset Circuit

After the measurement of a time interval both the gate unit and the display unit are prepared for the next measurement by operation of the reset switch S_6. This operates the reset relay B whose contacts perform the following functions:

Contact B_1 includes R_{28} in the anode circuit of V_4 and V_7, reducing the anode current to a value below that necessary to maintain ionization.

Contact B_2 raises the guides and cathodes of V_{10} to earth potential. The output cathode, however, remains connected to the -150V rail and therefore the glow invests the output cathode. Contact B_2 also raises the potential at the junction LP_1, LP_2 and this, coupled with the decrease in the cathode potential of V_4 and V_7, deionizes the indicator lamps.

Contact B_3 resets the display unit Dekatrons to zero in a manner similar to that described for V_{10}.

Contact B_4 energizes the display unit register causing it to step on one digit.

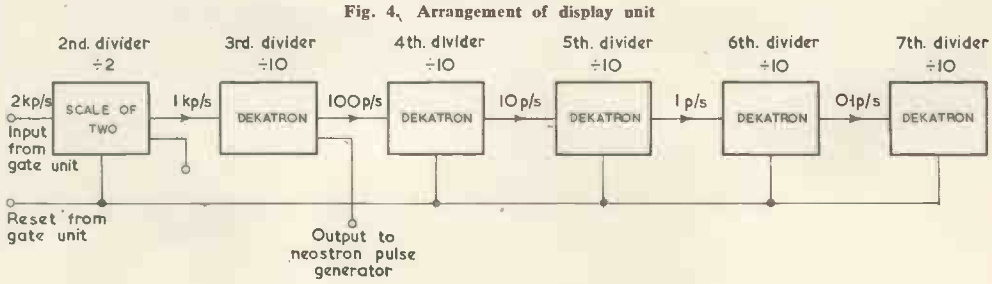
Power Supplies

As mentioned previously, power supplies for the timing camera are derived from T_2. All other supplies are derived from T_1 via circuits designed on conventional lines.

Display Unit

This is illustrated by the block diagram Fig. 4 and the circuit diagram Fig. 5.

Fig. 4. Arrangement of display unit



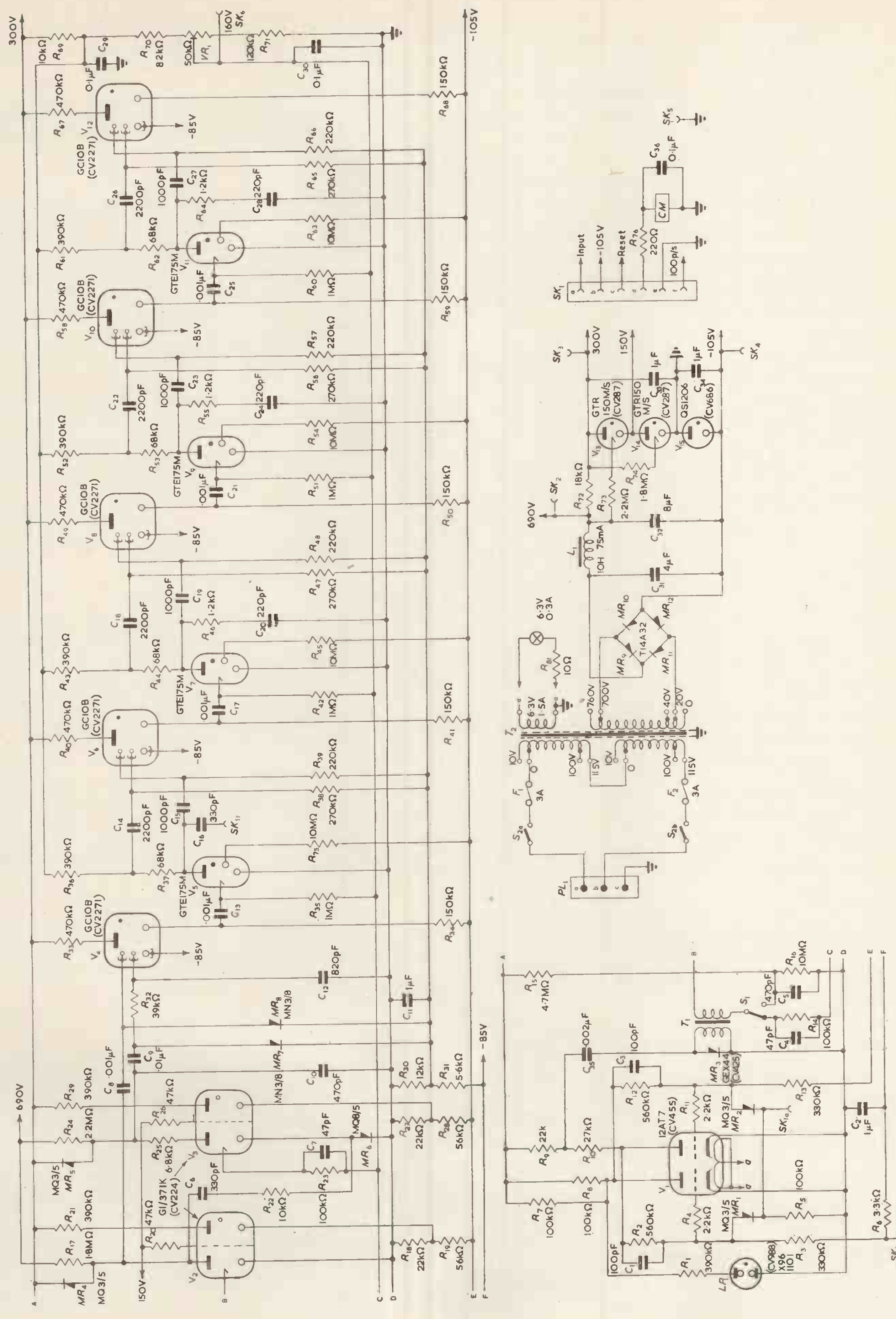


Fig. 5. The display unit

With the exception of V_1, only cold-cathode tubes are used in the unit. The high reliability and long life of these, as compared with corresponding thermionic types, is well known and contributes a great deal to the overall reliability of the timing unit. An additional factor of great importance in this particular application is the negligible amount of heat generated by the cold-cathode tubes.

SCALE-OF-TWO (2nd Divider)

The Eccles-Jordan circuit of V_1 forms the divider and a neon lamp, LP_1, is included to indicate the particular residing state at the end of the measured time interval.

Negative going pulses from the gate unit are fed via SK_{1a} and the coupling diodes MR_1, MR_2 to the grids of V_1. Output pulses, of 1kp/s repetition rate, are taken from a tapping on the right-hand anode load and applied to the phase inverting pulse transformer, T_1, via C_{35}.

THIRD DIVIDER

The double pulse Dekatron, V_4, forms the divider and it is coupled to the preceding stage via the cold-cathode trigger tubes V_2 and V_3.

The 160V trigger bias is applied to V_2 via R_{14}, S_1 and the secondary of T_1. An incoming pulse raises the trigger potential to a value high enough to cause ionization. The anode potential of V_2 falls to the maintaining value and, after the discharge of C_6 and the stray capacitance associated with the circuit, the current through R_{17} is insufficient to maintain ionization. The anode potential then commences to rise towards 690V, but is caught at 300V by the action of MR_4. This results in a substantially linear recharge of the circuit capacitance and the waveform at the anode has a steep fall and a linear rise. The initial steep fall is applied via C_6 and R_{22} to the cathode of V_3. This tube then ionizes and the cathode current flows through the low forward resistance of MR_6 after quickly discharging C_6 to chassis potential. V_3 deionizes in the manner described for V_2 and the waveform at the anode is similar to that at V_2 anode. The two waveforms are applied to the first and second guides of V_4 via C_8 and C_9 respectively. The diodes MR_7MR_8 d.c. restore the voltage waveforms to the guide bias potential. The function of V_4 is conventional and is described elsewhere[4]. Both V_2 and V_3 have auxiliary electrodes which, since they are under continuous discharge, stabilize the operation of the trigger triode portion of the tube. Interaction between the two sets of elecrodes is prevented by the screen.

FOURTH DIVIDER

The double pulse Dekatron V_6 forms the divider and it is coupled to the third divider by a cold-cathode trigger tetrode circuit employing a principle which has been described[5]. The capacitor C_{16} is provided to enable 10msec pulses (100p/s) to be taken from the anode circuit of V_5 and applied to the Neostron pulse generator in the gate unit via SK_{1f}.

SUBSEQUENT DIVIDERS

These are similar to the fourth divider except that there is no component similar to C_{16} in the anode circuits of the coupling tubes.

TEST CIRCUIT

The switch, S_1, is provided to enable the display unit operation to be checked independently of the gate unit. Operation of S_1 connects R_{15}, R_{16}, and C_5 to the trigger electrode of V_2. The potential applied is such that the tube oscillates at a frequency which depends on the time-constant of the trigger circuit and the difference between the striking and extinguishing potentials of the trigger. The resultant pulses in the anode circuits of V_2 and V_3 are applied to the Dekatron as described above.

RUN NUMBER METER

The meter, CM, is a P.O. electro-mechanical register which derives its energizing voltage from the gate unit via the reset relay contacts B_4. The meter steps on one digit for each operation of the reset switch and gives an indication of the number of time interval measurements made by the timing unit.

POWER SUPPLY

This employs gas-filled diodes as voltage stabilizers and is designed on conventional lines.

Conclusion

Both electrical and mechanical requirements were met by a prototype equipment which was subjected to searching tests at the Royal Aircraft Establishment. The results of these tests fully justified the measures adopted to meet the various requirements. Two other models were used during the record attempt and they proved to be entirely satisfactory under field conditions.

Acknowledgments

The authors are indebted to the Ministry of Supply and Dr. J. H. Mitchell, Controller of Research, Ericsson Telephones Limited, for permission to publish this article. They also wish to thank the Director and Staff of the Royal Aircraft Establishment, Farnborough, with whom the timing unit was developed.

REFERENCES

1. ACRED, N. B., BISHOP, G. An Electronic Timing Equipment for World Air Speed Record Attempts *Brit. Commun. and Electronics 3*, 288 (1956).
2. HILL, N. E. G. (To be published.)
3. ACTON, J. R. The Single-Pulse Dekatron. *Electronic Engng. 24*, 48 (1952).
4. BACON, R. C., POLLARD, J. R. The Dekatron. *Electronic Engng. 22*, 173 (1950).
5. *British Patent 712175.*

A New Valve Factory

A new factory for the manufacture of miniature and sub-miniature valves has recently been opened at South Ruislip, Middlesex by Hivac Ltd., the electronic division of the Automatic Telephone and Electric Co. Ltd.

Producing the very small electrodes and electrode assemblies involves such extreme mechanical control and absolute chemical cleanliness that every possible precaution has to be taken to maintain spotless working conditions.

The atmospheric pressure throughout the factory's air-conditioned assembly shop is kept above normal to ensure that no dust can enter from outside, and assemblers are provided with nylon headscarves and white overalls.

To dispense with dirt-collecting stanchions and roof girders, the barrel-vault type of construction has been adopted.

A unique feature in the factory's planning enables all electrical, gas, water and other essential supplies to be piped into the main assembly section from a specially constructed basement. The vacuum pumps are also housed below ground so that contaminating oil fumes and dust-harbouring pipes and wires are eliminated from the assembly shop. This enables regular maintenance to be carried out more easily and, at the same time, ensures the maximum freedom from distracting noise.

This type of construction also ensures that the greatest possible use is made of the floor space available in the air-conditioned part of the factory by enabling many operations not requiring a highly purified atmosphere to be carried out below ground.

Although every precaution it taken to keep the assembly area at the highest degree of cleanliness, so stringent have standards become that a special area has been erected and partitioned from the main assembly shop with individual pressurized benches for the assembly of close tolerance miniature and sub-miniature valves.

Electronic Methods of Analogue Multiplication

By Z. Czajkowski*, B.Sc.

(Part 1)

This article is a general survey of the principles used in analogue multiplication. An attempt has been made to classify the different methods of approach to the problem and to compare all systems as to accuracy, speed and complexity.

THE rapid scientific and technological progress of the last decade brought with it an enormous expansion of the field of electronic computation. Analogue computers play an important role in many fields of design and research. Special purpose machines have been also built to simulate aircraft, guided missiles, submarines and industrial plants. These are used both to study the behaviour of such systems and to train the personnel operating them without the use of the simulated aircraft, etc., which are usually very costly to operate. An important field of activity which is just opening for analogue computers is that of automatic process control.

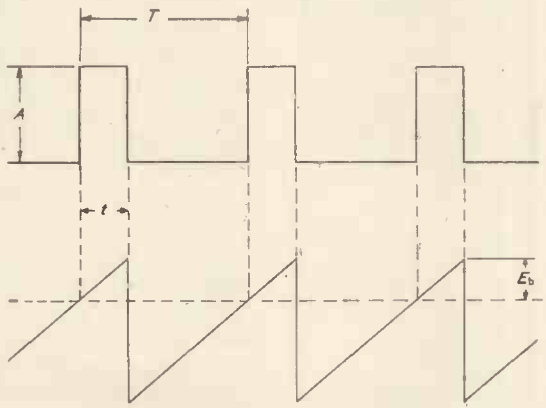


Fig. 1. Waveforms in pulse modulation

Of the six basic mathematical operations performed by computers the methods of adding, subtracting, differentiating and integrating have been known for a long time. A high gain negative feedback amplifier forms a standard part of any analogue computer. These amplifiers can be designed to give good accuracy, speed of response and stability. The servo multipliers can match the accuracy and stability of the functional amplifiers, but usually lack the speed of response, which makes them the slowest link in a computing machine.

The need for a multiplier device which would be fast, stable and simple has stirred the imaginations of many engineers and scientists. As this survey shows, all branches of electronic engineering have been called upon and techniques as different as radar, multi-channel telephony, cathode-ray oscillography and strain gauges have been used. No doubt, transistors and dielectric amplifiers will, in turn, make some contribution.

This article is divided into two parts. Part 1 deals with those devices which perform multiplication directly by application of some mathematical or physical formula. These instruments depend directly on the fact that the elements used follow exactly the law desired. Part 2 deals with those systems of multiplication which involve the principle of negative feedback in one form or another.

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The Time Modulation Multiplier

The principle of time modulation forms the basis of a large variety of multipliers. Several different systems are known and will be described in some detail. They all depend on finding the average value of a complex waveform, the amplitude, shape and repetition frequency of which may all vary. Fig. 1 gives the principle of rectangular waveform modulation. Pulse repetition frequency (f) is kept constant, i.e. $f = (1/T)$, where

A – is the amplitude of the pulse

t – the duration of one pulse

$t/T = K$ = coefficient of attenuation.

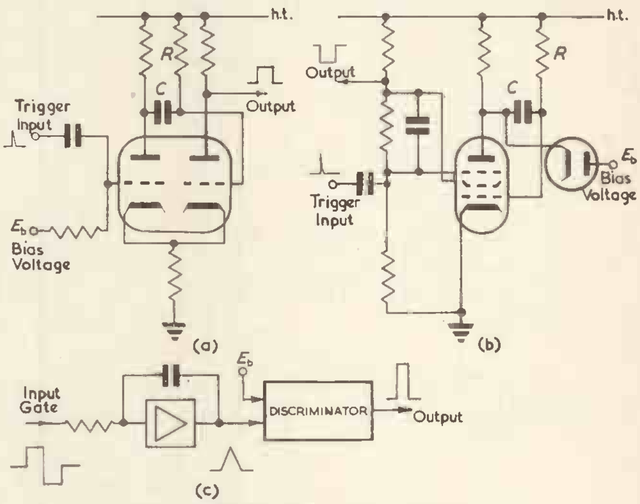


Fig. 2. Time modulation
(a) Delayed multivibrator circuit
(b) Simple Miller integrator
(c) High accuracy system

The average value of voltage therefore is:

$$V_{eq} = (t \cdot A)/T$$
$$= KA$$

If the true duration of the pulse is made proportional to one of the variables X, and the amplitude to the other variable Y, the average value will be proportional to their product XY.

Most methods for producing rectangular pulses, the duration of which is accurately controlled by a voltage, consist of generating a sawtooth waveform and then using this waveform to trigger a discriminator when the voltage reaches a value E, which is made proportional to one of the variables. There are very many circuits which could perform this function[1,2,3]. Probably the simplest of these is an unstable or delay multivibrator (Fig. 2(a)). The timing waveform is generated by charging the capacitor C through the resistor R. This waveform is exponential in form and therefore the accuracy of this circuit is of the order of a few per cent.

Fig. 2(b) shows a simple Miller integrator or phantastron. The timing waveform in this case is generated by the well-known Miller principle and is substantially linear. The length of the sweep is determined by the potential of the "catching" diode which is returned to the biasing potential

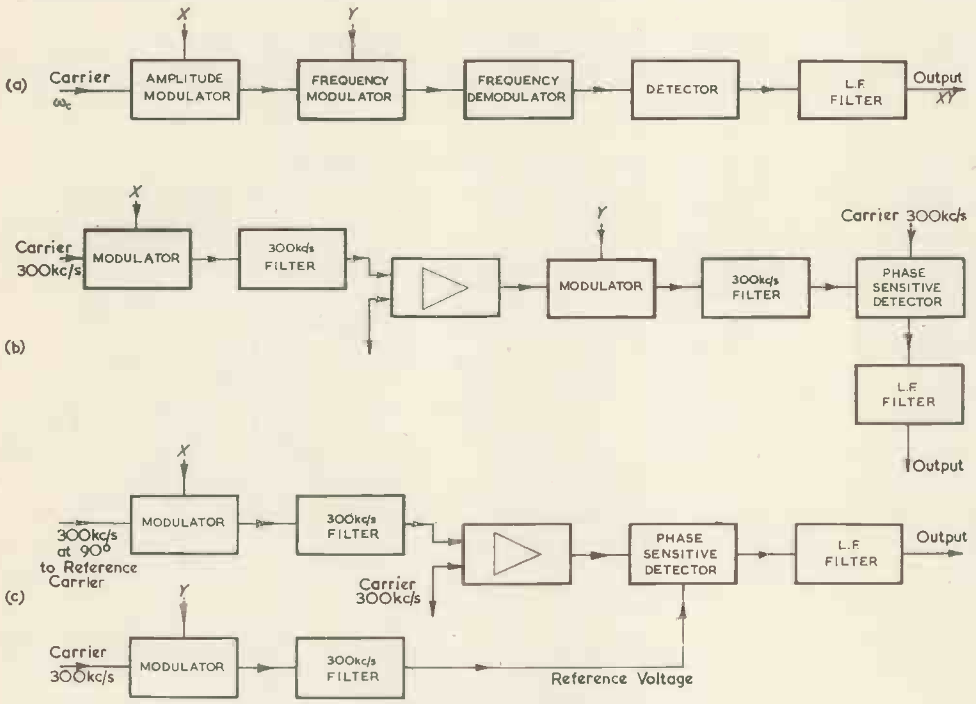


Fig. 3(a). f.m.—a.m. multiplier
(b). Double modulation multiplier
(c). Phase and amplitude modulation

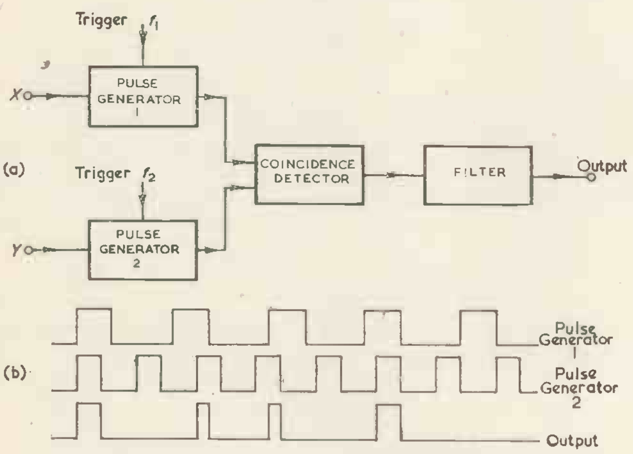


Fig. 4. Coincidence multiplier

E. For the detailed discussion of this and many similar circuits the reader may refer to works on radar measurements such as ref. 1.

A more elaborate method is illustrated in Fig. 2(c). A rectangular gate waveform is applied to a high gain integrating amplifier which produces a sawtooth voltage of considerable linearity. This voltage is then applied to a discriminator which produces a pulse of required length.

The accuracy of all these systems depends on:

(a) The linearity of the sweep;

(b) The stability of the discriminator in respect of such factors as heater voltage variation, valve and component changes etc.

The speed of response of time modulation multipliers depends on minimum length of pulse possible, the circuits in question and also on the speed of response of the l.f. filter which produces the average value of the waveform.

Amplitude, Frequency and Phase Multipliers

The simplest system of this nature uses a double modulation of a carrier waveform: amplitude modulation proportional to one of the variables X and frequency modulation proportional to the second variable Y (Fig. 3(a)). The resulting signal is then of the form:

$$x \sin [(\omega + y)t]$$

It is then passed through a frequency sensitive demodulator and a rectifier so that the demodulator signal is proportional to the product XY. The simplest system of this nature[3] uses modulators and demodulators of the type used in radio-communication. The resulting apparatus is quite simple and has a good response time. The linearity of the modulators could be improved by the use of negative feedback which reduces the error.

Another type of modulation multiplier[4] uses the modulation product of two waveforms of different frequencies which are both amplitude modulated. The block diagram is given in Fig. 3(b).

Both modulators marked in the diagram are of the diode limiter type and the clipping level of the modulator is proportional to the modulating voltage. The carrier waveform at frequency of 300kc/s is applied to the first modulator which is followed by a filter which extracts the 300kc/s component of the modulated waveform. This signal is then added to another carrier of frequency 750kc/s and the result is modulated by the second variable Y. The 300kc/s component is again extracted by a second 300kc/s filter and is applied to a phase sensitive rectifier, so that the polarity of the input can be recovered.

Another type of multiplier[4] is illustrated in Fig. 3(c). The variable X is used to modulate a 300kc/s signal which is phase displaced by 90° in respect of the carrier waveform. The first harmonic of the result is then extracted by means of a tuned filter and added to a constant amplitude carrier. The resulting waveform is used as a reference voltage in a phase sensitive rectifier. The voltage which is applied to the rectifier is the carrier modulated by the second variable Y. The result is proportional to the product XY.

For the mathematical analysis of the last two multiplier systems which is fairly complicated, the reader may refer to ref. 4 and also 5.

Coincidence Multiplier

This multiplier makes use of the theory of probability. If two or more rectangular waveforms of different but fixed frequencies and variable duty ratios are compared, the time during which the pulses coincide will be proportional to the product of their respective duty ratios.

The pulse generators which are illustrated in Fig. 4(a) can be of any form, similar to those described for use with the time modulation multiplier. They produce pulses of constant amplitude but of variable duration proportional to the input voltages X, Y. The pulses are fed into a coincidence detector, the output of which is a train of pulses corresponding to the time during which all the input pulses coincide. As the amplitude of these pulses is constant, their average value is proportional to the product of the input voltages. The advantages of this system are that more than two variables can be multiplied at the same time and also that somewhat difficult pulse amplitude modulation is avoided because all the amplitudes are kept constant and only the output amplitude is critical. The accuracy of the system is largely determined by the variable length pulse generators.

Multiplication by Squaring

All multipliers working on this principle make use of the relation:

$$XY = \tfrac{1}{4}[(X + Y)^2 - (X - Y)^2]$$

To achieve multiplication by this method the minimum of three adding or subtracting devices and two squarers are required. The method is essentially a single quadrant one. To restore the polarity of the output, special invertors are required which are operated by the input signals. High speed relays can also be used for restoring the polarity.

Of the various methods of squaring, that of segmented diode characteristics[6] is the most accurate and stable one. Though it requires a relatively large number of components the circuit parameters can easily be calculated as the method does not depend on the characteristics of valves or other non-linear elements. There is also no difficulty with initial setting or change of components. The general principle of the method is illustrated in Fig. 5(a).

A voltage A is applied to the potentiometer chain arranged as shown in Fig. 5(b). The points along this chain are connected to a number of diodes returned to bias voltages K_1A, K_2A etc. Neglecting the voltage drop across the diodes, the circuit responses will be along line $B = K_0A$ as marked on Fig. 5(a) for values of $0 < A < K_1A_1$. When the voltage A reaches value: $K_1A_1 < A < K_2A$ the response will be along the line K_1A. In this way a curve is obtained composed of a number of straight lines which can be made to approximate to the required parabolic law. An overall accuracy of better than 1 per cent can be obtained with only 11 diodes[6]. Thermionic diodes are preferable in this application to the crystal ones by virtue of their more predictable characteristics.

A large number of non-linear squaring elements have been developed mostly for use in the power and r.m.s. value measurements and can be applied to analogue multiplication. The non-linear elements which are used all have the disadvantage that the conditions of operation are very critical and usually not reproducible. Pre-set controls must be used to set the initial conditions and these have to be adjusted at intervals or when an element is replaced. Several of these elements will be briefly mentioned here.

A temperature limited diode[7] has the anode current proportional to the square of the heater current. Though this relation is a fairly accurate one, the time response of such a device is necessarily limited.

A number of multi-grid valves exhibit the square law characteristics[8,9]. Usually the input voltage is applied to one or more of the grids and the output is in the form of the anode current. To convert it to voltage again this current must be passed through a resistor. The potentials of all the electrodes are critical and must be determined by experiment. The range of such a device is usually very limited. Two or more valves can be connected in parallel to obtain more average characteristics.

Special cathode-ray tubes can be used for squaring. One of these[10] uses a beam of electrons in the form of a thin sheet which is deflected in a transverse direction across a suitably shaped mask. The current through the tube will, therefore, be proportional to the square of the deflecting voltage.

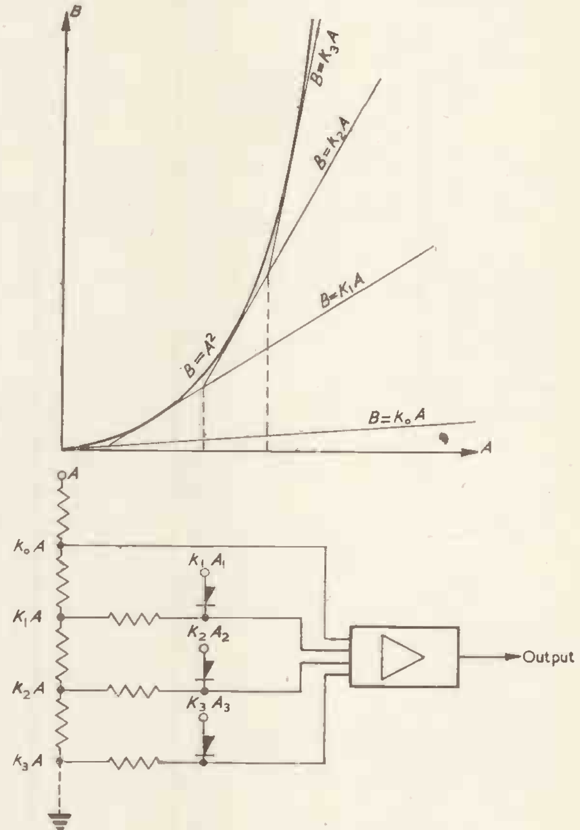


Fig. 5. Multiplication by squaring

An interesting method of squaring[11] is that of clipping a triangular waveform. A sawtooth waveform of constant frequency is clipped off so that the remainder consists of a number of triangles, the height of which is proportional to the quantity to be squared. The area under the waveform or its average value is then proportional to the square of the variable. The accuracy of this method depends on the same factors which were discussed in connexion with multiplication by means of square pulses, i.e. linearity of the waveform and the accuracy of switching.

The Logarithmic Multipliers

An apparently simple method of multiplication consists of generating voltages proportional to the logarithms of both variables, adding them and extracting the logarithm of the sum. The difficulty here, as in the case of squaring multipliers, lies in the construction of the non-linear elements.

The most common logarithmic elements are metal rectifiers and thermionic valves. All copper oxide as well as selenium and germanium rectifiers exhibit the logarithmic characteristics of the form:

$$v = K \log i + K_1 i$$

where v and i are the voltage and current through the rectifier. It may be said, therefore, that a metal rectifier consists of a true logarithmic convertor in series with linear resistance. The best results are obtained with currents which are much smaller than the rated current of the rectifier. For

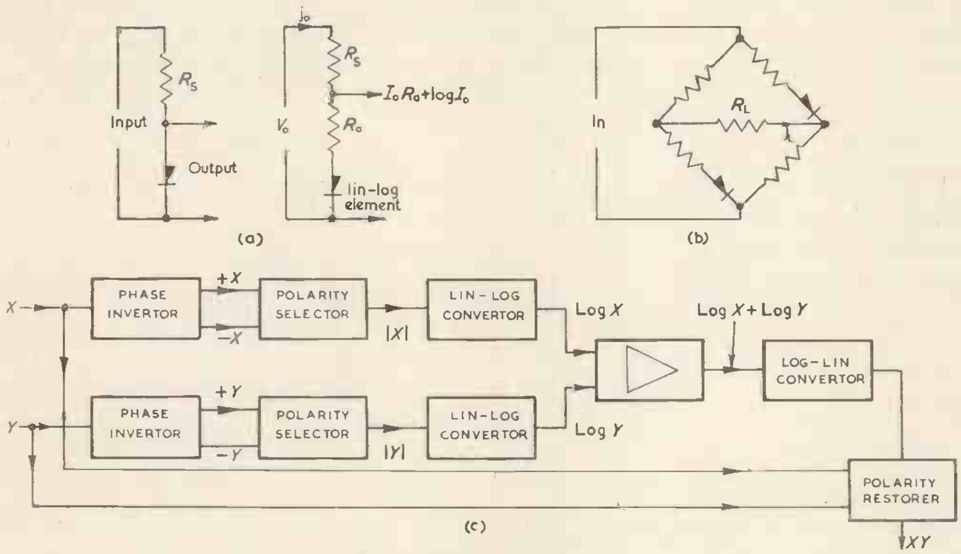


Fig. 6(a). Equivalent circuit
(b). Bridge connexion
(c). Logarithmic multiplier

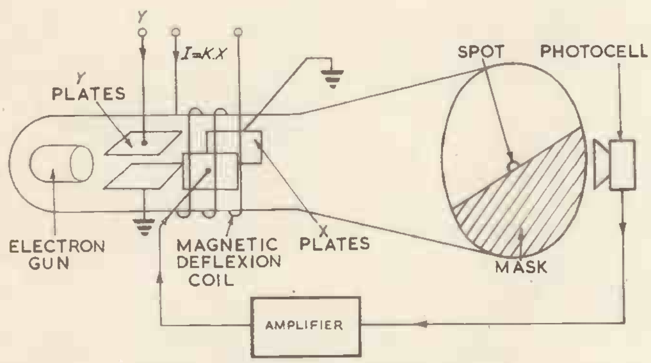


Fig. 7. Basis of an electron beam multiplier

instance, the author obtained a very accurate relationship between the voltage and current in the range of 1μA to 10mA for a copper oxide power rectifier normally rated at 0·5A. A fairly reasonable characteristic can be obtained over as many as five decades. The stability of such a device is, however, very poor, the influence of the changes of the ambient temperature being particularly troublesome. Also, the very small values of current and voltage involved make these elements somewhat difficult to apply in accurate analogue computation.

Another source of error which is quite independent of the logarithmic characteristics is the necessity of converting the output from the form of current into a voltage. A simple device of putting a large resistor with series with the logarithmic element (Fig. 6(a) introduces an error unless the voltage drop across it very much larger than across the rectifier. A bridge connexion of two diodes[2] can be used to compensate for the linear part of the characteristics (Fig. 6(b)).

Thermionic valves[12] can also be used to obtain logarithmic characteristics. Diodes or multi-grid valves connected as diodes, can be used in a manner similar to metal rectifiers[13]. Such valves are operated in the region of very small anode currents (a few microamps). Triode valves in which the grid is used, both as a diode anode and a control grid, can also be employed with success[14].

On the whole, the logarithmic devices provide means of simple and fast multiplication of medium accuracy, but suffer from the disadvantage of the necessity of frequent re-calibration. Thermionic valves are generally more stable than semi-conductor elements, provided that the heater supplies are reasonably stabilized. The method is obviously a single quadrant one and some means of recovering the polarity of the output must be used. An example of such a method is illustrated in Fig. 6(c).

Cathode-Ray Tube Multipliers

Crossed Field Multiplier

This ingenious device[15] makes use of the laws which govern the movement of electrons in a combined electrostatic and magnetic field. Fig. 7 illustrates the principle. The beam of electrons passes between the Y-plates of a cathode-ray tube to which a potential proportional to the voltage Y is applied. Therefore, the velocity of electrons in the Y direction will be:

$$V_y = KY$$

A coil is arranged on the neck of the tube so that the lines of magnetic field are parallel to the axis of the tube. The current I through this coil is made proportional to the second variable Y. The magnetic field H will, therefore, produce a deflexion in the Y-direction proportional to: HV_y i.e. to: XY. The electron beam produces a spot on the fluorescent screen, half of which is covered by a mask with a straight edge along the X-axis. A photocell is arranged in front of the tube face and feeds an amplifier in such a way that the balanced conditions are obtained with only half of the spot visible above the mask. The output of the amplifier is connected to the Y-plates. When sufficient voltage is applied to reduce the Y-deflexion back to zero, this voltage will be proportional to the deflexion produced by the magnetic field[15], i.e. to the product XY.

A further development of the same principle uses a special tube in which the screen is replaced by two metal electrodes or collectors. The potential on these plates due to the electron current through the tube is balanced by an amplifier exactly as before.

Circular Beam Multiplier

This multiplier[16] which works on quite a different principle, uses a specially constructed cathode-ray tube. The beam of electrons is of large circular cross-section and of uniform density. It is allowed to fall on four quadrant plates, as illustrated in Fig. 8. The beam is deflected from equilibrium by voltages proportional to the variables X and Y. The currents collected by these plates are added algebraically with signs as shown. If the centre of the beam is deflected by a distance of X_1 and Y_1 it can be proved by simple geometry that the result of the summation will be an area of $4X_1Y_1$.

The accuracy of these devices depend mostly on the construction of the cathode-ray tubes concerned. The configuration of the magnetic field in the crossed field

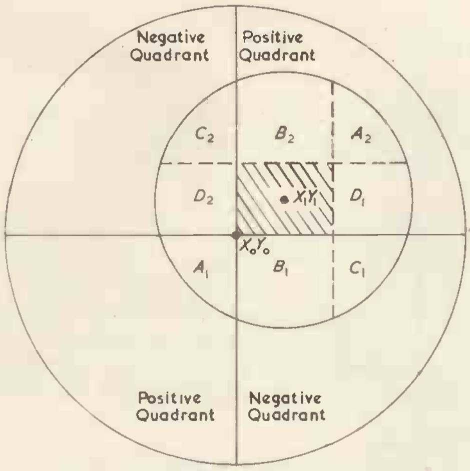


Fig. 8. Circular beam principle

multiplier and the uniformity of the electron beam in the case of the circular beam are of primary importance.

The speed of response of these instruments is very good and is limited mostly by the need to drive a magnetic coil at high frequency in the case of the crossed field multiplier. A four-quadrant operation is also a great advantage.

Multipliers Using Special Characteristics of Vacuum Valves

These devices are certainly the simplest that can be devised for analogue multiplication. Unfortunately, their range and accuracy is very limited. The simplest example of these is a variable-mu valve which accepts one of the signals in the form of an alternating voltage, whereas the other signal is in the form of d.c. bias. The amplitude of the output is proportional to the product of the two inputs, provided that the a.c. signal is kept very small and that the d.c. signal is applied with proper reference level and polarity. Tuned filters can be used to eliminate the second harmonic present due to the curvature of the grid characteristics, whereas two valves in push-pull can be used to reduce the common mode effect. The problem is well known in radio-frequency modulators and extensive literature exists on the subject.

Two control grids have been used to modulate the anode current of a multi-grid valve. Such a device, however, has a very limited range[1] and is sensitive to the setting of the d.c. voltages.

REFERENCES

1. CHANCE *et al.* Multiplying Devices Using Carrier Waveforms. Rad. Lab. Series 19 (Waveforms).
2. CHANCE, HUGHES, MACNICOL, SAYRE. Multiplier Using Tube Characteristics. Rad. Lab. Series 21, Ch. 3.
3. SOMERVILLE. An Electronic Multiplier. *Electronic Engng. 24,* 79 (1952).
4. MEYER, M. A., TULLER, H. W. Two New Electronic Analogue Multipliers. *Rev. Sci. Instrum. 25,* 1166 (1954).
5. STENBERG. Two Frequency Modulating Products. *J. Math. Phys. 32,* 233 (1953) and *33,* 68 (1954).
6. MARSHALL, B. O., Jr. An Analogue Multiplier. *Nature* 167, 29 (1951).
7. DRAPER, J. H. P., TUCKER, D. G. A Square Law Circuit. *J. Sci. Instrum. 24,* 20 (1947).
8. McG. ROSS, H., SHUFFREY, A. L. An Electronic Square-law Circuit, *J. Sci. Instrum. 25,* 200 (1948).
9. EL SAID. Novel Multiplying Circuits With Application to Electronic Wattmeters. *Proc. Inst. Radio Engrs. 37,* 1003 (1949).
10. SOLTES, A. S. Beam Deflection Non-Linear Element. *Electronics 23,* 122 (August, 1950).
11. NORSWORTHY, K. H. A Simple Electronic Multiplier. *Electronic Engng. 26,* 72 (1954).
12. MEAGER, R. E., BENTLEY, E. P. Vacuum Tube Circuit to Measure the Log of Direct Current. *Rev. Sci. Instrum. 10,* 336 (1939).
13. SNOWDEN, F. C., PAGE, H. T. An Electronic Circuit which Extracts Antilogarithms Directly. *Rev. Sci. Instrum. 21,* 179 (1950).
14. HOWARD, R. C., SAVANT, C. H., NEISUANDER, R. S. Linear to Logarithmic Voltage Converter. *Electronics 26,* 156 (July, 1953).
15. DEELEY, E. M., MACKAY, D. M. Multiplication and Division by Electronic Analogue Method. *Nature 163,* 650 (1949).
16. ANGELO, E. J., Jr. An Electron-Beam Tube for Analogue Multiplication. *Rev. Sci. Instrum. 25,* 280 (1954).

(To be continued)

The Effect of Sunspot Activity

A three or four year period during which appearances of sunspots on the sun's surface will be more frequent than at any time since the year 1949, is now beginning.

The reliability of long-distance wireless communications is affected by sunspots and the following notes have been prepared by Cable and Wireless Ltd.

It is correct to say that some sunspots interrupt long-distance wireless circuits; but incorrect to infer that the more numerous the sunspots the poorer the wireless conditions. On the contrary, it is generally true that the greater number of sunspots the better the performance of long-distance wireless circuits.

The degree of ionization and the quality of reflection of the ionosphere is affected by several causes. Among them are sunspots and magnetic storms.

Sunspots are a visible indication of changes in the sun. The frequency of these changes rises and falls with a rough time interval of 11 years between successive peaks and between successive troughs. The peak periods of change are known as periods of sunspot maximum, and the troughs as sunspot minimum periods. As energy from the sun is responsible for the formation of the ionosphere, the ability of this region to reflect radio waves back to the earth also changes steadily over the solar cycle, being greater at sunspot maximum than at sunspot minimum.

In addition to this steady, predictable change in solar activity, there occur relatively short-lived changes, usually of a violent character, which also affect the ionosphere—mainly by reducing its reflecting properties. Such short-lived changes, which may be due to a magnetic storm or flaring sunspot, are called ionospheric disturbances.

During a magnetic storm, for instance, the ionization often becomes so weak that radio waves pass right through the region and are lost in space. The circuit then fails. On the other hand, when a sunspot flares, the ionization often becomes so intense that the radio signals are absorbed before reaching their destination, and the circuit "fades out".

The most severe magnetic storms may also affect submarine cables and can cause short periods of interruption.

Disturbances occurring when the reflecting properties of the ionosphere are already low, have a greater effect on wireless circuits than disturbances occurring at sunspot maximum, when the reflecting properties of the ionosphere are at their greatest.

The ionospheric stroms which occur at sunspot minimum cannot as yet be associated with any visible solar characteristic. In addition, they are not great in absolute terms—i.e. when measured in an ionospheric observatory, they are less significant than those occurring during sunspot maximum. They are however more prolonged, and because of the minimum reflecting properties already prevailing, their effect on radio circuits is considerable. Circuits across the North Atlantic are often unworkable for several days at a time.

At sunspot maximum, however, the sun's disk shows a constantly changing pattern of sunspots. Some of these spots give rise to ionospheric disturbances which, although often violent, are much less prolonged than those at sunspot minimum. Others may erupt into gigantic flares, the largest of which cause total interruption of high frequency wireless communcation over the part of the world in daylight at the time of the flare.

These solar flares are of considerable scientific interest and the associated wireless "black-outs" are known as "Dellinger type" fade-outs. Although "Dellinger type" fade-outs are a nuisance, the period of circuit failure is usually about half an hour—a fraction of the lost time associated with the sunspot minimum disturbances.

A period of sunspot maximum is now rapidly approaching and "Dellinger type" fade-outs will be almost a daily occurrence. But the overall effect on wireless communications of both Dellinger and magnetic storms, will be off-set by the generally improved reflecting property of the ionosphere. As a result, radio circuit performance will continue to imorove compared with the sunspot minimum years of 1951 to 1954.

A Wide Range Photo-electric Automatic Gain Control

By C. Riddle*, B.Sc., A.M.I.E.E.

A photocell and valve are arranged in such a way that the output voltage is proportional to the light modulation, and independent of the value of the steady light flux. The circuit is extremely simple, and the range over which the light flux may vary is very large (100 000 : 1).

THE basic circuit is shown in Fig. 1, and consists of a photocell fed from a 90V d.c. supply through a high value resistor. When illuminated, the photocell delivers current to a diode which acts as the photocell load impedance.

The operation depends upon two separate factors:

(1) Over the range of currents used (0·03μA to 9μA), the diode voltage/current relationship is exponential, from which it follows that for small current changes the a.c. resistance formed by the diode is inversely proportional to the current. This means that a small percentage change of current will always produce the same diode voltage, irrespective of the actual d.c. value of the current.

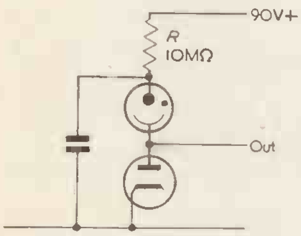


Fig. 1. Basic circuit

(2) The photocell sensitivity will fall as the light increases, due to the voltage drop across the resistor R. If a gas-filled photocell is used, the initial sensitivity at 90V may be 200μA/lm, but this falls by several hundred times when the voltage is reduced to, say, 0·5V.

The total range of operation of the circuit is therefore the combined effect of a loss of sensitivity of several hundred times, together with a current range of about 300 : 1, which gives an overall range of approximately 100 000 : 1.

For constant output the following conditions must apply:

(a) The diode characteristic must be exponential.

(b) The photocell light/current relationship must be linear.

(c) The load impedance formed by the diode must be small in comparison with the photocell internal impedance.

In fact, none of these conditions are quite maintained, but the variations are usually fairly small, and will be discussed later.

The operation of the circuit can now be described as follows:

At very low illuminations, the photocell will be operating at maximum sensitivity, and the diode a.c. resistance will be several megohms. As the light increases, the current increases and the photocell sensitivity falls. When the current reaches 6 or 7μA, the photocell gas amplification disappears, and thereafter the photocell behaves as a vacuum photocell. When the current reaches its maximum value of 9μA, the diode a.c. resistance is reduced to about 10kΩ. As the light increases still further, the photocell sensitivity is reduced in proportion to the increase in light. The limit is reached when the voltage across the photocell falls to a value comparable with the signal (see Fig. 9).

Design Principles

O. W. Richardson's formula for the thermionic emission current from a hot solid can be expressed as:

$$I_s = AT^2 \exp(-\phi/kT) \quad \ldots\ldots\ldots\ldots \quad (1)$$

where I_s = current/cm²
T = absolute temperature
ϕ = Work Function (i.e. the additional energy required to remove the fastest electrons from the parent surface)
k = Boltzmann's constant (0·0863 × 10^{-3}V/°K)

In a diode the anode current will be equal to the emission current from the cathode only if the anode potential is large and positive.

For smaller positive values of anode potential, the electron flow is modified by space charge effects, and in this region the well-known Child-Langmuir three-halves power law applies:

$$I = \text{constant} \times V^{3/2}$$

The space charge becomes smaller when the anode potential is reduced and becomes negligible when the anode is more than a fraction of a volt negative.

In the negative voltage region the three-halves power law does not apply. Instead, in equation (1) the Work Function may be considered to have increased by a factor eV, where:

e = charge of 1 electron.

The anode current is therefore:

$$I = AT^2 \exp\left[-\frac{\phi + eV}{kT}\right]$$
$$= I_s \exp[-(eV/kT)] \quad \ldots\ldots\ldots\ldots \quad (2)$$

where I_s is the diode saturation current.

Most small receiving valves have an oxide coating running at about 1 000°K, and the formula becomes:

$$I = I_s \exp[-(V/0{\cdot}086)] \quad \ldots\ldots\ldots\ldots \quad (3)$$

In the circuit described, the anode is kept sufficiently negative so that equation (3) always applies.

For any steady applied potential, the diode has a d.c. resistance which is given by:

$$R_{dc} = (0{\cdot}086/I) \text{ logh } [I/I_s]$$

For small voltage swings about this potential, the diode may be said to have an a.c. resistance which is given by:

$$r_{ac} = (dV/dI) = (0{\cdot}086/I) \quad \ldots\ldots\ldots\ldots \quad (4)$$

* *Ministry of Supply, A.R.D.E.*

If the current changes by a small amount ΔI, the voltage across the diode changes by:

$$r_{ac}\,\Delta I = 0{\cdot}086\,(\Delta I/I) = m \times 0{\cdot}86\text{mV},$$

where m is the percentage current modulation.
(It will be seen that the signal will have a maximum amplitude of 86mV).

The voltage change is therefore proportional to the current modulation, and independent of the actual value of the current, over the complete range at which equation (3) is valid.

The range of current over which the equation is valid depends upon the type of diode used. For convenience,

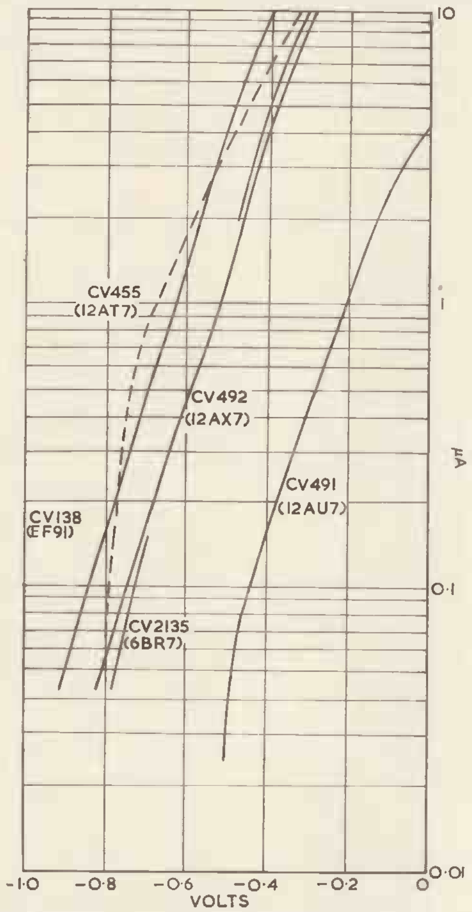


Fig. 2. Grid voltage/grid current characteristics of some triodes and pentodes

an ordinary small receiving triode or pentode may be used, with the control grid acting as the diode anode. The valve is otherwise used as an ordinary amplifying valve, with the anode current controlled in the usual way by the control grid voltage. (This is in contrast to some applications, where the valve is run at very low anode currents, and a logarithmic grid voltage/anode voltage relationship is obtained[1].)

The lower limit of validity of equation (2) depends upon the negative grid current flowing from the grid, due to ionization of the residual gas, and also to grid emission. This varies widely from one valve to another, and increases with high anode potentials. Its effect is to make the net grid current zero at about $-1{\cdot}0$V to $-1{\cdot}8$V for oxide-coated unipotential cathodes at temperatures[2] of 1 000°K to 1 100°K.

The upper limit depends upon the diode anode (or valve grid) area and spacing. For small diodes it may be as high as 500μA, and for small receiving triodes and pentodes it may be between 10μA and 100μA.

Fig. 2 shows the experimental results on a series of small receiving valves, from which it may be seen that a range of a few hundred may be obtained with several different types. In general, it is best to keep the power dissipation to a minimum.

Distortion

Since the transfer characteristic is logarithmic, harmonic distortion will be high at modulations exceeding 30 or 40 per cent. Fig. 3 shows the distortion produced on:

(a) A single sinusoidal waveform at 90 per cent modulation,

(b) Two mixed sinusoidal waveforms.

In the latter case the valve is operating as a class-A grid modulator (and incidentally can be deliberately used in this capacity for a.f. to r.f. modulations up to about 30 per cent). From this it is evident that some sort of correction is desirable if high values of modulation are to be used.

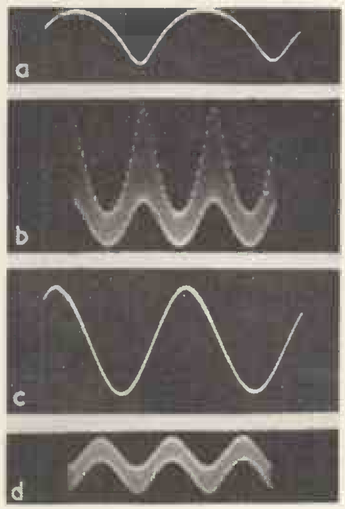


Fig. 3. Waveforms, showing distortions before and after correction

This can be done quite simply by adding a second stage identical to the first, and arranging that the signal amplitudes at both grids are the same.

The basic circuit is given in Fig. 4.

Let the initial currents and voltages be as shown. The relationship:

$$i = i_s \exp[-(v/0{\cdot}086)]$$

holds for both valves; or

$$\begin{aligned} 0{\cdot}086 \,\text{logh}\,(i/i_s) &= -v \\ &= 0{\cdot}086 \times 2{\cdot}3 \log i - 0{\cdot}086 \times 2{\cdot}3 \log i_s \\ &= 0{\cdot}2 \log i - k_0 \end{aligned}$$

If v_1 is large compared with v_2, then $i_1 = (v_1/R_1)$;

$$v_2 = 0{\cdot}2 \log(v_1/R_1) - k_1 = 0{\cdot}2 \log v_1 - k_2$$

Also:

$$v_4 = 0{\cdot}2 \log i_2 - k_3 = 0{\cdot}2 \log\left[i_{20} + \frac{v_3 - v_4}{R_2}\right] - k_3$$

Now let v_1 change to Mv_1.

$$\begin{aligned} \text{Change in } v_2 = \Delta v_2 &= 0{\cdot}2 \log v_1 - 0{\cdot}2 \log Mv_1 \\ &= 0{\cdot}2 \log M \end{aligned}$$

$\therefore \Delta v_3 = -0{\cdot}2\,A \log M$, where A is gain of 1st valve.

$$\begin{aligned} \Delta v_4 &= 0{\cdot}2 \log(R_2 i_{20} + v_3 - v_4) - 0{\cdot}2 \log R_2 - k_3 \\ &- 0{\cdot}2 \log(R_2 i_{20} + v_3 - 0{\cdot}2\,A \log M - v_4) + 0{\cdot}2 \log R_2 + k_3, \end{aligned}$$

(assuming Δv_3 is large in comparison with Δv_4).

If under steady conditions $v_3 = v_4$,

$$\Delta v_4 = 0{\cdot}2 \left\{ \log R_2 i_{20} - \log (R_2 i_{20} - 0{\cdot}2 A \log M) \right\}$$
$$= 0{\cdot}2 \log \left\{ \frac{0{\cdot}2 A \log M}{R_2 i_{20}} - 1 \right\}$$

Let the initial a.c. input resistance of the second valve be r_{20}.

$$r_{20} = (0{\cdot}086 / i_{20})$$
$$\Delta v_4 / \Delta v_2 = \frac{A r_{20}}{R_2 + r_{20}} \simeq \frac{0{\cdot}086\, A}{i_{20} R_2}$$

If R_2 is adjusted so that $\Delta v_4 = \Delta v_2$, then $i_{20} R_2 = 0{\cdot}086 A$;

$$\Delta v_4 = 0{\cdot}2 \log \; (\text{logh } 10 \log M - 1) \quad \ldots\ldots \quad (5)$$

In Fig. 5 is plotted $\Delta v_4 / M$, from which it can be seen that the transfer function is more or less linear up to fairly high modulation values.

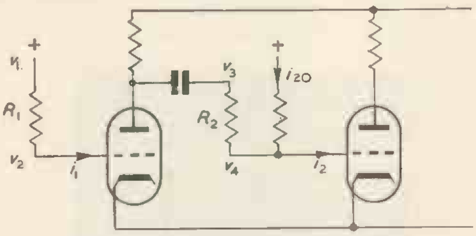


Fig. 4. Basic distortion reducing circuit

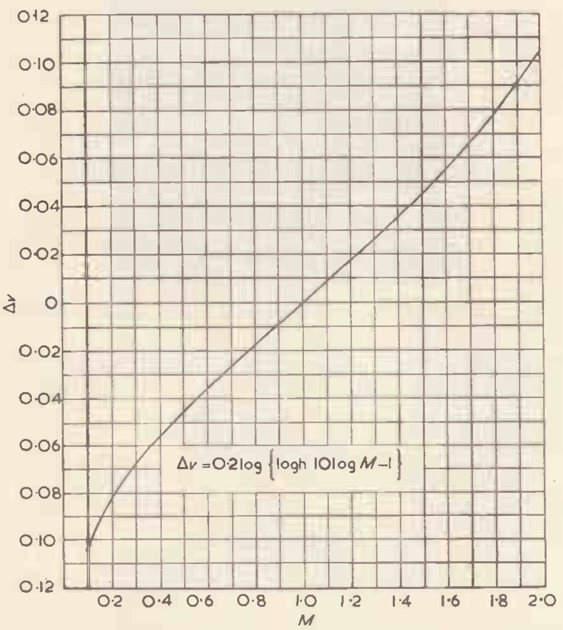


Fig. 5. Transfer characteristic of circuit with distortion correction

The result is independent of the value of i_{20}, and it should be noted that the values of i_s for the two valves need not be equal.

In practice, the second valve grid resistor may be connected to earth, and its value chosen so as to provide any convenient value for i_{20}.

Figs. 3(c) and 3(d) show the same signals as in Figs. 3(a) and 3(b), after correction in the second valve. The total harmonic distortion at 90 per cent modulation is about 5 per cent.

Fig. 6 gives the complete circuit, using a CV492 (12AX7) double triode.

Performance

There are many different types of valve which will give satisfactory performance up to 10μA. It is best to run the valve as cool as possible, both by ensuring adequate ventilation and by reducing the power consumption as far as possible. For operation at very low photocell currents, an electrometer type of valve might prove most satisfactory. A CV2135 (6BR7) has proved very satisfactory in an application where low microphony is important.

The photocell characteristics, however, are non-linear in almost every respect, although most of the non-linearities do not cause serious trouble.

Fig. 7 shows a family of curves taken on a particular CV2133 (90CG) photocell, with the d.c. load line superimposed. The a.c. load lines on this scale are almost vertical, which ensures that the a.c. output is independent of the shape of the curves, until the limit is reached when

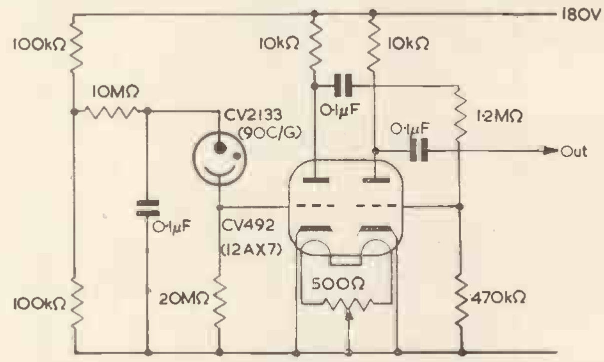


Fig. 6. Complete circuit, including distortion correcting circuit, using CV492 valve

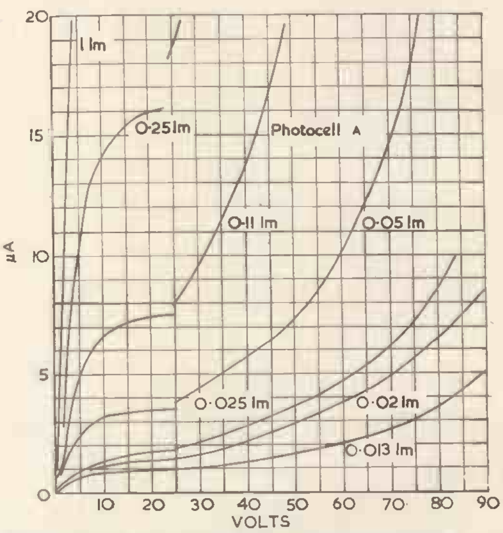


Fig. 7. Family of voltage/current characteristic curves for a particular CV2133 photocell, showing marked discontinuities at about 25V

the slope of the curves becomes comparable with the slope of the load line.

In some photocells, an abrupt change of current occurs at about 25V. This is shown to a marked extent in Fig. 7, which was much the worst out of about 24 tested. The shape of the curves varies slightly, depending upon whether the voltage is increasing or decreasing. The curves shown were taken with increasing voltage.

As a result of this discontinuity, there is a certain amount of instability at the point where the load line crosses a transition point. In most cases, the effect is not serious, and it can be avoided if necessary either by selection of photocells, or by reducing the photocell current by increasing the d.c. load resistance. Testing for stability can be carried out very rapidly by the use of a tungsten filament lamp with varying voltage supply, as described in the next section. Increasing the load resistance will reduce the range at the high brightness end, which for many applications is not serious. A range of 1 000:1 can be obtained with a d.c. load resistance of 200MΩ, and a maximum current of 0·45μA.

Fig. 8 shows the light/current relationship on a particular photocell at different voltages, from which it can be seen that the relationship is not linear at high or low voltages.

The non-linearity at high voltages is not important in this circuit, since the current is very small at high voltages. At low voltages the curve bends in the opposite direction. This means that for a given light flux the photocell voltage is higher than it would otherwise be. The effect is not so important as it would be in a linear amplifier, since in the latter the output is proportional both to the absolute amplitude and to the slope of the curve, whereas in this case the output is independent of the absolute amplitude, and proportional only to the slope of the curve.

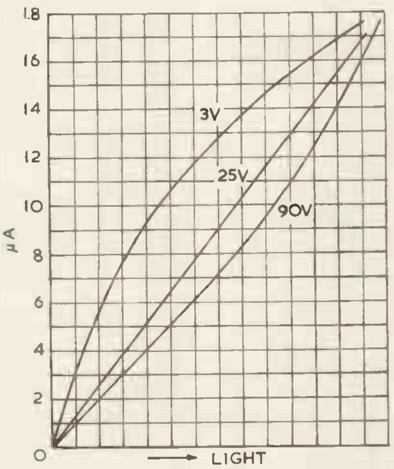


Fig. 8. Light/current characteristic of a particular photocell, showing non-linearities at high and low voltages. The light scale is linear in each case

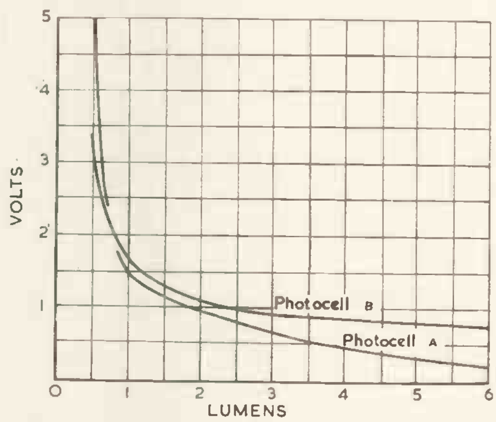


Fig. 9. Light/voltage characteristic of two different photocells, at constant current of 10μA

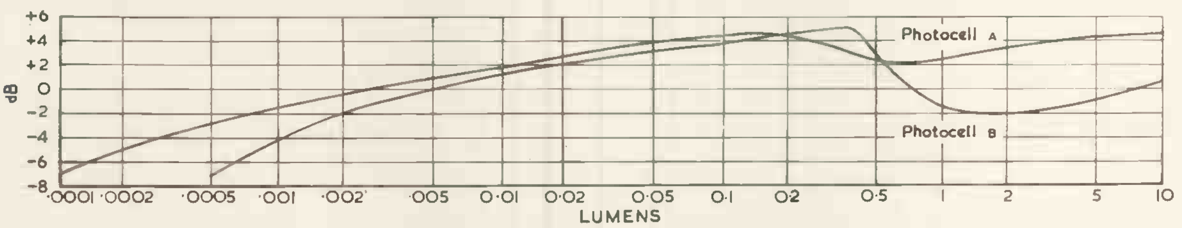


Fig. 10. Overall performance curves, using a 50c/s lamp source with 7 per cent modulation, showing the change in output due to the substitution of the two photocells A and B (*see Figs. 9 and 7 for characteristics of these two photocells*)

Different photocells may be considerably different in the low voltage portion of their characteristics, and this is reflected in the shape of the overall performance curves. Fig. 9 shows the light/voltage curves at constant current for the best and worst photocells out of 24 tested, and Fig. 10 shows the corresponding overall performance curves.

At very low voltages, the photocell E/I characteristic bends back into the negative voltage region, because at high illuminations some electrons have sufficient initial velocity to reach the anode even when the latter is negative.

Luminous flux values of up to 10lm are greatly in excess of those normally recommended for small photocells, particularly of the gas-filled type. Unless specially cooled, the photocell temperature rises considerably when exposed to such high illuminations for prolonged periods. At high temperatures the dark current increases considerably, and this wrecks the performance at the lower end of the range. The dark current, however, returns to its normal value as soon as the photocell is cooled, and the overall performance then remains unaffected. In a number of tests on CV2133 (90CG) photocells in which the temperature was raised by radiant illumination from a tungsten lamp to about 60°C or 70°C, the sensitivity in every case increased, sometimes by a considerable amount.

It would appear from this that short exposure to high illuminations does little damage, but if the full range of operation is to be obtained, it is necessary to keep the photocell cool, and to ensure that the dark current does not exceed about 0·01μA at 90V.

Bandwidth, Noise and Time-Constants

The bandwidth will vary with the illumination. At low intensities, when the photocell is operating at maximum sensitivity, the photocell response is down to about 70 per cent at 10kc/s, as compared with the response at low frequencies. When the voltage falls to 25V or less, the photocell operates virtually as a vacuum photocell, with a correspondingly high bandwidth. The valve input capacitance in parallel with the valve input a.c. resistance will then limit the bandwidth. The latter will therefore increase as the light increases and the load resistance decreases.

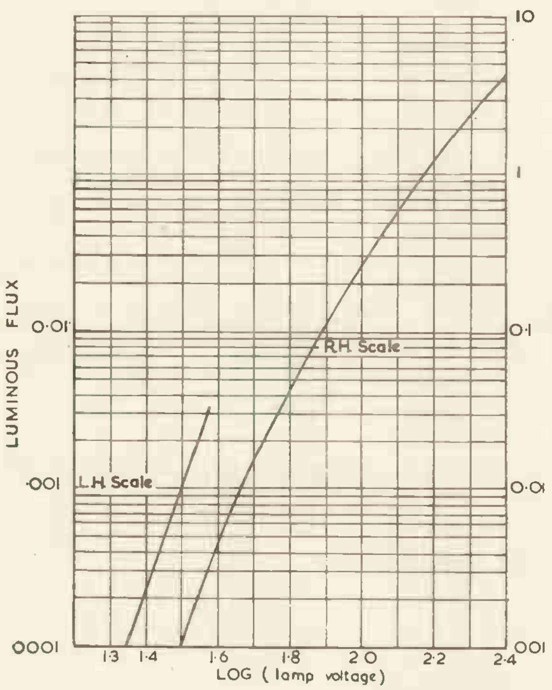
The signal-to-noise ratio will also increase as the light increases. This follows since for a given photocell voltage the signal current is directly proportional to the light flux, whereas the photocell shot noise is proportional to the square root of the light flux.

In Fig. 1, the decoupling capacitor serves a useful purpose only at high illuminations. For low illuminations it may be omitted, and the output will follow changes in d.c. level without delay. At high illuminations, when the photocell voltage is low, it is necessary to decouple the photocell anode, and this introduces a time-constant, the value of which can be chosen to meet the needs of any particular requirement.

Testing With Tungsten Lamps

An ordinary tungsten filament lamp run from a variable voltage 50c/s supply can be used to check the performance, since the percentage modulation at 100c/s due to the a.c. supply is constant for any given lamp. The percentage modulation varies with the type of lamp, being smaller with lower voltage lamps having thicker filaments. For a 200W 250V lamp, measurements show about 7 per cent modulation, and the efficiency is about 13lm/W.

If the light output is reduced by lowering the a.c. supply voltage, the frequency spectrum shifts towards the infra-red. Consequently, if a photocell is used to measure the light output, the readings are not directly proportional to the flux in lumens. This means that if different types of photocells are used to measure the light flux, the photocell current at any given lamp voltage will depend upon its colour response. Provided that the photocell used for measuring the flux is the same type as that used in the equipment, a direct estimate of luminous flux may be made.



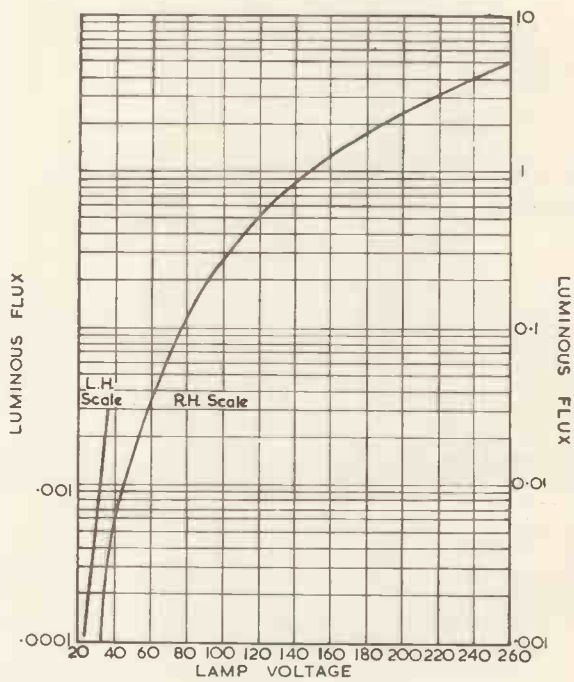


Fig. 11. Light output/lamp voltage for a 200W single coil tungsten filament lamp

Fig. 11 is a graph of lamp voltage/luminous flux as measured with an AgOCs photocell type CV2133 (90CG). It will be noticed that log (lamp volts)/log (flux) is almost constant over a wide range.

Equipment in Use

The circuit was originally developed for use in projectile velocity measuring equipment using two "skyscreens". Each skyscreen contains a lens and slit system, a photocell and an amplifier. The lens and slit system is designed so that the photocell is illuminated by a thin fan of light from the sky. A projectile passing through the field of view will interrupt some of the light, and so produce a signal. Two such signals, derived from two skyscreens placed some distance apart in the line of fire can be used to start and stop a chronometer. If the distance is known, the velocity can be calculated. The present design of skyscreen incorporates an automatic gain control so that the signal amplitude is independent of the sky brightness, and ensures constant output (for a given ratio of Height/Calibre) under nearly all English weather conditions from sunrise to sunset. The equipments have now been in use for several years and have given very satisfactory service.

Conclusions

The circuit can be used whenever it is desired to eliminate signal variations due to difference in brightness of the light source. (It does not cope with variations of stray incident light, which, of course, alter the modulation level.)

The circuit is simple and effective, and has the following advantages:

(1) The range of control is extremely wide.

(2) The voltage output is largely independent of the photocell characteristics.

(3) If the correcting circuit is not required, there are no additional components required as compared with a normal linear amplifier.

(4) With the aid of a second valve, the distortion can be reduced to reasonable proportions at high modulations.

(5) If an anti-microphonic valve is used on a suitable mounting, it can be made exceptionally anti-microphonic.

(6) No damage to the gas-filled photocell can result from accidental "glowing" due to high illumination.

(7) Vacuum type photocells can be used if necessary, with consequent improvement in bandwidth, at the expense of sensitivity.

(8) The current source need not necessarily be a photocell at all, since the diode fed through a high resistance for any positive current source will give a fairly wide range of operation.

(9) There is an automatic improvement in signal-to-noise ratios at high illuminations.

(10) None of the component values are critical, and wide tolerance components may be used.

REFERENCES

1. Valley, G. E., Wallman, H. Vacuum Tube Amplifiers. M.I.T. Radiation Laboratory Series No. 18, Section 11.4. (McGraw Hill).
2. Ibid Section 11.5.

Chrominance Circuits for Colour Television Receivers

(Part 2)

By B. W. Osborne*, M.Sc., A.Inst.P., A.M.Brit.I.R.E

Chrominance Demodulator and Matrix Circuits

The I-Q System

We have seen that the chrominance signal is obtained at the transmitter by combining the E_I' and E_Q' colour difference signals in phase quadrature, so that the instantaneous phase of the chrominance signal depends on the relative instantaneous amplitudes of the E_I' and E_Q' signals. It follows that, at the receiver, the E_I' and E_Q' components may be obtained separately by mixing the incoming chrominance signal with locally generated c.w. phase reference signals lagging the reference burst by 57° and 147° for the *I* and *Q* axes respectively (Fig. 1).

Various alternative methods are available for the generation of the local phase reference signal as described previously. This signal must be locked in phase at the desired angle relative to the phase of the reference burst, and a master phase control is then used, in conjunction with a 90° phase shift network to produce two outputs in phase quadrature and having the desired phase relationship to the burst. In the *I-Q* system, the synchronous demodulators operate along the *I-Q* axes (Fig. 1). These axes are so chosen that the E_I' signal distinguishes between orange-red and blue-green, while the E_Q' signal distinguishes between yellow-green and purple. In order to avoid crosstalk between the channels without increasing the overall video frequency bandwidth beyond that in use for monochrome transmissions, the E_I' signal is transmitted as a moderately wide band, partly single-sideband signal and the E_Q' signal as a narrow-band, double-sideband signal. In the proposals for the British adaptation of the N.T.S.C. system[2], the frequency response of the *I* channel taken to the −3dB point extents to approximately 1·2Mc/s and that of the *Q* channel to approximately 350kc/s.

Thus for fine structure, corresponding to video frequencies above 1·5Mc/s, there is no colour information. For very small objects the human eye has no colour sense, and the absence of fine structure colour information is no disadvantage[4].

For small objects, corresponding to the frequency band passed by the *I* channel but not by the *Q* channel, the eye is capable of distinguishing between orange-red and blue-green, but not, say, between blue and green, so that colours can be matched using only two primaries. In other words, the chromaticity diagram for small objects becomes a straight line, from orange-red to blue-green. The *I* and *Y* channels alone are then capable of providing adequate colour presentation.

For larger objects, a three-primary colour system is necessary, and this is effectively available over the video frequency band up to the limit of the *Q* channel response.

A block diagram of an *I-Q* demodulator system is shown in Fig. 14. The video signal is fed through a band-pass filter accepting the sub-carrier and its upper *Q* and lower *I* sidebands (i.e. from approximately 1·4 to 3·1Mc/s for a sub-carrier of 2·66Mc/s). The output is fed to the control grids of the two demodulators, which mix the chrominance signal with phase reference signals in quadrature lying along the *I* and *Q* phase axes (Fig. 1). The separate *I* and *Q* outputs are then amplified and the unwanted high frequency components of the mixer outputs are filtered out. As the *I* channel chrominance components are essentially single sideband, the *I* demodulator load should theoretically have a rising characteristic in the upper frequency region, whereas the *Q* demodulator should have a flat

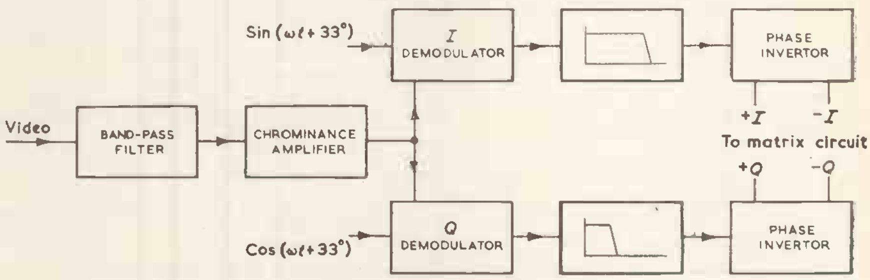


Fig. 14. Block diagram of I-Q demodulation system

response. *I* and *Q* signals of both polarities are required for the matrix circuit, so that phase invertor stages are included in both *I* and *Q* channels.

Any interaction between the two chrominance channels will result in the blurring of colour transitions. Precautions to be taken to prevent crosstalk between the *I* and *Q* channels include the use of circuits of relatively low impedance handling low level signals. The circuit layout must be arranged to reduce any coupling between the *I* and *Q* channels, and between either *I* or *Q* and the luminance channel. Poor amplitude linearity or variation in phase delay with frequency in the chrominance amplifier may also give rise to *I-Q* crosstalk.

In order to obtain correct relative *I* and *Q* signal levels at the input to the matrix system, the gain of one of these channels must be adjustable relative to the other. A chrominance control affecting both *I* and *Q* channels must also be provided in order to obtain the desired luminance/chrominance ratio at the matrix.

The Low-Frequency Colour System

The *I-Q* demodulator system described above is capable of using all the available signal information. However, the nature of the colour television signal makes it possible to obtain red, green and blue colour signals by demodulation along axes other than the *I-Q*, and this can lead to a considerable reduction in the number of valves and components required in the demodulator system.

If we consider only those colour difference signals at frequencies below 350kc/s, the expression for E_M quoted previously may be replaced, for instance, by the equivalent form:

$$E_M = E_Y' + (1/1{\cdot}14)\,[(1/1{\cdot}78)\,(E_B' - E_Y')\sin\omega t + (E_R' - E_Y')\cos\omega t]$$

* *Ultra Electric Ltd.*

By operating the synchronous demodulators along the *R-Y* and *B-Y* axes, it is thus possible to recover the $(E_B' - E_Y')$ and $(E_R' - E_Y')$ signals directly. Furthermore, as the *R-Y* axis lies at 90° to the reference burst, the *R-Y* phase reference signal is in the required phase for the input to the phase detector circuit in an a.p.c. loop. If a quadricorrelator phase lock system is used, the *B-Y* signal is in the required phase for the operation of the *S* detector. If a crystal ringing circuit is used, the output may be directly used to drive the *B-Y* demodulator, with a 90° phase shift network in the feed to the *R-Y* demodulator.

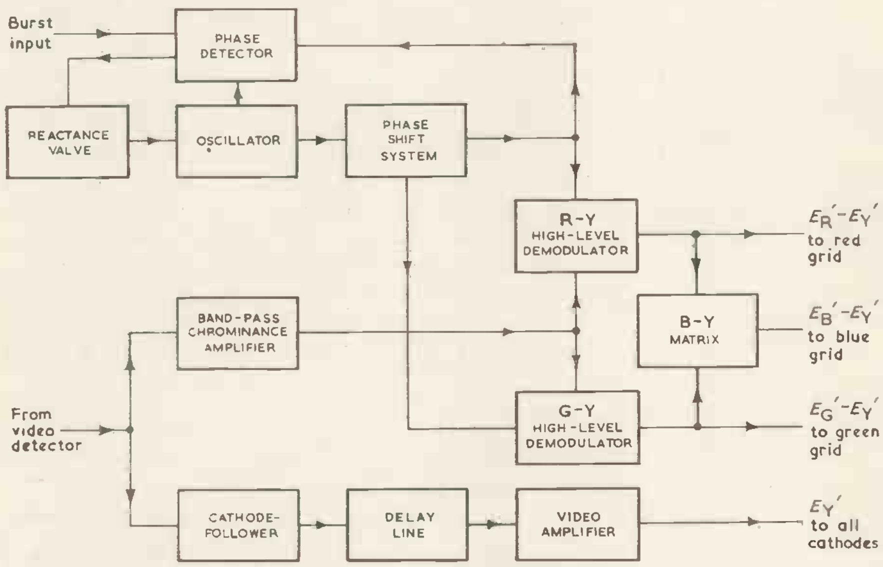


Fig. 15. The chrominance and luminance sections of a simple colour receiver

The green signal is obtainable from the blue and red colour difference signals by a simple matrix circuit, using the relationship:

$$(E_G' - E_Y') = -0{\cdot}51(E_R' - E_Y' - 0{\cdot}19(E_B' - E_Y').$$

Alternatively, in a receiver demodulating along the *R-Y* and *G-Y* axes, the red and green colour difference signals are obtained directly at the demodulator outputs, and the blue colour difference signal is obtained by matrixing the other two.

The output signals, limited in frequency to 350kc/s, have then to be separately added to the E_Y' luminance signal to obtain the red, blue and green outputs. The addition may be conveniently performed between grids and cathodes of a three-gun colour display tube.

The comparative simplicity of a receiver demodulating at a high level along the *B-Y* and the *G-Y* (or *B-Y*) axes is shown by the block diagram of Fig. 15.

The simple low-frequency system does away with the need for the *I* channel with its filter and delay, and simplifies the operation of the matrix circuits. It has been applied generally in medium-grade commercial receivers, where the slight loss in colour information is outweighed by the benefits gained by the simplification of the receiver circuits.

Some improvement in colour detail may be gained by accepting frequencies up to 500 or 600kc/s in the demodulator filters. Though the *I* signal is single-sideband at frequencies higher than 350kc/s and the acceptance of higher frequencies in a *R-Y*. *B-Y* receiver must lead to unwanted interaction between the demodulator channels, the cross-talk, however, is tolerable, and the acceptance of some of the *I* channel colour detail is beneficial.

Synchronous Demodulators

If we take a chrominance signal $A \cos(\omega t + \phi)$ and mix it with a reference signal $B \cos \omega t$ of the same frequency but of different phase, we get an output proportional to:

$$A \cos(\omega t + \phi) B \cos \omega t,$$

$$\text{i.e., to } AB/2\ [\cos\phi + \cos(2\omega t + \phi)].$$

If the high frequency $\cos(2\omega t + \phi)$ term is eliminated by the use of low-pass filters, we are left with an output proportional to $A \cos\phi$, which is a measure of the degree of saturation and the hue of the original colour.

When a valve with two control grids is used as a synchronous demodulator, the mixing process is in general non-linear. Livingston[16] has shown that as the output also contains a term, representing the distortion of the chrominance signal, and which is proportional to the square of the amplitude of the chrominance sub-carrier, it is advisable to operate the demodulators with as large a ratio of *B* to *A* as is practicable.

When the chrominance signal is applied to the control grid g_1 and the reference signal is applied to the suppressor grid g_3, the coefficient of the $\cos\phi$ term in the output is $AB/2\ (\partial g_{m1}/\partial e_{g3})$[16], where the transconductance g_{m1} is the rate of change of anode current with control grid voltage.

The value of $\partial g_{m1}/\partial e_{g3}$ for the 6F33 over a range of e_{g1} from -1 to -2V is approximately 700μA/V^2, appreciably greater than the value of 350μA/V^2 obtained for the 6AS6. The 6BA7 heptode is more linear as a demodulator than the 6AS6, but has lower conversion conductance. The 12AT7 (ECC81) double triode gives comparable performance to the 6BA7 when used in a cascade circuit[17].

Diode demodulators can be used[13], but suffer the disadvantages of being particularly sensitive to amplitude variations, and require unduly high reference carrier levels for satisfactory phase linearity, thus increasing screening problems.

Many demodulator circuits used to date employ a short suppressor base pentode in an unbalanced circuit, the chrominance signal being fed to the control grid at a level of about 1V and the reference carrier fed to the suppressor grid at as high a level as the grid base will accommodate.

A practical circuit employing a 6F33 valve is given in Fig. 16. The input is obtained at a conveniently low impedance from the chrominance band-pass filter, the resistive termination of the filter providing a suitable position for the chrominance controls. Self-bias is used, the control grid bias being set to the optimum working point at $-1{\cdot}5$V, for a peak-to-peak input chrominance signal of 1V, and the suppressor grid bias to -2V to enable the valve to handle a 4V phase reference signal on the suppressor grid. The output is fed into a four-terminal load network designed to give the desired low-pass characteristic. That shown in Fig. 15 is suitable for the *Q* demodulator of a 2·66Mc/s sub-carrier system.

Considerations influencing the design of synchronous demodulators include phase and amplitude linearity, high conversion conductance, ease of construction and adjustment, and the absence of any interaction with other circuits, particularly with the other demodulator. The circuits should be of small physical dimensions, so positioned that screening is either not required or is easily provided, should not be required to handle high level signals, and should have reasonably low input and output impedances.

However, the use of high-level triode demodulators in low-frequency colour demodulator systems makes it possible to d.c. couple the colour difference outputs to the c.r.t. grids, the addition of the luminance signal being performed between grids and cathodes of the display tube. Interaction between the two demodulators can be reduced

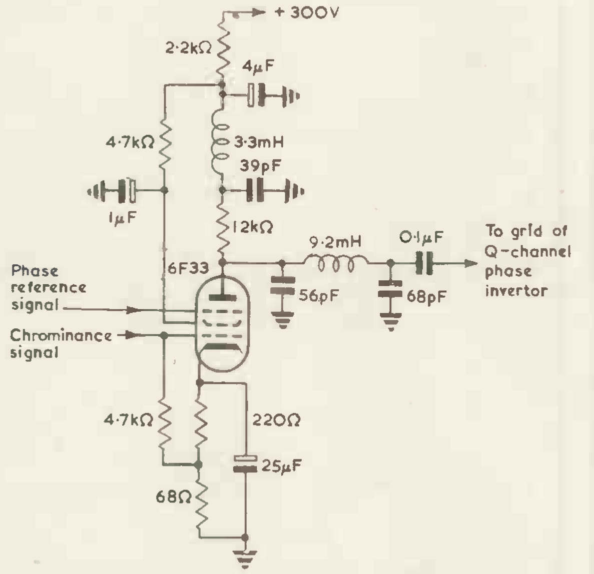


Fig. 16. A synchronous demodulator circuit

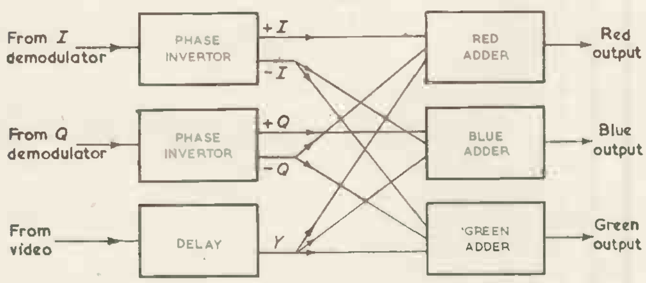


Fig. 17. Block diagram of I-Q matrix unit

to an acceptable level. In eliminating the necessity for using separate amplifiers in each colour channel, high level demodulation allows a reduction in the total number of valves in a colour receiver working along the *R-Y* and *B-Y* or *G-Y* axes, at the expense of some loss in phase and amplitude linearity.

In a receiver demodulating along the *B-Y*, *R-Y* axes the maximum demand from the *B-Y* channel is some 30 per cent greater than that from the *R-Y* channel, so that overloading of the *B-Y* channel limits the available output. Altes[18] has shown that an increase in the efficiency of a high level demodulator system may be obtained by choosing the demodulation angles so that the maximum outputs required in the demodulator output channels are equal, and has given curves from which the optimum demodulation angles for use in a particular system may be obtained. The optimum demodulation angles will in general vary with the relative drive requirements of the three guns of the colour display tube. For a tube requiring equal drive in each gun, an "*M*" signal may be derived by combining correct amounts of the demodulator colour difference outputs with the luminance signal, where:

$$M = (E_R'/3) + (E_G'/3) + (E_B'/3)$$

and

$$M\text{-}E_Y' = -0{\cdot}48\,(G - M) - 0{\cdot}19\,(R - M)$$

or

$$M\text{-}E_Y' = 0{\cdot}48\,(G - M) + 0{\cdot}29\,(R - M)$$

according to the polarity of the luminance signal at the *M* matrix. The demodulators are then operated along the *B-M*, *G-M* axes[18] and the maximum demodulator outputs are equalized. As the *R-M* axis lies close to the *I* axis, the full colour information may be used in a demodulator system of this type by feeding the wide-band chrominance signal to the *R-M* demodulator.

The new 6AR8 beam-deflexion valve has been found particularly useful in high-level demodulators, as both positive and negative colour difference outputs may be obtained from the one demodulator without the use of a phase-splitter stage[19].

Matrix Networks

The synchronous demodulators in a colour television receiver provide the colour difference signals which must be combined with the luminance signal to obtain the red, green and blue signals to be applied to the display tube. The circuits used for the addition of the signals in the correct sense and proportion are generally referred to as matrix networks.

Transposing the equations of the E_I' and E_Q' signals in order to express the E_R', E_G' and E_B' signals in terms of E_I', E_Q' and E_Y' we obtain:

$$E_R' = 0{\cdot}96E_I' + 0{\cdot}63E_Q' + E_Y',$$

$$E_G' = -0{\cdot}28E_I' - 0{\cdot}64E_Q' + E_Y', \text{ and}$$

$$E_B' = -1{\cdot}11E_I' + 1{\cdot}71E_Q' + E_Y'.$$

In a system demodulating along the *I*, *Q* axes, the E_I', E_Q' and E_Y' signals must therefore be combined in the above proportions in order to obtain the desired red, green and blue outputs (Fig. 17).

It is necessary to ensure that the *Y*, *I* and *Q* channels delay the signal equally. As the *Q* channel is narrow-band, the *I* channel is moderately wide-band and the *Y* channel occupies the full video bandwidth, additional delay must be inserted into the *Y* channel, and to a lesser degree, into the *I* channel, before matrixing, in order to equalize the delays. For this purpose, a section of delay cable, or alternatively a lumped constant delay line, is inserted into the *Y* channel. For equalizing the *I* and *Q* delays, it is possible to arrive at a satisfactory compromise in the design of the low-pass filters following the synchronous demodulators.

In a system demodulating along the (*R-Y*), (*B-Y*) axes the matrixing process is simplified, as the demodulator outputs (after filtering through low-pass networks accepting frequencies up to 350kc/s) consist of the separate $(E_R' - E_Y')$ and $(E_B' - E_Y')$ signals, which only require the addition of the E_Y' signal for the production of E_R' and E_B'.

The E_G' signal may then be obtained by combining the E_R', E_B' and E_Y' signals using the relationship

$$E_Y' = 0{\cdot}30E_R' + 0{\cdot}59E_G' + 0{\cdot}11E_B',$$

whence $E_G' = 1{\cdot}7E_Y' - 0{\cdot}51E_R' - 0{\cdot}19E_B'$.

Various techniques for accomplishing the necessary summation are available[20], and have been discussed by Quinn[21] and Feingold[22].

One method is to use valves with a common anode load, a small variable amount of current feedback being used to compensate for differences between the valve parameters. Alternatively, a common cathode resistor may be used, giving a voltage gain per valve for a three-valve circuit of only 0·3, but with the inherent advantages of independence of valve parameters, and of good frequency and transient response consequent in the low impedance of the output circuit.

A more favoured arrangement is that of the feedback summing amplifier (Fig. 18) which is capable of precision matrix operation and only uses one valve irrespective of the number of inputs. For a circuit of sufficient gain, the output E_o is given[21] in terms of the inputs E_1, E_2, E_3 ......, the associated resistors R_1, R_2, R_3...... and the feedback resistor R_F by the relationship

$$E_o = R_F(E_1/R_1 + E_2/R_2 + E_3/R_3 + \ldots\ldots)$$

The values of R_1, R_2, R_3 .... include the source impedances, and it is therefore convenient to ensure that the latter are very low. The input impedance looking in from any source is simply its associated series resistance.

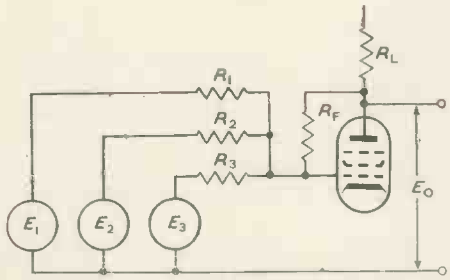


Fig. 18. A feedback summing amplifier

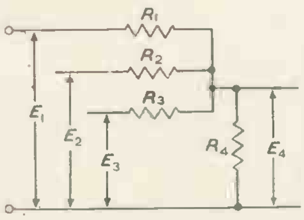


Fig. 19. A passive summing network

The cross-coupling between sources is generally below 1 per cent[21]. The frequency response, linearity and stability of the circuit are excellent.

The simple method of matrixing by a passive summing network, of the type shown in Fig. 19, has been commonly used. In terms of the admittances Y_1, Y_2, Y_3, Y_4 of the resistive arms R_1, R_2, R_3 and R_4, respectively (Fig. 18), the output E_o is given by

$$E_o = \frac{Y_1E_1 + Y_2E_2 + Y_3E_3}{(Y_1 + Y_2 + Y_3 + Y_4)}$$

Low impedance driving sources must be used, and it is common practice to feed the matrix circuit from cathode-follower stages. The value of R_4 must be kept low ($\sim 5\text{k}\Omega$) in order to obtain the necessary frequency response. The gain of passive matrix networks, limited by the necessity for reducing interaction between the sources and by the required frequency response, generally lies in the region 0·2 to 0·6. This is a lower figure than that obtained with the feedback summing amplifier, but for commercial receivers the elimination of one valve in each of the three colour circuits of an *I-Q* receiver using a passive network outweighs the slight advantage inherent in the use of feedback summing amplifiers. For laboratory or studio systems, however, the latter are likely to be favoured.

Matrixing at Sub-carrier Frequency

Recent work has demonstrated the possibility of matrixing before demodulation[23]. The chrominance signal may be divided into two paths containing wide band and narrow band filters, and the signals then recombined with suitable differences of phase and amplitude so that the E_I' component in the wide-band signal and the E_Q' component in the narrow-band signal are in the same phase and along either the (*R-Y*) or the (*B-Y*) axis. One colour demodulator is then used in each colour difference channel, one producing $(E_B' - E_Y')$ and the other $(E_R' - E_Y')$, for:

$$0{\cdot}36\,(E_B' - E_Y') = 0{\cdot}62E_Q' - 0{\cdot}4E_I' \text{ and}$$

$$(E_R' - E_Y') = 0{\cdot}62E_Q' + 0{\cdot}96E_I'.$$

We can thus obtain the $E_B' - E_Y'$ and the $E_R' - E_Y'$ signals by combining a common source of E_Q' with E_I' signals of both polarities and the correct amplitudes, provided that the requisite 90° phase shift is introduced between the *Q* and *I* channels. An advantage of this system is that the demodulators are then fed in phase from a single reference sub-carrier source.

An experimental circuit used for this purpose[23] is illustrated in Fig. 20. The common $0{\cdot}62E_Q'$ signal is obtained by suitable selection of the relative impedances of the

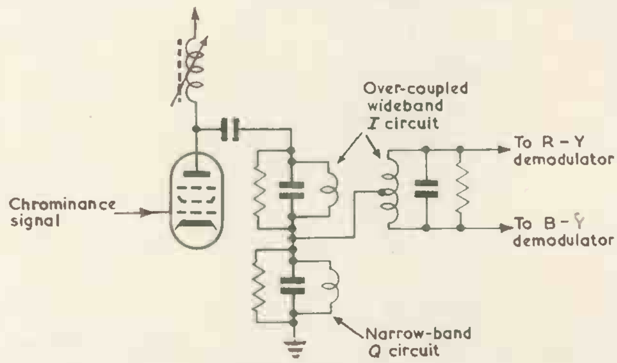


Fig. 20. Matrixing at sub-carrier frequency (after Hazeltine Corp.)

single-tuned narrow-band *Q* circuit and the over-coupled double-tuned *I* circuit. The tap on the secondary of the *I* transformer is positioned so as to obtain the $0{\cdot}96E_I'$ and the $0{\cdot}40E_I'$ signals, and the required 90° phase shift between *Q* and *I* is automatically obtained.

This circuit considerably reduces the number of valves and components in an *I-Q* receiver. However, it is necessary to compensate for the delay inherent in the narrow-band *Q* circuit relative to the *I* circuit.

Further experimental systems[23] have followed the lines described above, but with the *I* and *Q* passbands shaped independently and then recombined, using a delay line to obtain the desired phase and envelope delay characteristics.

It seems likely that, with further development, circuits of this type may be of considerable practical importance in the design of commercial high-quality *I-Q* receivers.

The Transient Response of the Intermediate and Video Frequeny Circuits

The i.f. amplifier in a colour receiver must be designed, as in a monochrome television receiver, to give the desired selectivity, not only to avoid the reception of unwanted signals, but also to prevent the occurrence of a beat between the chrominance sub-carrier and the sound carrier. The upper sideband of the *Q* signal extends to a video frequency of over 3Mc/s. It will also be seen that over the frequency range 2·3Mc/s to 3Mc/s, the frequency bands of the chrominance components overlap, so that to avoid picture degradation caused by the introduction of crosstalk between the *I* and *Q* channels, the circuit amplification and delay must both be constant. Small deviations from the ideal bandpass characteristic may cause a noticeable reduction in the transient response due to *I-Q* crosstalk[23,24,25]. Uniformity of delay is also needed in order to obtain coincidence of the luminance and chrominance contributions to abrupt transitions in colour. It is thus essential to design the amplifier stages for minimum phase distortion over the entire video band. The phase distortion

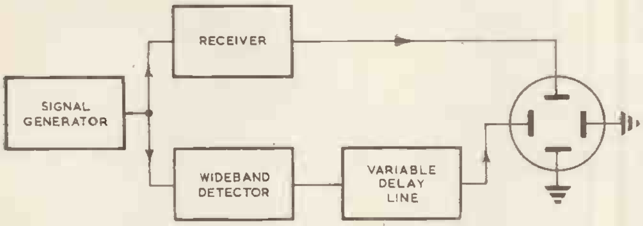


Fig. 21. Measurement of receiver phase delay

may be considered significant if the change in phase delay with frequency approximates to or exceeds the duration of a single picture element. Phase distortion up to 0·1μsec may be tolerated.

Curves showing the amplitude and phase characteristics of a flat-staggered quadruple have been given by Avins, Harris and Horvath[25], who have also stressed the importance of working in terms of phase delay rather than in envelope delay in investigations of the transient response of receivers.

The phase delay of a receiver may be measured (Fig. 21) using a variable delay line (of the same form as that described previously for the determination of phase detector characteristics) in conjunction with a wideband detector and a cathode-ray tube display with symmetrical horizontal and vertical amplifiers[25]. One of the amplifiers is fed from the receiver under test and the other from the controlled delay line, the latter being adjusted to equalize the phase at the output. The phase delay produced by the receiver, at the particular frequency, is then equal to that introduced by the delay line. Repetition of the procedure over the operating frequency range provides the required phase distortion information.

Acknowledgment

The author wishes to thank the Chief Engineer, Ultra Electric Ltd, for permission to publish this article.

REFERENCES

16. LIVINGSTON, D. C. Theory of Synchronous Demodulator as Used in N.T.S.C. Colour Television Receiver. *Proc. Inst. Radio Engrs. 42*, 284 (1954).
17. CLARK, E. G., PHILLIPS, C. H. Colour Demodulators for Television Receivers. *Electronics 27* ,164 (June, 1954).
18. ALTES, S. K. Determination of the Optimum Demodulation Angles in Colour Receivers. *Convention Record of the Inst. Radio Engrs.* Pt 7, 165 (1955).
19. ADLER, R., HEUER, C. Beam Deflection Tube Simplifies Colour Decoders. *Electronics 27*, 148 (May, 1954).
20. Waveforms, M.I.T Radiation Laboratory Series. 19, chap. 18 (McGraw Hill, 1949).
21. QUINN, W. M.. Methods of Matrixing in an N.T.S.C. Colour Television Receiver. *Convention Record of the Inst. Radio Engrs.* Pt 4, 167 (1953).
22. FEINGOLD, W. R. Matrix Networks for Colour Television. *Proc. Inst. Radio Engrs. 42*, 201 (1954).
23. Some Notes on the Sub-Carrier Matrix. Hazeltine Corporation. Rep. 7166R (1955).
24. KRONENBERG, M. H., WHITE, E. S. Design Techniques for Colour Television Receivers. *Electronics 27*, 136 (Feb., 1954).
25. AVINS, J., HARRIS, B., HORVATH, J. S. Improving the Transient Response of Television Receivers. *Proc. Inst. Radio Engrs. 54*, 274(1954).

A New Radio Simulator

The Mullard SL.21 Radar Trainer (RAF type 2292), developed in conjunction with the Ministry of Supply for the Royal Air Force, enables ground controllers of intercepting fighters to be trained on realistic targets without the expense of using actual aircraft and radar transmitters or dependence on favourable flying conditions. The trainer displays on the p.p.i. screen of a radar set up to 12 artificial echoes which can be controlled from dummy pilots' positions, the controller under training giving directions to the "pilots", following the trajectory of their echoes and guiding them to intercept other moving target echoes.

The pilot's control unit has controls for speed of flight up to 1 000 knots and rate of turn, rate ½ to rate 3 (90° in 10sec) in steps of ½. Indicators show direction of flight, "eastings", "northings" and a ground range of "aircraft" from the radar station of 3 to 200 nautical miles. The effect of a wind of up to 100 knots in any direction can also be simulated.

The trainer may be used in conjunction with a radar station so that live and artificial aircraft may be shown together on the screens.

The trainer consists of: the mixing rack, including the monitor p.p.i.; pilots' control units; computer racks. Each pilot's control unit controls the track of one target pip on the p.p.i.'s and one computer rack will accommodate up to 4 tracks, which are fed into the mixer rack. The position of the dummy aircraft on the display is indicated automatically in cortesian co-ordinates.

A voltage proportional to the setting of the aircraft speed control is applied to the leading sine/cosine potentiometer; the shaft of which is rotated at a speed determined by the setting of the rate of turn control, the actual shaft position being shown on the pilot's heading indicator. The output voltages from the sine/cosine potentiometer represent the velocity of the aircraft resolved along the East and North axes. These voltages are integrated in the X and Y integrators respectively giving signals which represent the position of the aircraft in rectangular co-ordinates. These voltages are applied to the azimuth triangle solver, which gives outputs corresponding to ground range and bearing. A voltage proportional to the ground range is applied to the range phantastron, triggered continuously by the radar trigger or the local trigger generator. The output from the phantastron is a pulse which is delayed with respect to the trigger pulse by a time proportional to range. These delayed pulses are fed to the azimuth gate which is opened when the bearing given by the azimuth triangle solver coincides with the bearing of the radar aerial or of the aerial simulator, the number of pulses passing through the gate depending upon the setting of the beam width control of the gate. These pulses, forming the simulated echo signal, are mixed with those representing other aircraft and then distributed to the monitor and other p.p.i. displays.

As well as supplying outputs to the azimuth triangle solver the X and Y integrators operate the northing and easting indicators in the pilot's control unit. To simulate the effect of wind on the movement of the aircraft, provision is made for voltages corresponding to the resolved wind vector to be fed into the X and Y integrators.

The pilot has a control for varying the intensity of the echo given by his target so varying the amplitude of the pulse fed to the p.p.i. displays, a calibration generator contained in the equipment giving the normal range marker rings.

The Short-Interval Timer

By G. Hitchcox, M.I.E.E., M.Brit.I.R.E.

Many industrial devices operate so rapidly that electronic time measurement is required, both in designing and in testing them. Examples are: fuses and contact-breakers, high speed relays, camera shutters, and explosive caps. This article describes the factors which govern the specification and design of a modern short interval timer intended for general purpose uses.

GENERALLY speaking, electronic techniques have only two advantages over their mechanical, magnetic, or physical counterparts: rapidity of response and potentially very high power gain. Rapid response follows from the low inertia of the only moving elements, the electrons themselves. Power gain can be high because grid current is incidental, not essential; the input stage of a high impedance d.c. amplifier, such as that in an electronic pH-meter, is capable of a power gain of ten thousand million times.

It is, of course, the first advantage of rapid time response that makes the electronic time interval meter so attractive. Before the last war, the field was monopolized by the mechanical stop watch and by instruments such as the "Synclock" in which a high speed electromagnetic clutch, controlled by the signal whose duration is to be measured, is the link between a constantly running synchronous motor and a gear-train connected to the timing indicator.

'Chronotron' measuring the acceleration of a car at speed

Both classes of instrument have both a proportional and a fixed error. The proportional error of a good watch can be as low as ±0·05 per cent; its fixed error depends upon the frequency of its escapement, falling from 100msec at the usual 10c/s to 10msec with the fastest escapement commercially available, 100c/s. The proportional error of a "Synclock" is that of the frequency of its motor supply; its fixed error, the variable difference between "on" and "off" operating times, can be reduced to a few milliseconds. The corresponding human error with trained recorders is about ±10msec.

The potentially high proportional accuracy of these instruments renders them admirably suitable for measuring fairly long intervals of one second or more. Their irreducible fixed error prevents their successful use below, perhaps, 0·1sec because of its intolerable proportional importance.

During the last war an urgent demand arose for the rapid, convenient, and fairly accurate measurement of much shorter intervals, of a few milliseconds or less. It arose in the development of very high speed predictors, correspondingly fast cameras, in the testing of explosives, and in ballistics generally. In projectiles, timing variations can appear in the fuse, in the gaine which links the fuse to the warhead, and in fragmentation itself. If a shell is travelling at several thousand feet per second, it is obvious enough that an operational error of a few milliseconds could well result in the explosion missing an aircraft altogether.

The first commercial timer appeared in 1943; laboratory models had preceded it by several years. It was hurriedly developed to meet an unusual demand. When the "Mosquito" fighter-bomber was designed, both its unprecedented speed, 100mile/h faster than any other service aircraft, and its wooden construction required the provision of knives along the leading edges of its wings, fitted to give it a chance of cutting the cables of any barrage balloons into which it might run. To make research and testing feasible, aerial conditions were simulated on the ground by mounting wings on either side of a trolley-car, rocket-propelled along a half-mile stretch of track, on each side of which lengths of cable could be suspended vertically. The car accelerated continuously to a terminal velocity of about 450mile/h and was then stopped in a few feet by a water buffer, certainly an awe-inspiring sight. Because of the very high acceleration it was essential to measure speed over a short distance by tripping contacts mounted only a few feet apart; to measure the correspondingly short time, the electronic timer had to be developed.

Counter and Integrator Systems

Electronic timers can be grouped into two classes: the pulse counter and the current integrator, both in their various forms. In general, other things being equal, the former is the more accurate and the latter the simpler and more convenient.

In counter instruments the output from a repetitive pulse generator of constant and highly stable frequency is fed through high speed switching circuits during the time interval to be measured to a counter system, which registers a total proportional to the duration of the interval. If the drive frequency is an exact multiple of ten, the counter can be directly calibrated in units of time.

As the electrical analogue of the stop watch, itself a counter, the electrical counter suffers from both proportional error and a fixed uncertainty interval. In a well-known commercial model, the drive frequency, which determines proportional accuracy, is 1Mc/s, controlled by a quartz crystal to within ± 1 part in 10^7. Fixed uncertainty, the interval between successive pulses, is necessarily the reciprocal of drive frequency, one microsecond. Each decade in the counter employs a scale-of-two followed by a scale-of-five; time presentation is singularly pleasing, the terminal reading of the counter appearing on individual meter scales, calibrated from 0 to 9, and one for each decade.

While the accuracy of integrator instruments varies widely with their design and construction, even at the highest figure commercially available, ± 1 per cent, it is well below that of the comparable counter, except when the control frequency is derived from the mains. The great merit of the integrator type is its simplicity; six or seven valves against perhaps from thirty to seventy, with the corollary advantages of lower cost, greater convenience, lower bulk and weight, the possibility of battery operation, and the increased reliability which should go together with simplicity. Both types have their place: the counter where maximal accuracy is essential, and the integrator where fair accuracy is sufficient and where cheapness and portability are appreciated. This article describes only the integrator type which, surprisingly, is the more recent development.

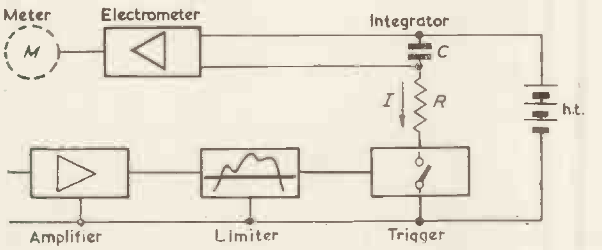


Fig. 1. Capacitance—integrator system

Basic Circuit

An integrator timer is shown in block form in Fig. 1. The trigger circuit is preceded by a high gain d.c. amplifier and a limiter stage, included to permit the acceptance of wide variations in amplitude of control signal without affecting accuracy of timing. The trigger circuit acts as a high speed electronic switch which allows the timing current to flow only for the duration of the amplified control signal. The magnitude of the timing current is separately controlled by a network R; it can be resistive as suggested in the diagram or it can be an electronic constant-current device such as the anode-cathode circuit of a cathode-follower, to make possible a strictly linear meter scale. Timing current is integrated by a precision low-leakage capacitance C and the voltage developed across this capacitor, proportional to charge and hence to the time interval to be measured, is fed to an electrometer d.c. amplifier and displayed on a meter directly calibrated in units of time.

In such a system a number of stage combinations are possible: the amplifier with the limiter, or the limiter with the trigger device, or the trigger with the control network. In the practical instrument described later, limiter and trigger are combined in an electronic switch unaffected by control amplitudes above a low minimum level, while input amplifier and current control circuits are separate.

The basic theory of an integration timer is simple; how satisfactory a design is in practice depends very much upon details of technique and upon careful selection of component values. The general character of the design is largely determined by the mechanism adopted to select the ranges.

If time intervals are to be measured over the whole useful range, from a fraction of one millisecond up to, say ten seconds, this total coverage must be split into a number of separate ranges, large enough to permit easy reading well up the meter scale, and not too large, to avoid confusion. Eight is a reasonable number, the shortest from 0 to 4msec, the next from 0 to 10msec, then from 0 to 40msec, and so on up to the longest range, from 0 to 10sec. In this way the greatest interval between successive ranges is 4 to 1 and the ratio of the longest to the shortest is 2 500 to 1.

The fundamental equation governing a capacitance integrator is, of course:

$$v = q/C = It/C$$

where q is the charge inserted into the integration capacitance C by a current I flowing for the time interval t and where v is the voltage developed across the capacitance and measured by the electrometer amplifier. Relating the electrical quantities to a time interval t_{fs} intended to be the full-scale reading of any one range gives:

$$t_{fs} = CV_{fs}/I$$

where V_{fs} is the electrometer input to produce full-scale deflexion of the meter. It might seem that range coverage of 2 500 to 1 could be provided by varying any or all of the three quantities on the right-hand side of the equation, but in practice two have sharply optimum values and it is usual to vary only the third, the timing current I.

This point needs clarification. If the integration capacitance C is too large, then the amplitude of timing current required to charge it on the shortest range can be inconveniently high. If it is too small, decay of charge and therefore of reading when a timing operation has been completed, can be so severe that effective accuracy is lost. Rate of decay is determined for all ranges by the product CR_p, where R_p is parallel leakage resistance across the capacitance together with valve, wiring, and switching. A possible minimum value for R_p on a very damp day is 1 000MΩ. A suitable compromise value for C can be 4·0μF; with this value, maximum timing current, on the shortest range, becomes 12mA, and the product CR_p becomes 4 000sec, implying a reading decay of $1\frac{1}{2}$ per cent each minute. This is perhaps as good a compromise as can be obtained.

The figure of 12mA for the shortest range assumes a full-scale voltage drop across the timing capacitance of 12V. This value is determined by the probable zero instability of the electrometer d.c. amplifier against changes in mains voltage; a normal zone of drift for a properly designed single valve bridge circuit is $\pm$50mV. To "swamp" this drift to not more than $\pm\frac{1}{2}$ per cent of full-scale requires a full-scale input of 10V, obviously enough; it cannot be made much more than this because of increasing difficulty in acceptance. A fully symmetrical voltmeter circuit and thorough power stabilization could reduce instability and hence minimum input by perhaps twenty times, but at too great a cost.

It will be understood that the precise choice of some quantity values, such as 12V for amplifier sensitivity, is dictated by the need to use standard components, rather than by technical nicety.

From this wearisome but essential prelude it appears that both the value of the integration capacitance and the sensitivity of the electrometer voltmeter must be regarded as invariable outside fairly close limits, and that it is by variation of timing current I that range coverage over 2 500 times must be provided. The importance of proper choice of operational values has been stressed; they are summarized below.

Integration capacitance $C = 4{\cdot}0\mu F$ for all ranges
Electrometer sensitivity $V_{fs} = 12{\cdot}0V$ for all ranges
Integration quantity $q_{fs} = 48\mu C$ for all ranges

The last equation is derived from the first two. Knowing this quantity, corresponding to full-scale of any range, and having defined what these ranges are to be, the timing currents can be derived:

0 – 4msec	$I = 12mA$	0 – 400msec	$I = 120\mu A$
0 – 10msec	$I = 4{\cdot}8mA$	0 – 1sec	$I = 48\mu A$
0 – 40msec	$I = 1{\cdot}2mA$	0 – 4sec	$I = 12\mu A$
0 – 100msec	$I = 0{\cdot}48mA$	0 – 10sec	$I = 4{\cdot}8\mu A$

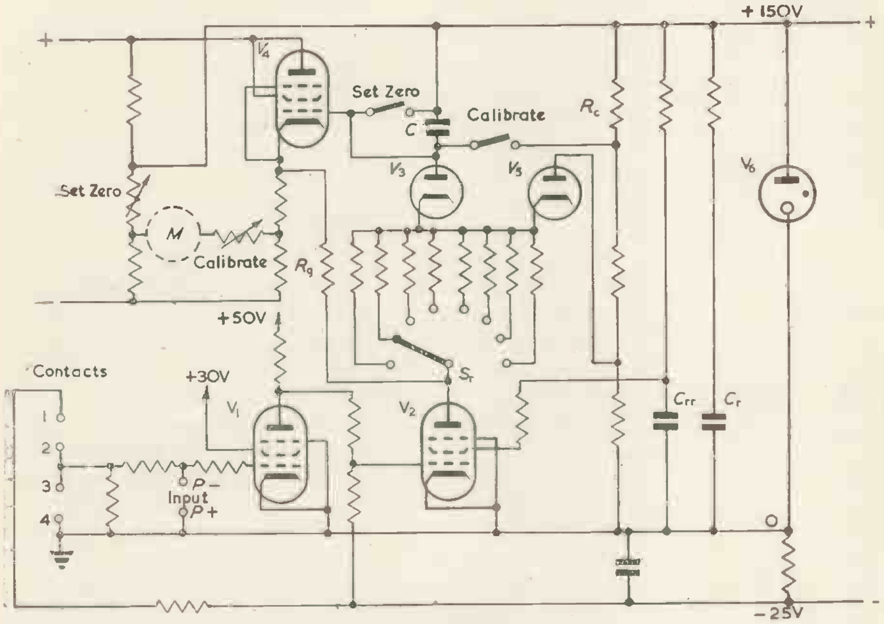


Fig. 2. Practical instrument

From these figures it follows that reading decay should not exceed 0·025 per cent per second and that zero drift due to electrometer instability should not exceed $\pm 0{\cdot}5$ per cent of full-scale deflexion.

The Practical Instrument

The circuit of a commercial instrument appears in Fig. 2. At the bottom left is the input amplifier with a stage gain of about 30. The timer can measure either the duration of an externally generated signal, fed to the terminals *P-P*, or it can measure the duration of opening or closure of a pair of contacts, or the interval between the successive operation of any combination of two pairs. Here terminals 1-2 and 3-4 are used, together with an internally generated negative test voltage derived from the h.t. supply.

With no input signal, valve V_1 is fully conductive, its anode nearly at earth potential, and trigger valve V_2 is cut off. The (negative) input signal raises the anode potential of V_1, overcomes the hold-off bias on V_2, and allows timing current to flow. Valve V_2 is fully operated by a change of + 15V, any excess being absorbed by grid current and having no further effect; since the amplifier gain is 30, minimum signal input to the instrument is $15/30 = 0{\cdot}5V$, negative to earth. Threshold level, below which input signals are too small even when amplified to affect V_2 at all, is about 0·2V.

Resistive control determines the amplitude of timing currents, with a separate high stability control resistor for each range. Electronic constant-current devices could have been used to permit absolute linearity in meter calibration; their disadvantage is the need for individual "fine" adjustment of current each time the range is changed, a result of their imperfect stability. With high stability resistors and an accurately controlled h.t. supply, after one initial adjustment common to all ranges, only operation of the range switch is required.

With resistive control, the timing current necessarily falls exponentially as the voltage across the timing capacitor rises. However, if the maximum "lost" voltage is small relative to that of the h.t. supply, visible non-linearity can be slight. In this circuit, the h.t. voltage is 150, while the full-scale drop across capacitor *C* is 12V; the total fall in current is therefore not more than 8 per cent below its initial value. It can be corrected by distorting the meter scale, maximum departure from linearity being 2 per cent at mid-scale, which is hardly perceptible.

Timing current flows through valve V_2, through the selected control resistor, through the diode V_3, and into the timing capacitor *C*. The voltage developed across it, proportional to the time interval, is fed to the electrometer voltmeter; V_4 is a triode-connected pentode, its grid current being less than $5 \times 10^{-10}A$. Both Set Zero and Calibration controls are provided, their adjustment common to all ranges. The main h.t. supply is stabilized by a 150V high-stability reference tube; on the shortest time range its current output is supplemented by the large reservoir capacitance C_r. In the same way, the screen supply to trigger valve V_2 is largely derived from capacitor C_{rr}.

Both diode valve V_3 and guard resistor R_g are essential to the system. With no signal input, trigger valve V_2 is cut off; the insulation resistance across its base cannot, however, be perfect. If it were as low as 1 000MΩ, with an h.t. supply of 150V a leakage current of $0{\cdot}15\mu A$ could flow through it and therefore into the timing capacitance $C = 4\mu F$. This builds up a signal at the rate of 2·25V/min, appearing as a wholly intolerable zero drift.

Such leakage can be heavily reduced by the arrangement shown in Fig. 3. The upper end of the guard resistor R_g is taken to the cathode of voltmeter valve V_4, which "tracks" its grid, and is always about 1V positive to it and therefore to the anode of diode V_3. While R_g can be high, perhaps 1MΩ, it is still minute compared to the minimum probable leakage resistance R_v across the trigger valve V_2 when it is cut off. It follows that in the quiescent (no signal) phase, very little voltage is dropped across R_g and that the potential of the cathode of diode V_3, to which the lower end of R_g is connected through the control resistor, is also held at about +1V to its anode.

The effect of the arrangement is to substitute a leakage path where 1V is applied across the base resistance of V_3 for one where 150V is applied across V_2, a 150 to 1 improvement . The presence of the guard resistor R_g does not affect accuracy in timing since, although a fairly heavy current of $150\mu A$ then flows through it, this current is derived directly from the h.t. source and does not enter the timing capacitance C.

Accuracy of Calibration

The accuracy of an instrument such as this is affected by changes in the integration capacitance C, or in the electrometer voltmeter sensitivity, or in the magnitude of the timing currents; these last are determined both by

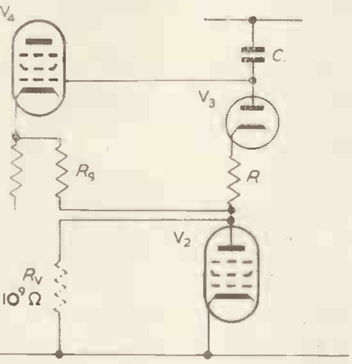


Fig. 3. Guarding circuit to minimize effect of leakage path across trigger valve V_2

EIL 'Chronotron'

overall h.t. voltage and by the values of the individual control resistors. It is happily quite practicable to incorporate timing capacitors and control resistors of adequate initial precision and long-term stability, certainly remaining accurate within $\pm 0{\cdot}75$ per cent over a wide range of ambient temperatures. But since both h.t. voltage and electrometer sensitivity depend upon the characteristics of electronic valves, some means of checking and adjusting their values is required.

If a voltage derived from the h.t. supply and developed across the calibration resistor R_c is injected into the voltmeter, its magnitude can be so chosen that the electrometer indicates full-scale, or some other convenient reference value, when both h.t. voltage and sensitivity are at their design-centre norms. If either should subsequently change, the calibration reading can be returned to its proper value, perhaps by adjusting amplifier sensitivity only, as shown in Fig. 2. Whatever change produced the shift in reading would have affected the accuracy of timing to the same extent, a change in h.t. voltage by modifying amplitude of timing current, and a change in electrometer amplifier sensitivity obviously enough by varying the sensitivity of display. It follows that re-adjustment of the calibration control can be used to correct any errors other than those due to changes in the control resistors or in the timing capacitance itself, both in practice most unlikely.

To avoid significant dependence upon valve parameters, it is essential that, in operation when timing current flows, the voltage dropped across the anode-cathode circuits of V_2 and V_3 should always be very low. Such potential drops are largely determined by proper choice of valves and electrode supplies. On the shortest range, where the timing current is highest at 12mA, the drop across V_2 is about 2·8V; the unusually high screen voltage helps to keep it low. For diode V_3 the drop is about 1·9V for the same timing current. On the longer ranges where the timing currents are much lower, the total drop across the electronic and therefore variable part of the chain is insignificant; on the longest range it can actually reverse in sign as contact potential overrides voltage drop due to the passage of current.

Overloads in Time

Overloads in time cannot damage the meter; in operation, the grid of the electrometer valve V_4 is driven negative to its cathode and the circuit is so arranged that even at the cut-off point, beyond which further signals have no effect, current through the meter is well below the destruction level.

A second diode V_5 is so connected that it becomes conductive if the voltage across the timing capacitor exceeds 18V, an overload of 50 per cent beyond full-scale; such conduction locks the system to prevent any further rise. This is not an additional precaution to protect the meter; it is to avoid the long discharge interval which would be required if a prolonged overload could drive the voltage across capacitor C up to 150V. Discharge time would be long because of dielectric hysteresis, which can maintain a significant fraction of the original excessive charge even after C has been short-circuited for one minute; the effect, of course, is a slow upwards creep of zero level.

Choice of Valves

The amplifier valve V_1 should be capable of high voltage gain to avoid an inconveniently high minimum input signal. The electrometer valve V_4 must have very low grid current, which can cause zero drift, and it must have high mutual conductance to permit adequate degeneration and therefore high stability in voltmeter circuit sensitivity. For both applications the EF37A is suitable.

The trigger valve V_2 should combine a short grid base, sharp cut-off, and the ability to pass heavy momentary currents with a low anode to cathode voltage drop, this last to minimize dependence of timing current amplitudes upon variable valve parameters. All glass construction is desirable to provide high anode to cathode leakage resistance; the EF95 or 6AK5 is an appropriate type.

Since the cathodes of the two diode valves V_3 and V_5 are at the same potential, a combination double-diode such as the EB91 or 6AL5 can be used, all-glass construction again being an advantage. In no case is the operation of this instrument affected by ordinary changes during the life of any valve or by its replacement. Changes in main supply are unimportant since their effects are covered by the calibration procedure; they are in any case minimized by the use of a stabilized h.t. supply.

Acknowledgment

This article is based upon the "Chronotron", a commercial instrument designed for and manufactured by Electronic Instruments Limited, and thanks are given to that Company for granting permission to describe it.

An Improved Type of Differential Amplifier

By J. C. S. Richards*, B.Sc., Ph.D.

A differential amplifier stage capable of giving a high rejection ratio with unselected valves and components and without a balance control is analysed, and a particular amplifier is described in some detail. The stage is particularly suitable for converting balanced to unbalanced signals.

CONSIDERABLE attention has been given to the design of differential amplifiers—that is, amplifiers which have two ungrounded input terminals and which give zero or negligible output when like signals are applied between each of the input terminals and earth. The ability of a differential amplifier to discriminate between like and unlike input signals is expressed in terms of its rejection ratio, defined more precisely below. To obtain high values of rejection ratio with most designs of differential amplifier, it is necessary either to use specially selected valves and components, or to have some form of balancing control.

It is shown below that by means of a feedback circuit, it is possible to make a differential amplifier having an inherently high rejection ratio without the use of selected valves or a balance control. The addition of a single balance control makes it possible to attain an effectively infinite rejection ratio. The high rejection ratio is maintained over a wide band of frequencies. The output of the amplifier can be push-pull or single-ended.

Performance Parameters

If a differential amplifier has a push-pull output its performance may be described in terms of the differential gain M_d, the in-phase gain M_i, and the inversion gain M_v. (Johnston[1], Parnum[2]). If the input signals are e_1 and e_1', and the output signals are e_2 and e_2' (all measured with respect to earth), these parameters are defined as follows:

$$M_d = (e_2 - e_2')/(e_1 - e_1') \text{ for } e_1' = -e_1$$

$$M_i = (e_2 + e_2')/(e_1 + e_1') \text{ for } e_1' = e_1$$

$$M_v = 2(e_2 - e_2')/(e_1 + e_1') \text{ for } e_1' = e_1$$

If a push-pull output is used, the rejection ratio is M_d/M_v and the value of M_i is comparatively unimportant. However, if the early stages of the amplifier have a small value of M_i, the design of later stages is greatly simplified. In addition, the simplest way to obtain a single-ended output is to design a push-pull amplifier for which the overall value of M_i is very small, so that the differential properties of the amplifier are retained if the output signal is taken between one of the output terminals and earth.

Standard Techniques

The conventional type of differential amplifier stage is shown in Fig. 1. For this stage we have:

$$M_d = \frac{\mu R}{R + r_a} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(1)}$$

$$M_i = \frac{\frac{1}{2}(\mu_1 - \mu_2)(R_1 - R_2)R_3 + \mu R(R + r_a)}{(R + r_a)\left[R + r_a + 2(\mu + 1)R_3\right]} \quad \ldots\ldots \text{(2)}$$

$$M_v = \frac{r_a^2(g_{m1}R_1 - g_{m2}R_2) + (\mu_1 - \mu_2)R(R + 2R_3)}{(R + r_a)\left[R + r_a + 2(\mu + 1)R_3\right]} \quad \text{(3)}$$

These expressions are slightly simplified versions of the exact formula. R, r_a and μ are the mean values of $R_1 + R_2$, r_{a1} and r_{a2}, and μ_1 and μ_2 respectively. They have been used where the exact values are not critical.

When R_3 is made large, both M_v and M_i become small, although neither tends to zero. Now for the whole amplifier, the value of M_i may be made very small by cascading several stages. On the other hand, M_v must be made small for each stage where the in-phase input signal is appreciable (in practice generally the first stage only) since a large value for M_v means that an appreciable anti-phase signal is passed on to later stages. The main problem, therefore, is to make M_d/M_v for the first stage sufficiently large.

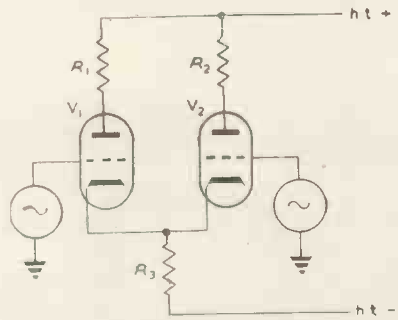


Fig. 1. Simple differential amplifier stage

M_v depends on the balance between the parameters of V_1 and V_2 and between R_1 and R_2, but if R_3 is made very large M_d/M_v becomes approximately $(\mu + 1)\,\mu/(\mu_1 - \mu_2)$, which is large if μ is large even when $(\mu_1 - \mu_2)$ is appreciable. In addition, μ is generally the most constant of the valve parameters, while g_m changes appreciably when the valve ages, or the heater voltage is varied. To obtain a large value of μ, V_1 and V_2 must be pentodes with the screen to cathode potential fixed, and arranged so that the screen currents do not flow through R_3. Fig. 2 shows a commonly used circuit, in which the anode slope resistance r_{a3} of V_3 provides a large common cathode resistance. If r_{a3} is assumed to be infinite, then:

$$M_d/M_v \simeq \frac{(\mu' + 1)\mu'}{\mu_1' - \mu_2'}$$

where μ_1', μ_2' are the inner amplification factors of the valves.

μ' has been written as before for the average value of μ_1' and μ_2'. This equation is derived on the assumption that the anode voltage does not affect either the screen or anode current of the valve, and that the ratio of the anode to the screen current is constant. It predicts a slightly larger value of M_d/M_v than the exact equation. Since the inner amplification factor of a pentode is relatively low, the value of M_d/M_v is small unless μ_1' and μ_2' are nearly equal.

The screen to cathode potential of a pentode valve may be held constant by connecting a large capacitor between screen and cathode in the circuit of Fig. 2. However, the effective resistance in the cathode circuit now becomes r_{a3}

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in parallel with R_4, which is comparatively small, so that M_v no longer depends mainly on the difference between μ_1 and μ_2, and M_i becomes relatively large. Andrew[3] has overcome these difficulties by using the circuit of Fig. 3 for which the rejection ratio is commonly greater than 10 000. The circuit of Fig. 3 has two practical disadvantages; a floating battery is required and, unless elaborate precautions are taken, the high anode impedance of V_3 is shunted by various leakage resistances and stray capacitances. Since the heater to cathode impedance of most valves is relatively small, the heater supply to V_1 and V_2 must have a low resistance and, what is more difficult to achieve in practice, a low capacitance to earth.

The Feedback Circuit

Fig. 4 shows a circuit which has none of the disadvantages of that shown in Fig. 3, and in several respects gives a better performance. Its operation may be briefly explained as follows. The essential function of a large value of cathode load is to maintain constant the total cathode

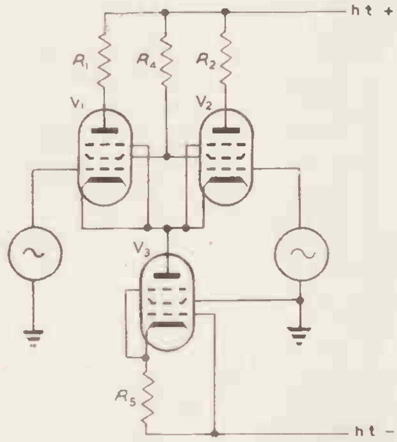


Fig. 2. Conventional differential amplifier stage using pentodes

current, and hence the mean anode potential of V_1 and V_2. In the circuit of Fig. 4 this action of the common cathode resistance is assisted by feeding to the grid of V_3 a potential equal to the mean anode potential of V_1 and V_2. This negative feedback loop tends to stabilize the mean anode potential of V_1 and V_2. Although this type of feedback is not new[4], its possibilities do not appear to have been fully explored.

At frequencies where capacitive impedances may be neglected, and if $R_7 = R_8$, it may be shown that the parameters of the circuit M_d, M_v and M_i are given by equations (1), (2) and (3), except that R_3 is now replaced by $R' \, (1 + \frac{1}{2} g_{m3} R)$, where R' is the impedance of R_4 and r_{a3} in parallel. Since $g_{m3}R$ may readily be made 50 or 100, the effective cathode load is now very high, and at least as large as r_{a3} alone. It may also be shown that if R_7 differs by only a small amount (say 20 per cent) from R_8, the operation of the circuit is not appreciably affected. Not only is the floating battery eliminated, but any stray impedance between the cathode or screen of V_1 and V_2 and earth need be greater than R' only. In particular, an earthed heater supply can be used for V_1 and V_2.

If we make the reasonable assumptions that $(\mu + 1) \simeq \mu$, $g_{m3}R \gg 2$ and $\mu_1 \, g_{m3} R R' \gg (R + r_a)$, then we can write:

$$M_v / M_d \simeq \frac{g_{m1}R_1 - g_{m2}R_2}{g_m R} \cdot \frac{1}{g_m R g_{m3} R'} + \frac{\mu_1 - \mu_2}{\mu} \cdot (1/\mu)$$

Typically we may take $g_m R g_{m3} R'$ as about 10 000 and μ as about 2 500 so that if $R_1 = R_2$ we find:

$$M_v / M_d \simeq (1/2\,500) \; [(\Delta g_m / 4 g_m) + (\Delta\mu / \mu)]$$

where $\Delta g_m = g_{m1} - g_{m2}$ and $\Delta\mu = \mu_1 - \mu_2$. Hence it would be expected that the rejection ratio of this circuit would be at least 2 500, and that the effect of valve ageing etc. would be small.

Practical Circuit

Fig. 5 shows a practical circuit embodying the above principles. The feedback voltage for the first stage is conveniently taken from the cathodes of V_4 and V_5. The in-phase gain of the first stage is approximately 1/60, and of the second stage 1/3, so that the output signal may be

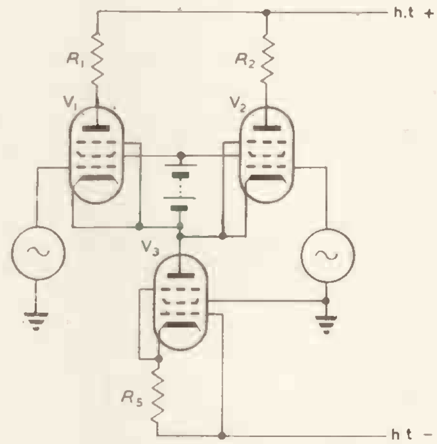


Fig. 3. Andrew's differential amplifier stage

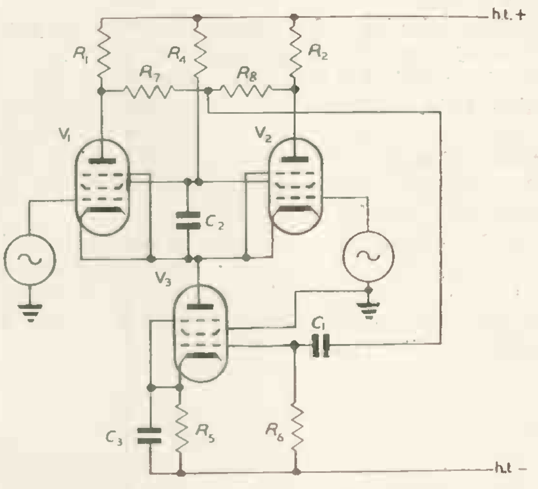


Fig. 4. Differential amplifier stage with additional feedback

taken either between the anodes of V_4 and V_5 or between one anode and earth, without appreciably affecting the rejection ratio. The anti-phase gain is over 3 000 for frequencies between 5c/s and 10kc/s. To check the effect of valve parameters on the discrimination ratio a random sample of eight EF37A's was taken, including both new and aged valves. The inversion gain was measured for an input of 1V r.m.s. at 500c/s. The corresponding rejection ratios are given in line 1 Table 1.

Most of the low values of rejection ratio given in line 1 of Table 1 arise from the presence in the sample of one valve with anomalous characteristics. If that valve is removed from the sample, the rejection ratios become as shown in line 2. Almost half the valve pairs then give rejection ratios of more than 5 000, more than a third give

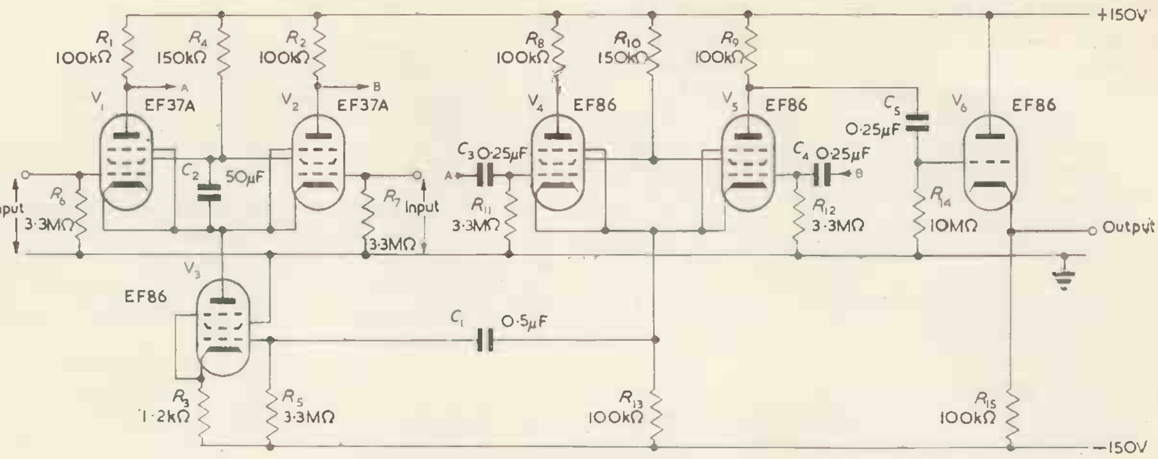


Fig. 5. Practical differential amplifier stage

V_6 EF86, triode connected. R_1, R_2 are high stability carbon resistors.

TABLE 1

REJECTION RATIO	1 000–2 000	2 000–5 000	5 000–10 000	10 000–20 000	OVER 20 000	TOTAL
Number of valve pairs	1	17	2	3	5	28
Number of valve pairs	0	11	2	3	5	21

ratios of over 10 000, and almost a quarter give ratios of over 20 000.

Balance Control

Although rejection ratios adequate for most purposes can be obtained with only moderate care in selecting valves, it is sometimes desirable to have a means for increasing the rejection ratio. Only at relatively low rejection ratios (<5 000 say) is it likely to be true that the first term on the right-hand side of equation (3) is unimportant. If either R_1/R_2 or g_{m1}/g_{m2} is made variable, effectively infinite rejection-ratios can be obtained.

It is found experimentally that if the ratio R_1/R_2 is adjusted by means of a suitably connected potentiometer so as to give a very high rejection ratio, then this high ratio is not maintained when the heater voltage supplied to the amplifier is varied appreciably.

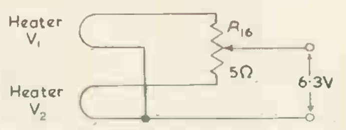


Fig. 6. Optional balance control for circuit of Fig. 5

The ratio g_{m1}/g_{m2} may be varied by means of a potentiometer connected so as to vary the relative heater voltages of V_1 and V_2. By this technique, due to Aitchison[5], the characteristics of the valves can be almost exactly matched, and the balance is maintained over a wide range of supply voltages and input signals. With the control shown in Fig. 6, the rejection ratio may be made greater than 50 000 for most valve pairs, and this ratio does not fall below 10 000 for heater voltage changes of ± 10 per cent. At rejection ratios greater than 50 000, the output is almost wholly second harmonic.

Frequency Response

The particular amplifier shown was designed for use with a high impedance bridge network operating over the frequency range 20c/s to 10kc/s, and the performance figures quoted for 500c/s hold good over this range. The anti-phase gain and rejection ratio fall by 3dB at about 25kc/s. The high frequency range could easily be extended by a factor of at least five by using valves of higher mutual conductance (e.g. Mullard EF91, EF95) and lower values of anode and screen resistors. The anti-phase gain falls 3dB at about 0·5c/s, but this figure could easily be lowered to 0·1c/s (say) simply by increasing the time-constants of the coupling network.

The rejection ratio increases by a factor of between 2 or 3 at 10c/s, but does not deteriorate by a factor of more than 10 for frequencies as low as 1c/s. The exact performance depends on the balance of the valves and their associated circuits. For many applications, the amplifier is required to discriminate against mains interference, and a fall in rejection ratio for frequencies lower than 50c/s is unimportant. Although some improvement can be attained by increasing the various capacitors in the circuit, for operation down to very low frequencies it is probably better to use direct coupled amplifier techniques.

Miscellaneous Details of Performance

The rejection ratio measured at an in-phase input signal of 1V r.m.s. is fairly well maintained for input signals up to about 5V r.m.s., but decreases markedly for signals greater than 10V r.m.s. The exact figures depend to some extent on the working range over which the valves remain balanced.

The circuit is relatively insensitive to supply voltage changes, and to the impedance of and ripple in the h.t. lines, provided the impedance between the negative h.t. line and earth is not too large (< 1kΩ say). The performance given was obtained by using simple gas tube stabilized supplies.

Applications

The particular circuit described was intended for use with a high impedance bridge network, eliminating the need for a screened and balanced transformer, and giving a single-ended output to feed a tuned amplifier and detector. For most applications there is no reason why miniature valves should not be used throughout. In the present instance EF37A's were used because the top-cap grid connexions were convenient in the layout adopted, and because octal based valves were less liable to damage when repeatedly inserted and removed while the effect of valve matching was being observed.

With slight modification (e.g. increasing C_3R_{11} and C_4R_{12} to about 3sec each) the amplifier should prove suitable for many biophysical applications.

The ability of the feedback circuit to give low in-phase gain in a single stage makes it useful in applications where a push-pull to single-ended stage is required. Such a stage, with the anti-phase gain stabilized at unity, has been designed to enable commercially made measuring instruments with a single-ended input to be used with a push-pull input. It is hoped to describe this device in a separate article.

REFERENCES

1. Johnston, D. L. Electro-Encephalograph Amplifier. *Wireless Engr. 24*, 231 (1947).
2. Parnum, D. H. Transmission Factor of Differential Amplifiers. *Wireless Engr. 27*, 125 (1950).
3. Andrew, A. M. Differential Amplifier Design. *Wireless Engr. 32*, 73 (1955).
4. Communication from E.M.I. Laboratories. Balanced Output Amplifiers of Highly Stable and Accurate Balance. *Electronic Engng. 18*, 189 (1946).
5. Aitchison, R. E. A New Circuit for Balancing the Characteristics of Pairs of Valves. *Electronic Engng. 27*, 224 (1955).

A Periscopic Television Unit

A new type of periscopic television equipment has been installed for furnace observation at Barking "C" power station. Developed by Babcock & Wilcox Ltd, in conjunction with the British Iron and Steel Research Association, Marconi's Wireless Telegraph Co. Ltd, and the Foster Instrument Co. Ltd, the unit incorporates entirely new optical principles enabling the whole of the burner wall and all burners to be observed simultaneously from a single position, and televised to any part of the station. With furnaces of the latest type, such as are installed at Barking "C", it is essential that the jet-injected pulverized fuel remains ignited at all burners, otherwise a serious explosion might result. This has hitherto been ensured by constant observation through small inspection doors, but modern boilers have now become so tall that a complete tour of inspection could take half an hour and involve climbing more than 100ft of galleries and ladders. The furnace wall aperture required is only four inches in diameter, and the unit has been developed to suit pressurized furnace operation, so that it can be inserted or withdrawn with ease and safety against furnace pressures of up to 20in water gauge. It can be left in position continuously, if desired, and has many other interesting features including automatic or semi-automatic "turn down" operation which renders the camera tube quiescent until a picture is required.

The periscope, which is completely water-cooled, with incorporated air-blast to keep fly ash from settling on the lens windows, is mounted, together with its associated television camera, on a carriage which can be advanced into the furnace or retracted by means of a remote-controlled air cylinder. Twin-entrance apertures are used, with lenses and prisms so arranged that the two fields of view are combined in panoramic form, with a dividing line at the centre of the picture. A field of 90° in one plane and 45° in the other plane is thus obtained, permitting the observation of the complete burner wall from a position above the burners. If applied to a pressure-operated furnace, the nose of the periscope water jacket is fitted with a conical fairing to avoid interference with the flow of air in the air-blast door through which the unit enters. This takes the form of a circular opening with gas-tight flaps that are pushed open by the nose of the periscope as it advances. At the same time, high-pressure air is automatically admitted to a ring of nozzles which converge on the opening surrounding the doors. The velocity of these air jets is sufficient to induce air into the furnace and form a "curtain" which prevents any tendency towards the emission of gas or flame.

The television apparatus is the industrial type BD835 as developed by Marconi's Wireless Telegraph Co., Ltd. and comprises two units, the Camera and the Control Unit.

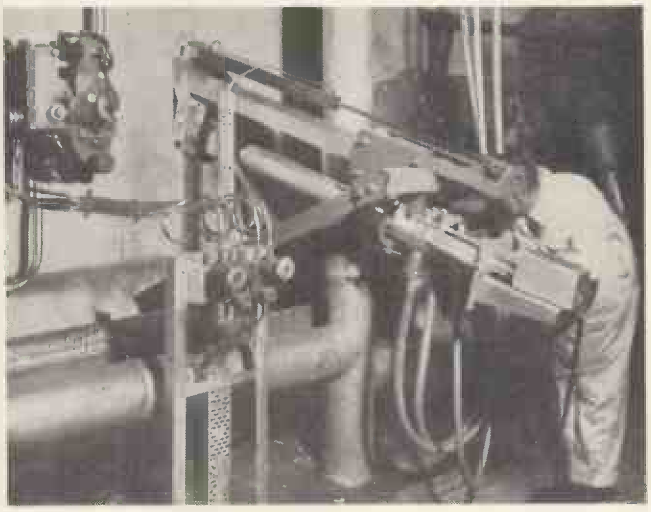
The Camera

A Vidicon pick-up tube is employed, the signal from which is capacitance-coupled to a three-stage video amplifier and then fed to a mixer stage. In this circuit the outputs of the vision and sync pulse circuits are combined, and the resultant signal further amplified. This combined signal is then fed through a sync clipper stage which keeps the horizontal and vertical sync pulses equal in amplitude regardless of varying signal conditions, thereby maintaining stable monitor synchronization.

A grounded cathode "Hartley" oscillator circuit provides any frequency between 54-88Mc/s, which is modulated by the vision signal. The modulated signal is then fed, by coaxial cable, to the Control Unit.

The Control Unit

This contains the horizontal and vertical deflexion circuits for the pick-up tube, and a circuit which prevents burning of the video target coating in the event of scanning failure. In addition, it includes circuits for providing the necessary waveforms for vertical scan, horizontal scan, vertical and horizontal blanking and sync pulses.



The periscope unit about to enter the furnace

The periscope unit is normally used only when lighting-up operations are in progress. When the furnace is operating under stable conditions, the unit is withdrawn, and monitoring is taken over by a photo-electric cell. Should the overall brightness of the furnace decrease, the photo-electric cell operates an alarm on the control panel; at the same time the periscope nose is automatically advanced into the furnace, and a picture obtained to show which burner has been extinguished.

A Variable Multiple Pulse-Stream Generator

By W. Woods-Hill*

During the development of circuits intended for use in electronic computers there is a need for checking, with a minimum of apparatus, the logic of circuits which require numerous pulse streams for their operation.

This article describes such an apparatus using an electrostatic pick-up previously described[1,2].

IT was noticed that a large part of engineers' time was devoted to designing and building complex pulse stream generators for testing discrete computer circuits. These generators were often as complex as the section of machine they were to supply; further, each one had to be tailor made to suit whatever section of a proposed machine (arithmetic unit, division sub-routine, etc) was to be tested next.

The obvious need was for a piece of apparatus capable of altering the pattern and number of pulse streams it emitted by means of switches or other external controls.

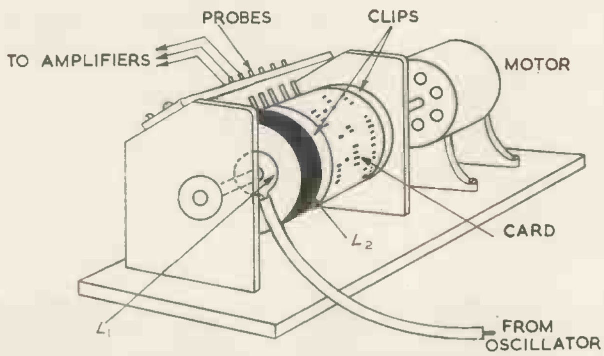


Fig. 1. Motor-driven drum mechanism

Because the requirements of yet undesigned equipment are virtually unpredictable, the pulse patterns obtained must be completely flexible within the capability limits of the apparatus and, of course, quickly alterable from one pattern to another.

If these requirements were to be interpreted in a purely valve and switched electronic form, the bulk of equipment involved would be far in excess of the tolerable maximum.

For this reason, a combined mechanical and electronic device was designed which has, as a result of the mechanical component, a speed a good deal less than the ideal, but in view of the reduction in cost and that the aim was to test the logic more than the electronic limitations, had quite satisfactory characteristics.

Description

As shown diagrammatically in Fig. 1 the apparatus consists of a motor driven drum revolving at high speed, to the surface of which are approached 11 metal probes. The drum is provided with two clips capable of holding in place a card wrapped round the periphery. This card is coated on one side with thin metal foil which makes electrical contact with the holding clips.

The card, which is of the type used in Hollerith punched card accounting systems (plus metal backing), is punched with holes at positions representative of the desired pulse stream coding.

The metal backing to the card is connected, via the clips, to a radio frequency coil L_2 wound round one end, and rotating with the drum. This coil is supplied with radio frequency via a low impedance coupling loop L_1 concentric with, but not attached to, the driving shaft on the same end of the drum as the coil is wound.

This coupling loop induces a high r.f. voltage in the coil L_2 which is resonated by the capacitance of the metal backed card and a swamping capacitor. This r.f. raises the potential of the card to 200V on peaks. The probes P_1 to P_{11} are approached close to, but not in contact with the card and as they are electrically screened along their whole length, can only pick up r.f. from their tips held close to the card surface.

When a punched hole passes immediately under the tip of any probe the amount of r.f. picked up by that probe drops sharply.

This reduction in energy reaching the probe is interpreted at the output of the amplifiers and rectifiers attached to the probe, as a positive pulse. This pulse is clipped and shaped by the following circuits, and delivered at low impedance to the output lines.

Circuits

The oscillator supplying the card with radio frequency is quite straightforward, and is shown in Fig. 2.

It consists of a 6V6 valve in a Hartley oscillator circuit inductively coupled by a three-turn loop L to an 80Ω coaxial cable. This coaxial cable feeds the drum tuned circuit via the fixed loop previously mentioned.

The probe amplifier V_2 has been described previously[1], but is included to show the coupling to the next clipping stage. Briefly V_2 combines the functions of rectifier and amplifier in the following way.

R_1, the negative bias control to the grid of V_2, is adjusted so that the radio frequency peaks picked up when a hole is stationary under the probe just does not cause any anode current flow.

Rotating the drum till the metalizing is under the probe raises the amplitude of the r.f. peaks reaching the grid of V_2 so that it now conducts on positive half-cycles. The r.f. component is filtered out in the anode circuit and the resultant waveform is applied via R_2 to the grid of V_3, the clipper stage.

The positive waveform reaching the grid of the clipper each time a hole is sensed, has an amplitude of approximately +30V.

As the grid is maintained at −20V and a 12AU7 has an I_aV_g cut-off of −10V under these conditions, the first part of the positive pulse is lost together with any noise picked up. The peak of the wave on the other hand is truncated as soon as the grid is driven positive to earth because of the grid stopper R_3.

In this way anode current can only change in V_{3a} during the centre swing of the positive pulse.

As shown (Fig. 3), a somewhat irregular shaped pulse has become a clear square wave in the anode of V_{3a}, with a peak-to-peak amplitude of approximately 130V.

The second half of V_3 is working as a cathode-follower and has the waveform from the anode of V_{3a} applied, via C_3 and R_5, to its grid.

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The voltage dividing network $R_4R_5R_6$ reduces the voltage swing available at the anode of V_{3a} to 70V at the grid of the cathode-follower (V_{3b}) and hence approximately 70V is available across the cathode load R_7, at low impedance.

What percentage of this waveform is required can be adjusted by R_7. Capacitor C_5 is included to compensate for any high frequency loss due to the series component of R_7 when the slider approaches the centre and lower settings.

An optional output has been provided to give the opposite phase and takes the form of R_8 in the anode of V_{3b}. It should be noted that the circuits of Fig. 2 are a.c. connected so that the drum can be slowed down to zero speed with no great change in pulse amplitude available at the output of each channel. This does not apply to the inverted phase available from the anode because a capacitor C_4 has to be included to block the high voltage.

Fig. 4 shows a card punched with holes which will give an output in accordance with the pattern specified.

Notice how by breaking through from one hole to another where required, pulses of long duration can be catered for. This should not be carried to an extreme however, because the strength of the card may be over weakened if too many bridges are removed.

In conclusion the following points are worth noting:

(1) The principle of using radio frequencies allows the use of a tuned circuit in the grid of V_2 and is an effective

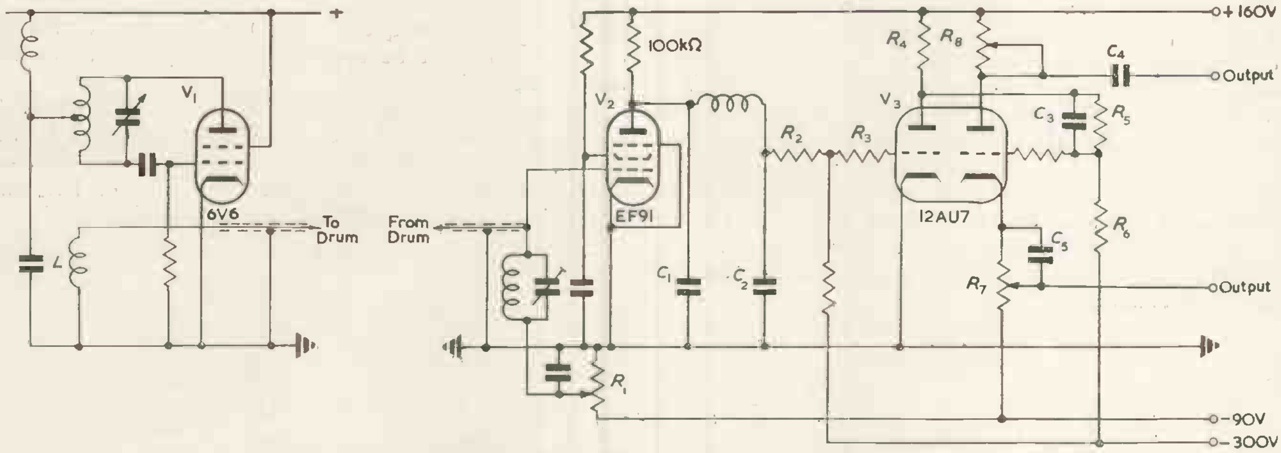


Fig. 2. Oscillator, probe amplifier, pulse clipping and shaping stages

short-circuit to interference, such as hum, electrostatic or induced. The same tuned circuit is adjustable so that the grid to earth capacitance of both the valve and the cables are "tuned out" and the change of r.f. amplitude at the probe is not by-passed, even with quite long connexions between drum and amplifier.

(2) Extremely long connexions (100ft) can be tolerated if the probe output is first transformed to low impedance by a tuned transformer before connexion to the cable.

(3) By means of parallel connexions to the r.f. generator and amplifier, the same lead which carries the r.f. to the device can be also used as the signal lead, but this way the moving member causes a reduction of r.f. reaching the amplifier when the capacitance increases instead of an increase, as in this case.

(4) If low frequency phenomena are being studied, a relay can be connected directly in the anode of V_2 (Fig. 2) and the clipper and cathode-follower dispensed with.

The specifications of the device just described are listed below.

P.R.F.: 8kc/s to zero.

Positive and negative output adjustable from 70V to zero volts.

Negative pulse, d.c. components retained.

Swing +35V −35V (R_7 set at maximum)

Total number of valves (excluding power supply) 23.

Time required to change from one pulse pattern to another 5min approx.

Number of channels: 11.

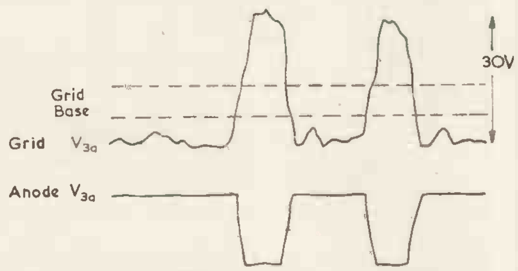
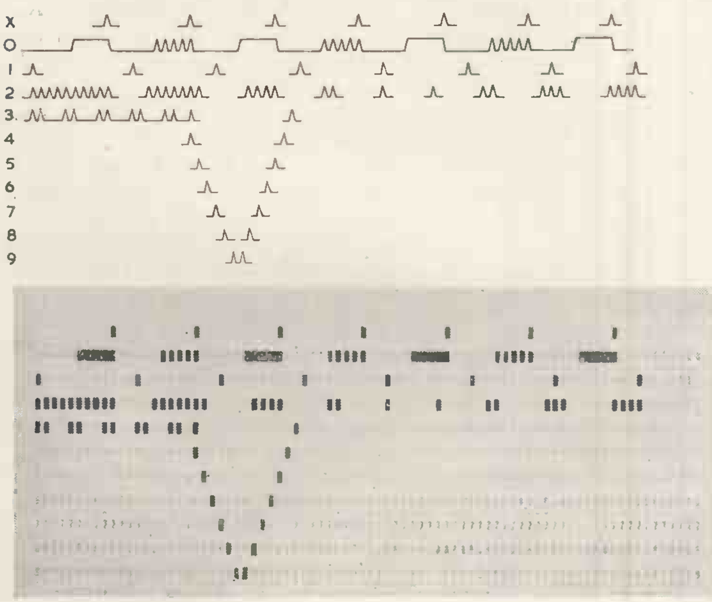


Fig. 3. Waveforms produced at grid and anode of V_{3a}

Fig. 4. Output pattern given by punched card (below)



REFERENCES

1. Woods-Hill, W. An Electrostatic Pulse Generator. *Electronic Engng.* 28, 122 (1956).
2. British Patent 703 038.

An A.C. Potentiometer for Measurement of Amplitude and Phase

By M. J. Somerville*, B.Sc.

A simple a.c. potentiometer circuit using a.c. coupled amplifiers is described. The design utilized enables the generation of in-phase and quadrature components which remain accurately in quadrature with one another, even though substantial phase shifts will occur in the a.c. couplings involved.

IT is often desirable in experimental work on servo-mechanisms, and in other fields, to be able to measure quickly and accurately the phase and amplitude of one a.c. voltage relative to another. The voltage to be measured may contain harmonics, in which case it is essential that the measurement is of the fundamental of the waveform. The potentiometer described herein enables measurement of fundamental amplitude and phase to be made over a range of frequency between 1·5c/s and 5kc/s, with an accuracy corresponding to ± 1 per cent of full scale on in-phase and quadrature balances.

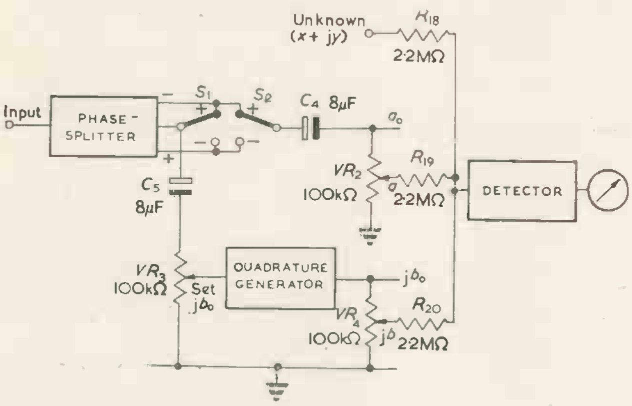


Fig. 1. The a.c. potentiometer

Basic Circuit (Fig. 1)

A sinusoidal input to the potentiometer is required. This is normally available from the input to the system under test. Phase splitting of this voltage is performed, and a means of setting the output (a_o) to a fixed level (10·0V r.m.s.) is provided. The switch S_2 enables selection of either positive or negative a_o.

Generation of the quadrature component is by means of a negative feedback integrator, with coarse adjustment of the time-constant for full coverage of the frequency range. The use of an integrator rather than a mutual inductance for generation of the guadrature term enables operation at low frequencies and has the advantage that any harmonics which may be present in the input are attenuated at the output. Fine adjustment for setting jb_o to the fixed level (10·0V r.m.s.) is by the potentiometer VR_3, and S_1 selects either positive or negative jb_o. The phase shifts due to C_4VR_2 and C_5VR_3 are arranged to have the same value, so that a_o and jb_o remain in quadrature at low frequencies. Due to extra phase shifts in the phase splitter and the couplings C_4VR_2 and C_5VR_3, $(a + jb)$ will have an undesirable phase shift relative to the input. This phase shift may be calibrated against frequency on a low frequency correction chart, which must be applied if the input is to be the reference signal.

The balance potentiometers, VR_2 and VR_4, provide signals a and jb, which are added to the unknown by the adding chain $R_{18}\,R_{19}\,R_{20}$ feeding a sensitive detector. Balance is obtained by alternate adjustment of VR_2 and VR_4 until the detected voltage is a minimum.

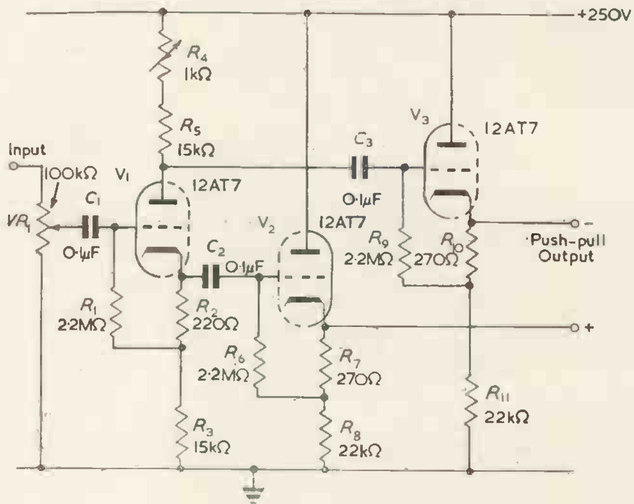


Fig. 2. Phase splitter

Phase Splitter (Fig. 2)

Equality of the push-pull output signal is adjusted by R_4, and will occur when $(R_2 + R_3) = (R_4 + R_5)$, provided the output circuits impose negligible loading. Signals at V_1 anode and cathode will be in anti-phase, and out of phase with the input by an amount dependent on R_1 and C_1. The input time-constant to the cathode-follower is $R_1C_1/(1-\beta)$, where β is the gain from grid to bias point. V_2 and V_3 are cathode-followers enabling loading of the push-pull signal. By making time-constants $C_3R_9/(1-\beta)$ and $C_2R_6/(1-\beta)$ equal, the outputs are still in anti-phase, though displaced from the input by a small angle. At low frequencies (1·5c/s) this displacement will be of the order of 3 degrees. Combined with the phase-shift due to C_4VR_2, a total shift of 11 degrees at 1·5c/s may be obtained, though a and jb remain accurately in quadrature.

Quadrature Generator (Fig. 3)

The component values for the amplifier (V_4) have been calculated to give a high gain (400 from grid to anode). V_5 cathode follows the high impedance output signal. The effect of phase-shift in the amplifier, due to C_8VR_4, is rendered negligible by taking the feedback capacitor C_9 to the output point, jb_o. If the amplifier gain is

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A, then the error in output phase is approximately $(1/A).(e_o/e_i)$ radians, where e_o and e_i are the output and input amplitudes respectively. For the circuit of Fig. 3, the maximum value of this error will be about $\frac{1}{3}$ of a degree. In addition, there will be a phase-shift if C_9 has any leakage. The phase-shift due to this effect is equal to the loss factor for the capacitor, and for those used this did not amount to any more than $\frac{1}{4}$ of a degree.

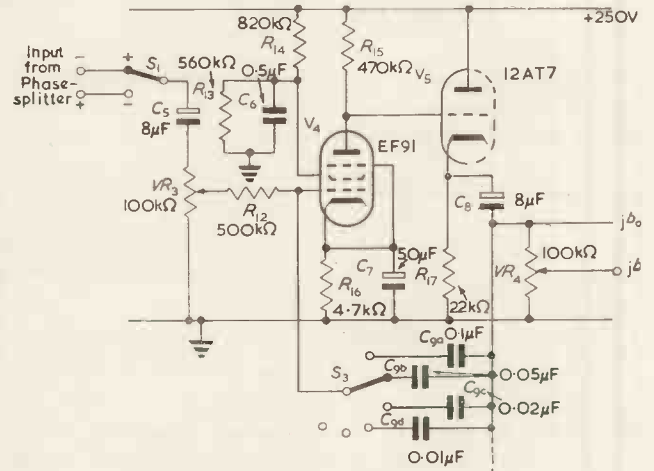


Fig. 3. Quadrature generator

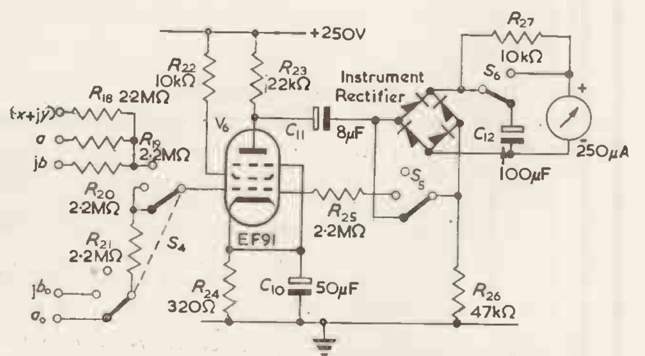


Fig. 4. Detector and voltmeter

Detector and Voltmeter (Fig. 4)

S_4 is a switch used for setting up the potentiometer. Positions 1 and 2 arrange the circuit as a negative-feedback voltmeter, when R_{26} is adjusted to give a full-scale deflexion for 12·5V r.m.s. input (with S_5 in position 2). With S_4 in position 3 the circuit is used as a detector, with sensitivity set by S_5, for which the three positions are OFF, VOLTMETER and DETECTOR. The inclusion of a smoothing circuit to eliminate pointer wobble on the detecting instrument needle is controlled by S_6, and need be used only at lower frequencies.

Balance Conditions

The procedure when using the instrument is as follows:

(1) Set S_5 on VOLTMETER and S_1 and S_2 as required.

(2) With S_4 in position 1, set a_o to 10·0V.

(3) With S_4 in position 2 and S_3 on the correct frequency range, set jb_o to 10·0V.

(4) With S_4 in position 3, adjust VR_2 and VR_4 alternately to give a minimum, using S_5 on DETECTOR for the final balance.

The operations (1) to (3) are only repeated when the supply voltage or frequency is altered.

It can be shown that, providing the balance components a and jb contain no harmonics, the balance obtained with an unknown containing harmonics is a balance with the fundamental of the unknown.

Suppose the unknown be:

$$e = \sum_{n=1}^{n=\infty} (x_n + jy_n) \sin n\omega t \ (n = 1 \text{ is the fundamental})$$

The signal from the potentiometer is $-(a + jb) \sin \omega t$.

The signal to the detector is therefore:

$$e_1 = [(x + jy) - (a + jb)] \sin \omega t + \sum_{n=2}^{n=\infty} (x_n + jy_n) \sin n\omega t.$$

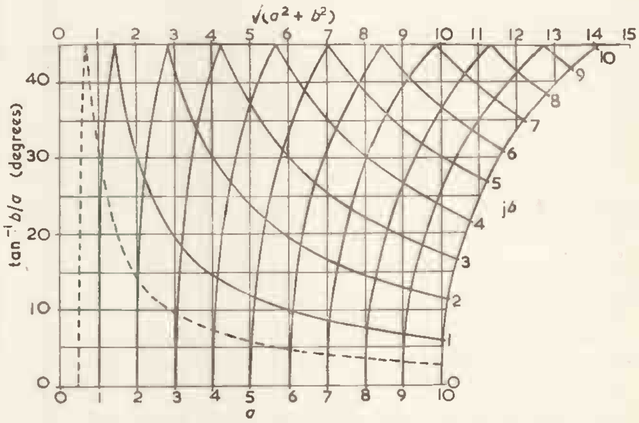


Fig. 5. Chart for obtaining the modulus and phase-angle

Let $E = [(x + jy) - (a + jb)]$

The r.m.s. value of the detected signal is then:

$$e_{r.m.s.} = \sqrt{[E^2 + \sum_{n=2}^{n=\infty} (x_n^2 + y_n^2)]}$$

$\therefore$ $e_{r.m.s.}$ is a minimum when E is a minimum, i.e. when $x + jy = a + jb$ (i.e. balance with the fundamental). It can be shown that the presence of harmonics of the unknown has negligible effect on the validity of the null balance on an average reading instrument as being balance with the fundamental, provided one of these harmonics predominates.

For extremely accurate measurements, however, a pre-filter should be used, or a sensitive r.m.s. reading instrument substituted for the bridge rectifier and moving-coil instrument of Fig. 4.

Conclusion

The accuracy of this instrument is dependent to only a second order upon the component values used. If greater accuracy of balance readings is required, the balance potentiometers VR_2 and VR_4 may be replaced by a conventional decade arrangement using a 12-way 2-gang switch for coarse balance and a continuous potentiometer for fine balance.

In practice the use of a chart such as that shown in Fig. 5 has been found very useful in obtaining a rapid conversion from $(a + jb)$ into $Ae^{j\phi}$ form. To obtain a good balance the unknown should not be too small, and the amount of harmonics present in it should not be too large, since the effect of these harmonics is to reduce the sensitivity of balance. Where the harmonic content is greater than 30 per cent it may be advisable to use a simple pre-filter for the unknown.

Reliability as a Design and Maintenance Problem

By R. Matthews*

Attention is drawn to the critical importance of reliability in complex electronic equipment. Valve performance is suggested as being largely responsible, and an estimate is made of this factor in equipment failure. A design philosophy is proposed which could improve operational (i.e. user) reliability, provided it was backed by the appropriate maintenance system. Valve by valve negative feedback and marginal checking techniques are then examined as related solutions to the problem.

THE introduction of electronics into the field of instrumentation throughout industry, its probable introduction into commerce, and its revolutionary contribution to the arts of war have forced the electronic designer to regard the reliablity of complex electronic equipment as his most outstanding problem.

This problem was once the province of the engineering and development teams who "engineered" the original circuits into a useable form. Today the inevitable comparisons are being made between electronic devices and the older mechanical, or even human, methods, with respect to reliability and robustness. And it is clear that future expansion in the use of electronic devices will be a function of how well they can attain the freedom from breakdown that characterizes the majority of machines already in use.

The Central Reliability Problem

Electronics is the most versatile of all the engineering sciences and a major factor contributing to this unique position is the relative ease with which the component parts can be acquired and assembled into a circuit. Nearly all the parts that go to make up a computer, radar set, gun control set or industrial control and sampling system are obtainable in any local radio shop.

This state of affairs brings with it a series of problems directly due to this widespread use of the component parts. These problems can be identified by comparing the tolerances in performance permitted by a design specification and the tolerances of the components which the designer is expected to use. With performances better than 1 per cent and resistors at 10 per cent, capacitors at 20 per cent, and valves at 20 to 30 per cent, the problem of designing repeatable circuits is not easy. Reliability of operation adds to the difficulty especially when the equipment is expected to perform for extended periods, under extreme climatic conditions[1].

The designer can, of course, overdesign on capacitor and resistor ratings and so attempt to increase the trouble-free life of an equipment. When the specification (size, weight, etc.) allows this to be done, a limit will be imposed by valve life. In what follows two suggestions are made for dealing with the problem posed by present day valve performance.

The first suggestion is based upon the proposition that breakdown is inevitable with present-day valves, and that a preventative maintenance technique can help avoid critical failure while in operation.

The second accepts the fact that valves contribute to unreliable performance, but seeks to show that reliability increases if more valves are used in the way prescribed.

That both these propositions gainsay popularly held intuitive notions should not detract from their more serious consideration.

Valve Life Estimates

From an analysis of past experience in using numbers of similar equipments it has been shown that the most regular failure is usually the valve. The type of statistic that is obtained is shown in Table 1.

TABLE 1

	PROTOTYPE EQUIPMENT	PRODUCTION EQUIPMENT
	(Per cent of	total failures)
Valve failures	60	90
Other Components	30	5
Plugs, cables and mechanical ..	10	5
Total	100	100

If the first item is further analysed then one can estimate that about 70 per cent or more of all these valve failures are characterized by the onset of a gradual decline in their performance.

These figures are intended to show the present-day dependency of reliability upon valve life and do not apply to the higher-standard valves which are beginning to be produced in numbers.

A Design Technique to Increase Reliability

In any circuit configuration one can generally identify the amplifier with a gain figure. If this gain figure falls because of valve deterioration then the output of the overall circuit falls in the same ratio.

The effect of the decrease in performance can be checked from interfering with the operation of the whole equipment by using negative feedback techniques in the original design. Now, providing the feedback is applied over each separate stage and not over the whole configuration, the overall performance can be made to remain sensibly constant until catastrophic failure occurs due to non-linearity or saturation of the channel. Thus it is possible to design circuits which will work with gradually failing valves until long after they would have otherwise become useless.

The argument against such a deliberate design policy is based upon the concomitant increase in the number of valves so needed to restore the external gain.

The following example shows the order of improvement that can be expected, and demonstrates that the freedom from gradual valve failure so achieved more than compensates for the fact that more valves are used.

Example

A first order assessment is made of the relationship between parameter dependency and the number of valves used when single stage feedback is applied.

Firstly, a single stage pentode amplifier is considered and an expression for the relationship between gain and the principal valve parameter is stated.

Secondly, the same overall gain is assumed to be achieved by two similar valves, each having a feedback resistor of the appropriate value in the cathode circuit, and the gain/parameter dependency in terms of the same incremental change is shown.

Lastly, the relationship is shown when both valves suffer

* *Decca Radar Ltd.*

the same change. The ratio of these relationships is discussed for some representative values.

Consider a pentode amplifier with an anode load R_L, a gain A, and a mutual conductance g_m.

Then:

$$dA = R_L \cdot dg_m \quad (1)$$

If the circuit includes a cathode resistor R_k, then in the usual notation:

$$A = \frac{R_L \cdot g_m}{1 + R_k g_m}$$

whence:

$$dA = \frac{R_L}{(1 + R_k g_m)^2} \cdot dg_m \quad (2)$$

Now the special condition in the equivalent two-stage amplifier is that the gain should just equal that of the single valve and is given by:

$$A_2^2 = A$$

where A is the overall gain and A_2 is the stage gain.

Thus since $dA_2 = \dfrac{R_L}{(1 + R_k g_m)^2} \cdot dg_m$

$$dA = A_2 dA_2 \quad (3)$$

where one A_2 is assumed constant.

Now the value of A_2 must be determined by finding the value of R_k which gives a stage gain equal to the root of the total gain.

Thus:

$$A_2 = \frac{R_L g_m}{1 + R_k g_m} \text{ and } A = A_2^2 = R_L g_m$$

$$R_L g_m = \left[\frac{R_L g_m}{1 + R_k g_m}\right]^2 \text{ or } R_k = \frac{\sqrt{(R_L g_m)} - 1}{g_m}$$

Substituting in the expression for dA_2:

$$dA_2 = (1/g_m) \cdot dg_m \quad (4)$$

and putting this in equation (3):

$$dA = A_2 \cdot (1/g_m)\, dg_m \text{ or using } A_2 = \sqrt{A} = \sqrt{(R_L g_m)}$$

$$dA = \sqrt{(R_L/g_m)}\, dg_m \quad (5)$$

which is comparable with equation (1)

The improvement thus effected insofar as relative independence from changes in g_m are concerned is the ratio of the coefficients in expressions (1) and (5) for dg_m.

That is:

$$\sqrt{(g_m R_L)} \text{ or } \sqrt{A} \quad (6)$$

It may be argued, however, that a variation may occur equally well in both valves, so the previous reasoning can be developed to show the improvement in parameter dependence even when both valves of the amplifier are subjected to the same change as in the single valve they replaced.

Since $A = A_2^2$ we have, for both $A_2 \rightarrow A_2 - dA_2$

$$A = A_2^2 + (dA_2)^2 - 2A_2 dA_2$$

Thus:

$$dA = 2\sqrt{A} \cdot dA_2$$

or:

$$dA = 2\sqrt{(R_L/g_m)}\, dg_m$$

and the ratio representing the inputs is now:

$$\frac{R_L}{2\sqrt{(R_L/g_m)}} = \tfrac{1}{2}\sqrt{A} \quad (7)$$

To recapitulate:

For one stage the gain varies directly as the g_m,

$$dA = R_L\, dg_m$$

For two stages of equivalent gain, the gain now varies

(a) for one valve changing its g_m,

$$dA = \sqrt{(R_L/G_m)}\, dg_m$$

(b) for both valves changing their g_m,

$$dA = 2\sqrt{(R_L/G_m)}\, dg_m.$$

The extension of this reasoning to n valves is, of course, limited only by the considerations relating to the relative frequency of faults due to slowly changing g_m and all other faults such as short-circuits, breakages, etc., and all component failure in respect of the added components.

A Preventive Maintenance Technique

The basis of a preventive maintenance technique is the acceptance of the fact that, due to causes outlined at the introduction to this note, failure is inevitable in complex equipments and that it is therefore better to anticipate trouble rather than wait for it.

Ideally, the maintenance crew would check all valves (since they have already been shown to be potentially the most likely cause of failure) on a valve tester and reject those which had become unacceptable through usage. This is not very practicable considering the numbers involved, and could mean small changes in balance, etc. To test the valves *in situ* is a better scheme, but not always possible for some circuit configurations. In any case one does not require information about the good valves, only the poor ones; and since it is to be hoped that they are relatively rare this implies a proportional waste of time.

What is really required is the rapid indication of only the poor valves so that they may be replaced prior to using the equipment.

To do this one can vary the applied potentials on the valves in their circuits and detect valves approaching the limit of usefulness by the failure of the equipment. Two general types of equipment are considered in which different methods of testing are used.

The first type of equipment comprises the groups, including computers, which are characterized by large-scale repetitive circuits. It is possible to have here a second h.t. line which can be switched in so that the valve parameters are substantially changed. Failure to operate in this condition is taken to indicate a gradually failing characteristic and the valve is replaced. It is assumed that upon restoring the circuit to its normal condition the equipment will function very reliably, insofar as emission failures of the gradual type are concerned, for the period immediately after the check.

The second general type of equipment is the non-repetition circuit type of equipment such as is exemplified by large radar sets, where the provision of separate h.t. lines and switching will considerably complicate the system. It is here suggested to exploit the heater dependency of valve parameters, and by reducing the heater current to a predetermined value, establish which valves are prone to failure

AN EXAMPLE USING HEATER CURRENT REDUCTION

So far as can be gleaned from the position of a "consumer", emission failure of the gradual type is caused by poisoning of the vacuum by gasses released at the high electrode temperatures in the structure, by gradual "stripping" of the cathode emitting surface, by slow leakage at the pin entry, or by mechanical movement of the electrodes due to excessive thermal stresses. In all these cases the primary effect is upon the current capacity of the valve and manifests itself as a commensurate change in the g_m. The change in grid base, both signal and compressor grids, is much smaller and may, in some circumstances actually alter in the opposite sense.

Some defects such as the development of an interface resistance is due to core-matrix interaction with residual gases and vapours and does not give rise to simple g_m current relationships.

These variations can be very closely simulated by heater current reduction as shown in Table 2.

From this table it can be seen that a controllable reduction in some important parameters can be achieved by reducing the applied heater potentials.

(c) Alteration of heater current makes a smaller difference to the cut-off characteristics than to the other parameters of the valve.

(d) It lends itself to an automatic switching-on sequence in which heater burn-out is avoided and the test position automatically "consulted" before going into the final operating state.

TABLE 2

VALVE Heater Potential Percentage Reduction	TYPE	CURRENT			g_m			CUT-OFF		
		6·3V (0%)	5V (16%)	4V (33%)	6·3V (0%)	5V (16%)	4V (33%)	6·3V (0%)	5V (16%)	4V (33%)
1. High Slope Pentode	CV138 (6AM6)	0%	32%	61%	0%	39½%	81%	0%	12%	21%
2. High Slope Triode	CV455 (12AT7)	0%	23%	48½%	0%	23%	43½%	0%	9%	20%
3. Low Slope Triode	CV491 (12AU7)	0%	23½%	56%	0%	27%	57½%	0%	7%	17%
4. Power Pentode	CV2127 (6CH6)	0%	21%	47½%	0%	13%	47%	0%	10½%	19%
5. Heavy Current Pentode ..	CV345 (12E1)	0%	5½%	9%	0%	3%	18½%	0%	3¼%	3¼%

Information is also needed as to the principal parameter of the valves used in the different configurations in an equipment.

Now, in order to estimate the effectiveness of a test consisting of a reduction in heater voltage, one must assess first how many circuits are being correctly tested as opposed to those in which the principal parameter is not being so drastically altered.

These figures must now be weighted in terms of their relative frequencies in the equipment, and finally it must be remembered that in the case of cut-off, the tendency is to make conditions more favourable instead of less. That is, where circuits are "cut-off" sensitive, it is the quality of "shortness" that is important and the heater current reduction actually does shorten the grid base.

The advantages to be gained from such a checking procedure are:

(a) The system is simple and applicable to equipment already designed.

(b) The system affects all valves whether fed by stabilized h.t. supplies or not.

Conclusions

Systems of marginal checking have, of course, been in use in the laboratory for some time and there are some examples of its application in computers both here and in the States[2,3].

The result of a marginal checking procedure being adopted for the SEAC equipment gives a 60-fold increase in the trouble-free operating period. That is, 60 times as many failures occurred in the prescribed marginal check period as did when operating.

Other results from a radar display system in daily use give a figure of between seven and ten.

If such a preventative procedure could be adopted on an equipment in which component over-design and valve by valve feedback had been incorporated, then the possibility of failure in action could be reduced to a level acceptable to all classes of user.

REFERENCES

1. Hunt, G. L. A Survey of Quality and Reliability Standards in Electronic Valves for Service Equipment. *J. Brit. Instn. Radio Engrs.* *11*, 519 (1951).
2. Forrester, J. W. Digital Computers Present and Future Trends. Joint AIEE-IRE Computer Conference. (Feb. 1952).
3. Slutz, R. J. Engineering Experience with SEAC. Joint AIEE-IRE Computer Conference. (Feb. 1952).

Electronic Photography*

"Electrofax"[1] is a process developed at the RCA Laboratories which shows promise of becoming a useful alternative to processes using conventional photo-sensitive materials, for photography, office copying and duplication, photo-engraving and other photo-resist applications.

Electrostatic systems of photography are dry and rapid, the best known system, "Xerography" being described as[2] employing:

> "A re-usable plate consisting of a thin layer of a photo-conductive material on an electronically conducting base. The plate is sensitized by electrostatic charging immediately before use. After exposure, the image is developed by dusting the plate with a micronized powder. Prints are made by transferring and fixing the powder image to paper or other materials."

"Electrofax" Process

In the "Electrofax" process, the photo-sensitive element is a mixture of powdered zinc oxide in a resin binder, which can be coated on practically any metallic or non-metallic base. Zinc oxide is relatively inexpensive and white in colour, an ordinary paper base with the applied mixture possessing the photographic sensitivity of silver halide printing papers, with the distinct advantages of low cost, long shelf life and insensitivity to light until electronically charged.

This charge is applied in the dark or under suitable safe-light conditions by means of one or more fine wires which, at a d.c. potential of 6kV negative to the metal plate beneath the paper are swept across the coated surface, transferring ions by corona discharge. The paper is now sensitized and ready for exposure and the usual precautions must be taken to shield it from stray light. Exposure by any conventional means reduces the charge in the exposed areas in proportion to the amount of incident light, leaving a latent electrostatic image on the coating corresponding to the dark areas of the original.

Development is achieved by applying a positively charged pigmented resin powder, called a toner, which adheres by electrostatic attraction to the negatively charged areas on

* *This article is based on part of the Peter Le Neve Foster Lecture given by C. G. Mayer, of the Radio Corporation of America, before the Royal Society of Arts, on 16 May, 1956.*

the coated surface. An ingenious method of applying the toner is by means of a magnetic brush consisting of a permanent magnet carrying at one end a mass of iron powder loaded with the fusible toner powder, positively charged due to the surface phenomena known as "tribo-electric effect". When the brush is swept over the coated paper the toner particles are attracted to the charged areas of the electrostatic image and give a reproduction of the original image, the quantity of charge at any point determining the resultant density. The iron carrier powder remains on the magnet.

Fixing of the image is effected by baking for a few seconds at a temperature sufficient to fuse the toner to the paper surface. This may be done in the light, since the powder image stays in place for some time. After baking, the image is as rugged and permanent as an impression in printer's ink. Multi-colour prints are possible by using coloured toners. "Electrofax" paper is stable, keeping indefinitely without deterioration[3].

Photographic Characteristics

The sensitivity and contrast of Electrofax coatings processed in the way described are about equivalent to a relatively high contrast silver halide contact paper possessing an ASA speed of 16. Resolution already obtained with a 200-mesh developing powder is of the order of 300 to 500 lines in such resolution with a relatively coarse developing powder being attributed to selection of the finer particles by the highly resolved charged image. (See Table 1).

TABLE 1
Approximate exposure requirements for Electrofax printing

LIGHT SOURCE	EXPOSURE CONDITION	TYPICAL EXPOSURE TIMES FOR A SATISFACTORY PRINT	
		UNSENSITIZED ZnO PAPER	ROSE BENGAL SENSITIZED PAPER
TUNGSTEN 100W bulb 2ft. from exposure plane	Contact through positive transparency	½sec	1/50 to 1/25sec
BLACK LIGHT Two 4W fluorescent tubes 2ft. from exposure plane	Contact as above	1/10sec	1/10sec
ELECTRONIC FLASH UNIT 10ft. from exposed plane	Contact as above	Single flash 10^{-6} to 10^{5-}sec	Single flash 10^{-6} to 10^{-5}sec
BRIGHT DAY-LIGHT	Adapted camera lens at f 4.5	½ to 5 sec	1/10 to ½sec

Dye Sensitization

Zinc oxide is sensitive primarily in the ultra-violet and far blue regions of the spectrum. This range may be extended into the visible by the addition of suitable common dyes, which raise the sensitivity due to increased energy absorption over a wider band, enabling special purpose papers to be designed for almost any desired spectral response. A typical Electrofax orthochromatic sensitizer is Rose Bengal, which gives an increase in effective speed of about 30 times for tungsten light.

Projection printing is easily possible, a prototype 35mm microfilm printer having produced 8½in by 11in prints, using a 1sec exposure with an $f/8$ lens and 300W projection lamp, while in bright sunlight, using an adapted camera, prints have been produced on Rose Bengal sensitized paper with exposures of 1/5 to ½sec at $f/4{\cdot}5$.

Special Applications

Since a variety of resins and waxes may serve as photoconductor binders and image materials, many chemical processing techniques become possible in a wide range of special applications, three of which are outlined below.

PRINTED CIRCUIT BOARDS

These may be quickly prepared by the above photoresist method, a thin Electrofax coating being applied to the surface of the copper-clad laminate. The desired circuit pattern is exposed on to the charged surface and the image developed with an acid resistant toner and fused in place in the usual way. After removal of the coating in the non-image areas by differential solvent action, the copper thereby exposed is dissolved away, the required copper circuit being ready for use on removal of the remaining fused coating by a second solvent.

Experimental or production circuit boards may be produced in a matter of minutes with an enlargement or reduction as desired. No intermediate photographic transparencies are needed, as the resist image may be laid down by contact printing from any circuit drawn on tracing paper.

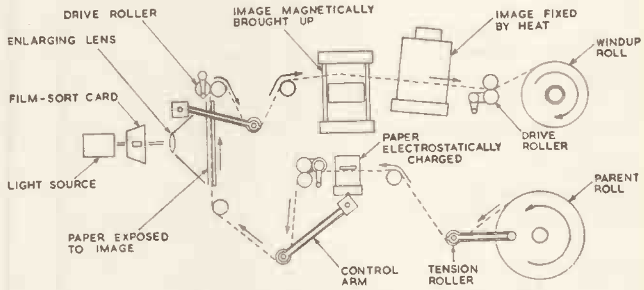


Fig. 1. Automatic enlarger printer

PHOTOGRAPHIC TRANSPARENCIES

If an Electrofax coating is applied to a transparent base, lantern slides, films or other transparencies may be produced. One method is to give a sheet of glass with a transparent conductive coating an Electrofax coating, print in the usual way and dissolve away the zinc oxide with a suitable solvent, leaving the fused image on a transparent background, ready for projection.

AUTOMATIC ENLARGER PRINTER

A recent commercial application of the Electrofax process is a fast automatic dry direct enlarger-printer capable of reproducing fifteen 17in by 22in engineering drawings per minute from microfilm originals. The machine accepts 35mm positive film either in roll form or in punched cards containing selected picture frames. A schematic diagram is shown in Fig. 1. This system releases the space needed for full size originals and permits centralization of vital drawings for maximum security and preservation. A rack above the lens system can be loaded with 500 cards, which are automatically dropped in front of the lens at the rate of one every 4 seconds, while the roll film in lengths of up to 100ft is being fed in at the same picture rate.

Other Applications

The use of Electrofax is being explored for such diverse purposes as making of templates for shipbuilding and aeroplane fabrication, mural decorations, nameplates and package labelling.

REFERENCES

1. YOUNG, C. J., GREIG, H. C. 'Electrofax'—Direct Electrophotographic Printing on Paper. *RCA Rev.* 15, 469 (1954).
2. SCHAFFERT, R. M., OUGHTON, C. D. A New Principle of Photography and Graphic Reproduction. *J. Opt. Soc. Amer.* 38 (1948).
3. SUGARMAN, M. L. 'Electrofax'—A New Tool for the Graphic Arts. Proc. of Seventh Annual Meeting of the Technical Association of the Graphic Arts (May 1955).

BOOK REVIEWS

Ionized Gases

By A. Von Engel, 281 pp. 144 figs. Demy 8vo. Oxford University Press. 1955. Price 42s.

BASED on lectures given to undergraduates at Oxford, this book is of an introductory nature, but is none the less complete in itself and draws information from many recent original papers. Emphasis is on the physical processes involved rather than on mathematical theory or practical techniques, and the arrangement of material in the book makes it both readable and convenient for reference.

The author has restricted the mathematics to what can be expressed concisely, and the proofs are carefully explained. There are however a few instances of slips with symbols which may prove troublesome to the beginner. Familiarity with simple kinetic theory of gases is assumed but the introduction to Chapter III enumerates some properties of the Maxwellian distribution and defines some kinetic theory terms.

In general, methods of measurement are described only briefly at the end of each chapter, and the reader is referred to the literature for details. High frequency discharges and plasma oscillations are not treated, nor are applications of the subject to atmospheric studies, discharge tubes or high voltoge engineering.

The text is well illustrated by graphs and tables of measured values selected from the literature. In addition there are numerous small diagrams which are of considerable aid to the memory (for example those showing particles colliding). A valuable feature is that a short numerical example is given to illustrate each formula derived.

Following a historical introduction, a chapter is devoted to conduction in the absence of multiplication processes.

There follows a long chapter on the "Production of Charged Particles", describing excitation, ionization and emission phenomena. The results of complex quantum mechanical derivations are presented in relatively simple language and the limitations of classical collision theory made clear.

In Chapter IV, entitled "Mobility and Charge Transfer", Langevin's theory of ionic mobility, taking induced dipoles into account, is explained, and a simplified theory of electron drift velocities derived. Charge transfer is considered qualitatively, and the results of wave-mechanical calculations are compared with measured values.

The chapter on 'Diffusion and Mutual Repulsion" includes discussions of ambipolar diffusion and the effect of a magnetic field on electron diffusion, while Chapter VI is devoted to recombination effects, treated both analytically and in terms of a coefficient.

Chapter VII, "Ionization in an Electric Field" is concerned mainly with electron multiplication and the breakdown process. Townsend's equation is introduced, and the primary and secondary coefficients, whose mechanisms were analysed in Chapter III, and here considered phenomenologically.

Time-lags and the high pressure spark mechanism are the subjects of the final section.

Dr. von Engel has himself contributed greatly to the understanding of the glow and arc discharges, and Chapters VIII and IX are devoted to these topics, which are developed in terms of the ideas evolved in the earlier chapters. Theories are given of the mechanisms operating in each of the regions into which these forms of discharge can be divided, and deductions from these theories are compared with experimental results.

A set of short appendices contains accounts of special theoretical points raised in the text, tables of integrals, etc. There are over 200 references and a good index.

"*Ionized Gases*" is clearly a valuable text book for the student whose degree syllabus includes a course on the subject, while for the physicist or engineer it provides a concise and illuminating account of the processes governing conduction in gases. It is no small achievement in the field of gaseous electronics to have conveyed so much information with so little strain on the reader's credulity, patience, and mathematical ability.

C. H. CARMICHAEL

Basic Processes of Gaseous Electronics

By L. B. Loeb. 1 012 pp. 320 figs. University of California Press. Cambridge University Press, London. 1955. Price 100s.

SINCE the war there has been an increased interest in the study of electrical discharges in gases, or "gaseous electronics" as it is now called, evident from the number of papers which have been published. Considerable advance has been made both in the theoretical understanding of the subject, and in the application of new experimental techniques. In spite of this, very few text books have been published during the last fifteeen years, and therefore this book by Professor Loeb, who is recognized as one of the foremost authorities on gaseous electronics, is particularly welcome.

The book deals with the kinetic mechanism and basic processes involved in the interaction between electrons, atoms, and molecules which lead to the behaviour of gases in electric fields. It is aimed at critically presenting the fundamental theory and experimental data necessary for the understanding of electrical breakdown in gases, and is primarily a reference book for physicists, engineers, or chemists working in this field.

The subjects considered in detail are summed up in the chapter headings: 1, Ionic mobilities; 2, Diffusion of Carriers in Gases; 3, The velocities of Electrons in Gases; 4, The Distribution of Energy of Electrons in a Field in a Gas; 5, The Formation of Negative Ions; 6, The Recombination of Ions; 7, Electrical Conduction in Gases Below Ionization by Collision; 8, Ionization by Collision of Electrons in a Gas—Townsend's First Coefficient; and 9, The Second Townsend Coefficient.

Readers who are familiar with Professor Loeb's earlier book *Fundamental Processes of Electrical Discharge in Gases* published in 1939 will recognize that the above general chapter headings follow the same pattern as the first nine chapters of his previous work. However, the author stresses that this new book in no way represents a revised edition of "Fundamental Processes", although the reviewer was a little disappointed that most of the experiments and theory of a more classical nature dealt with in detail in his first book, have been included with equal thoroughness. As an example, thirty pages are devoted to experimental methods of measuring mobilities which can be found in "Fundamental Processes" and Chapters II and VII remain unaltered since there has been little to add.

This does not imply that the book is not up to date or comprehensive in the subject it covers. On the contrary, the author has included and commented on all relevant papers, particularly presenting latest techniques and critically evaluating the more recent results. In presenting the more advanced material Professor Loeb has enlisted the help of co-authors; Professor Brown wrote on microwave studies of recombination, Dr. Wannier wrote on the theory of ion mobilities, and Dr. Molnar and Dr. Hornbeck contributed to "Metastable Action" and "Dynamic Time Studies of Townsend's Second Coefficient" respectively. Each co-author is an expert in his own field and the book therefore gives the opinion of foremost American scientists working on gaseous electronics.

Each topic is introduced in terms of a simplified kinetic theory, and is fairly readily understood. However, the more stringent treatment that follows is more difficult, and in the reviewer's opinion, requires a greater knowledge of atomic structure and kinetic theory than might be suggested by the author's preface.

To the physicist or engineer concerned with the application of the electrical behaviour of gases to glow discharge devices, the last two chapters dealing with ionization are probably of most interest. In these chapters, to which a third of the 1 000 page book is devoted, Professor Loeb gives a clear picture of present theoretical knowledge and helps to resolve the conflicting results that have appeared in the literature, especially those relating to secondary ionization processes. By drawing upon unpublished as well as published work, reliable and useful data on Townsend coefficients are given for a number of standard gases of known purity.

In these chapters the author gives his first hand experience of experimental procedure gained from forty years of research in gaseous electronics. Thus on page 664 one can read of his reasons for preferring mercury vapour diffusion pumps for gas discharge work to the reputedly more efficient oil pumping systems; and later of electrode dimensions for obtaining uniform fields.

Unlike the author's earlier book, there are no chapters on glow or arc discharges, which would be of more use to the electronic engineer engaged in development of, or circuit work with, gas discharge devices. However, the author states that it is hoped his book will lay a foundation for a second book by himself, and possibly books by his colleagues, that will deal more directly with the various breakdown processes. Such a book would have a wider appeal, and is awaited with interest.

The general presentation of the book is good, although some of the diagrams were not as clear as in "*Fundamental Processes*": for example in figure 8.23, the shapes of the curves around 25V, referred to in the text, were quite impossible to see in the reviewer's copy. The bibliography is most excellent and there is no doubt that for workers in gaseous electronics the book is well worth its rather high cost.

G. F. WESTON

Noise

By A. van der Ziel. 450 pp. 97 figs. Demy 8vo. Chapman and Hall Ltd. 1955. Price 60s.

THIS book is concerned with the effects in electronic circuits of natural spontaneous fluctuations. The theoretical study of these phenomena began before the invention of the thermionic valve in 1904 but the subject was largely of scientific interest until the nineteen-thirties when the sensitivity of radio sets became high enough for the basic fluctuations in circuit resistance and valve current to show themselves as noise in the audio output. Since then the necessity for high gain amplifiers for radar, television, and other purposes has given rise to considerable study of noise problems and there has been for some time the need for a comprehensive work on the subject. The present book fulfils this need and it is fortunate that Professor van de Ziel, who has contributed so much to the study, should have undertaken its preparation.

Starting with thermal noise, not only in resistors but also in microphones, the book goes via a discussion of "noise figure" to methods of noise measurements. This is followed by chapters on valve noise at low and high frequencies, including the effects of secondary emission and gas current, and on low noise, mixer, and feedback circuits. To deal adequately with mixer circuits a chapter is interposed on noise in semiconductor devices such as crystals and transistors.

This general discussion fills the first two thirds of the book, and, in it, the author has taken care to restrict the mathematics to the type of equation and method of working that will be familiar to the competent engineer. The more advanced mathematical treatments of statistical methods and Fourier analysis of fluctuating quantities, and their application to noise in detector circuits, are fully dealt with in separate chapters.

This is followed by an excellent condensed treatment of the laws of valve electronics, leading to a discussion of space charge waves and noise in beam devices. The terminating chapter deals with the effects of fluctuation phenomena in forming the limit to the sensitivity of various physical measuring instruments.

Two valuable features are noticeable in the book; the first is the very clear cross referencing between the various parts which deal with the same subject; the other is the inclusion of the related device theory where it is necessary for an understanding of the noise material. Thus, before the treatment of electronic beam devices there is the chapter devoted to the laws of valve electronics, while noise in semiconductor devices is preceded by a discussion on solid state theory.

The order of presentation of subjects in an interwoven field must always present difficulty and this is so in the present case. The attempt to treat thermal noise fully before going on to the theory of shot noise has led to the necessity for consideration of the application of shot noise formulae before they have been formally developed. In fact, although Chapter III is entitled "Applications of Nyquist's and Schottky's Theorems", no indication is given as to which of the formulae presented represents Schottky's theorem. To find out, it is necessary to read on for some sixty pages.

This is, however, a minor criticism in a book that is well produced and almost devoid of inaccuracies. A treatise on noise can never be regarded as simple reading, but, as a reference or a text book, the present volume should be on the bookshelf of all those concerned in any way with the valve or circuit aspects of noise.

W. H. ALDOUS

Electric Motors and Controls

278 pp. 14 figs. Demy 8vo. The British Electrical Development Association, 1956. Price 8s. 6d.

This book has been written with two aims in view. First, to demonstrate how the intensive use of electric motors and controls has helped to raise the productivity of both manufacturing plants and individual machines and, secondly, to give the non-technical reader an insight into the principles of electric motors and control gear, the variety of types that are available, and how to choose the right motor for a particular drive.

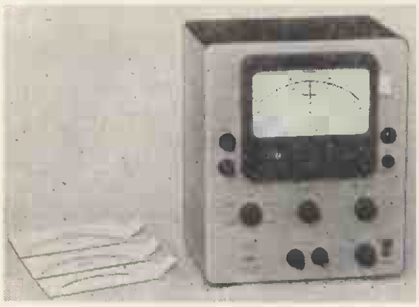
ELECTRONIC EQUIPMENT

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

Brüel and Kjær Deviation Bridge

(Illustrated below)

DESIGNED for use primarily in the production inspection of resistors, capacitors and inductors, Type 1504 uses a large illuminated moving-coil meter to indicate the percentage deviation from standard in terms of impedance and phase-angle. Positive and negative impedance deviations of 1, 5 and 20 per cent and phase-angle deviations of 10^{-2}, 5×10^{-2} and 20×10^{-2} radians can be read from separate scales inserted at will at the side of the indicator. The very high indicating speed of the bridge enables 4 000 units per hour to be tested within the ranges 10Ω to 10MΩ, 50pF to 10μF and 2mH to 100H. A demodulation circuit provides a zero stability such that readjustment of the standard controls is generally necessary only once per day.



Level Recorder

TYPE 2304 is a high speed graphic rectilinear recording instrument for signal level variations in the frequency range 20c/s to 200kc/s. The recorder is useful for the recording of any phenomenon which can be converted into voltage variations, such as reverberation decay, noise levels, voltage levels and the response of microphones, loudspeakers, filters, etc. The principle of operation is that of a balanced type of stylus positioning made possible by a patented potentiometer-controlled moving-coil drive system. Recordings are made by either a sapphire stylus on wax coated paper or an ink pen, the effective recording width being 50mm. with a choice of 10 paper speeds between ·003 and 100mm/sec. The input potentiometer provides full scale recording ranges of 10, 25, 50 or 75dB for a logarithmic characteristic.

Frequency Analyser

(Illustrated above right)

THIS instrument, Type 2105, operates on the degenerative principle and has a constant percentage bandwidth. It is applicable to all kinds of acoustic and vibration-technique research, being used for voltage measurement and analysis, as an indicator in accurate bridge measurements and for measuring sound pressure and levels.

There are 8 frequency ranges between 47c/s and 12kc/s read directly from the scales. Variable selectivity is provided in 5 steps, the maximum position corresponding to 40dB, which facilitates distortion measurements of the order of 1 per cent. The frequency accuracy is better than 1 per cent and as a linear amplifier in the range 30c/s to 15kc/s and as an analyser, the amplification is constant within ±1dB. As a sound level meter the frequency response characteristics are the internationally proposed characteristics at 30 to 60 phons and 60 to 130 phons.

Brüel and Kjær, Denmark.

Brush Surindicator

MODEL BL110 is a practical workshop instrument designed for the accurate measurement of surface finish roughness and simple enough to be used by personnel after only a few minutes instruction. The compactness and light weight of the equipment enable measurements to be made in the production machine if necessary. The surface roughness range is 1 to 1 000μ in arithmetic average deviation from the mean surface. A variable cut-off switch permits the separation of waviness and roughness characteristics of surfaces by filtering out wavelengths exceeding ·003, 0·01 or 0·03in, which is of importance in applications such as bearing surfaces and highly stressed parts. Calibration is carried out with the aid of precision reference specimens.

Brush Electronics Company, Ohio.

These two pages are devoted this month to a selection of the overseas equipment shown at the 2nd International Instrument Show which was organized by B & K Laboratories Ltd, 57, Union Street, London, S.E.1, who are the distributors for all the equipment described.

Disa Electronic Tachometer

(Illustrated above)

THIS is a 2-channel tachometer requiring no mechanical coupling between the rotating shaft and the pick-up head, a small permanent magnet being attached to the shaft whose speed is to be measured and the pick-up head placed close to the magnet. The instrument is sensitive to frequency and virtually insensitive to voltage and rotor-stator misalignment. The speed of rotation is read directly from a dial calibrated in revolutions per minute. The device is of particular importance in turbo operation but is easily adapted for any rotating shafts within the specification.

Disa Electronik, Denmark.

Krohn-Hite Push Button Oscillator

(Illustrated above)

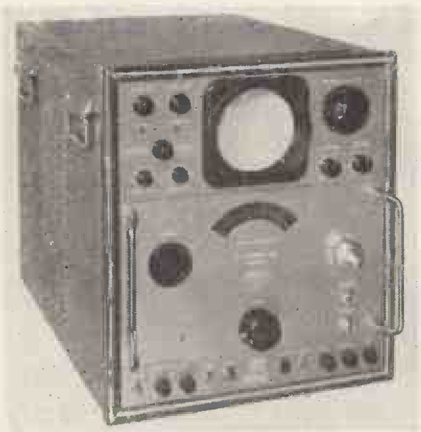
MODEL 440A provides both sine and square waves over the range 0·01c/s to 100kc/s continuously variable, controlled by a fine position decade multiplier switch, 3 banks of decade push buttons and a vernier control to cover the range between adjacent buttons in the third bank. Distortion and hum are less than 0·1 per cent at any output level, the maximum being 30V peak-to-peak. Frequency calibration accuracy is ± 1 per cent from 1c/s to 10 kc/s and ± 3 per cent over the whole range, with an hourly drift of less than 0·02 per cent. The amplitude variation is less than ± 0·25dB and switching transients are eliminated by special circuits. Special uses for this oscillator include bridge and distortion measurements, tuned filter alignment and rapid spot frequency checks.

**Krohn-Hite Instrument Co.,
Cambridge,
Mass.**

**Nuclear Instruments
Ampli-Count Scaler**
(Illustrated above)

MODEL 182 is a laboratory instrument for manual or automatic measurement of radio activity. A high voltage supply changing less than ·002 per cent for 1 per cent change in supply voltage and an input sensitivity range of 1mV to 1V permit its use with all commercially available Geiger, scintillation or proportional detectors. Additional features of this equipment are:— a 2μsec resolving time; 8 binary stages providing a maximum scaling factor of 256; a built-in mechanical register and timer; power supplies for an external detector; a calibrated, continuously variable linear amplifier; and a mains frequency test signal to permit checking of the scaling stages at any time.

**Nuclear Instrument and Chemical
Corporation,
Chicago.**



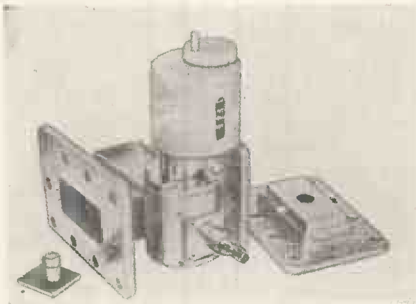
**Polarad
Spectrum Analyser**
(Illustrated above)

MODEL TSA is a broad band r.f. analysis instrument providing a 5in cathode-ray display of the complete frequency distribution of energy of an r.f. signal, 5 interchangeable tuning units covering the range of 10 to 44 000Mc/s. The high resolution of the instrument enables stable and accurate measurements to be made of signals at increments of less than 40kc/s. Typical applications: the observation and measurement of a.m. and f.m. signal sidebands; the accurate frequency measurement of radio and radar signals; the checking of magnetron spectra and the tracking of radar r.f. components; the determination of r.f. pulse characteristics by spectrum analysis.

**Polarad Electronics Corporation,
Long Island City, N.Y.**

**Sivers Lab
Frequency Meters**
(Illustrated below)

TYPE SL5212 can be used as a quick, direct reading laboratory instrument where intermediate accuracy is required,



or built into broad-band radar or instruments. The frequency meter has a patent cavity and plunger which cause the frequency, but not the wavelength, to vary linearly with the plunger length over a frequency range of 2 500 to 4 000Mc/s. The frequency is easily read on a counter giving the value directly in Mc/s. One turn of the knob corresponds to the same frequency change anywhere in the band, so that a flexible shaft or servo system can be used for connexion to a remote counter or chart recorder.

The cavity of the meter has two coupling windows to which the following coupling elements can be connected: End-connected or passing waveguide; end-connected or passing coaxial; crystal diode mount. For one coupling circuit, accuracy is better than ± 0·1 per cent and with the supplied correction curves, better than ± 0·05 per cent resetting accuracy being about ± 0·01 per cent. The loaded and unloaded Q are approximately 2 500 and 3 000 respectively.

**Sivers Lab,
Sweden.**

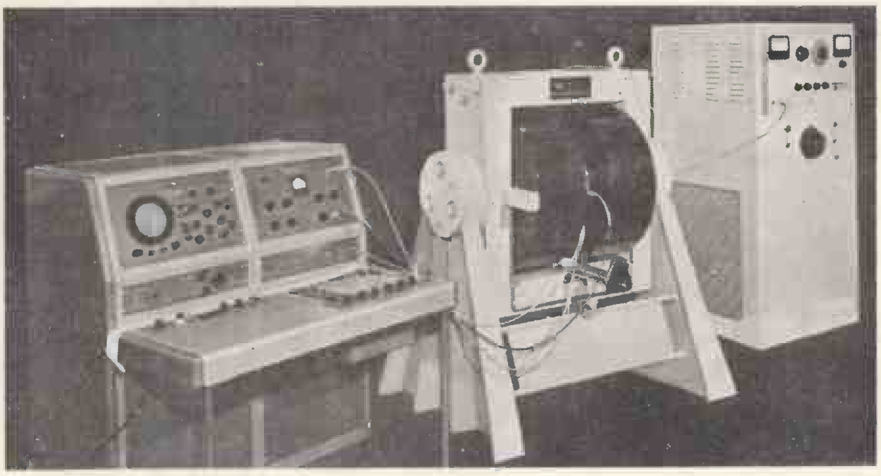
**Varian
N-M-R Spectroscope**
(Illustrated below)

NUCLEAR Magnetic Resonance spectroscopy is based on the fact that the various isotopes of the elements can be separately identified according to their differing nuclear gyro-magnetic constants.

There is an angular momentum associated with each isotope, also magnetic moment, the ratio of these two being the gyro-magnetic ratio for that particular nucleus. This ratio differs for each isotope, so permitting separation and identification according to the scheme of N-M-R spectroscopy. In addition, high resolution spectroscopy represents a new method for determining molecular structure and indentifying and measuring components in a mixture. The equipment consists of (1) a magnet whose gap field can be varied from zero to 10 000 gauss (2) a low power transmitter supplying r.f. energy to (3) a small transmitter coil in the magnet gap (4) a small receiver coil located within the transmitter coil and which surrounds (5) a Pyrex test tube containing the sample being investigated (6) a radio receiver tuned to the transmitter frequency and amplifying signals induced in the received coil by the gyroscopic precession of the nuclear "magnets" and (7) an indicator such as a graphic recorder, oscillograph or voltmeter which registers the presence of the signals. The spectrum is displayed dynamically in a recurring fashion by superimposing a small periodic sweep field on the strong magnetic field.

High Resolution N-M-R- techniques are finding applications in organic chemistry, biology and medicine, for example. qualitative and quantitative analyses of functional groups in chemical and petroleum research, study of hydrogen bonding, molecular rearrangements, fluorine chemistry and the silicones.

**Varian Associates,
California.**



LETTERS TO THE EDITOR

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

Convex Resistance Functions

Dear Sir,—Shannon and Hagelbarger[1] have recently enunciated a principle of impotence in two-pole resistance network design, namely that the resistance $R(R_1, R_2 \ldots, R_n)$ of a two-pole network N of non-negative linear resistances R_1, $R_2, \ldots, R_n$ is a concave downward function $R_1, R_2, \ldots, R_n$. This means in effect that for any two sets of non-negative values $R_1, R_2, \ldots, R_n$ and $R_1', R_2' \ldots, R_n'$ we have:

$$R(R_1 + R_1', R_2 + R_2', \ldots, R_n + R_n') \geqslant R(R_1, R_2, \ldots, R_n) + R(R_1', R_2', \ldots, R_n') \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

It may be of interest, therefore, to demonstrate the existence of a family of physically realizable, reciprocal, *resonant* passive linear two-pole networks N whose resistances violate the concavity condition (1).

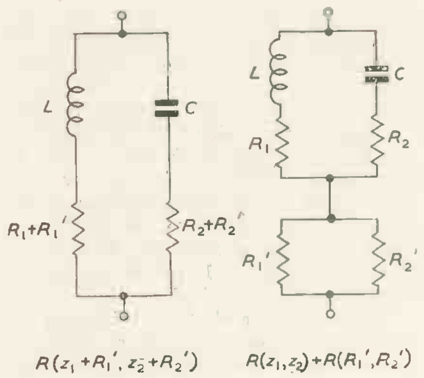


Fig. 1.

Consider the shunt combination of two conjugate complex linear impedances $z_1 \equiv R_1 + jX_1$, $z_2 \equiv R_2 + jX_2$, where $R_1 = R_2 = R > 0$, $X_1 = -X_2 = X$, say. This network is resonant and its resistance $R(z_1, z_2) = R/2 + X^2/2R$. Suppose now that the branch resistances R_1, R_2 are increased to $R_1 + R_1'$, $R_2 + R_2'$, respectively, where $R_1' = R_2' = R'$ say, $R' > 0$. Then the modified network is resonant and its resistance is given by:

$$R(z_1 + R_1', z_2 + R_2') = \tfrac{1}{2}(R + R') + X^2/2(R + R').$$

And since $X^2/2(R' + R') < X^2/2R$ for all $R, R' > 0$ it follows that:

$R(z_1+R_1', z_2+R_2') < \frac{1}{2}(R+R') + X^2/2R$, i.e.:

$$< R(z_1, z_2) + R(R_1', R_2') \quad \ldots\ldots (2)$$

This proves the existence of physically realizable, reciprocal, resonant passive linear two-poles violating the concavity condition (1).

For example, if (with the usual connotation) $z_1 = j\omega L + R$, $z_2 = -j/\omega C + R$, where $\omega L = 1/\omega C$, $R > 0$, then the corresponding network satisfies equation (2). [Cf. Fig. 1.]

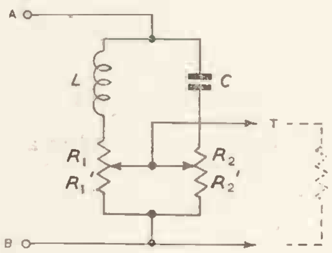


Fig. 2.

Shannon and Hagelbarger were concerned originally to design a rheostat with linear resistance components satisfying a convexity condition exemplified by equation (2). They do not, however, state whether the input to their rheostat was intended to be a.c. or d.c. If the former is the case, then the network of the above example, assuming convenient component values can be found to give resonance at the input frequency, may provide an acceptable solution to their problem. If, on the other hand, d.c. was intended, the above network would need to be preceded by a linear d.c.-a.c convertor and to be followed by a linear a.c.-d.c. convertor, and it is impossible to state (without further information) whether the resulting network could be designed on an economic basis or not.

The circuit connexions in the a.c. case could take the form illustrated in Fig. 2. R_1 includes the effective series resistance of the coil L. Evidently, if the input resistance of any circuit to the right of the tapping point T is large compared to $\frac{1}{2}R_{max}'$, the input resistance found between terminals A and B is independent of the position of T.

REFERENCE

1. Shannon, C. E. Hagelbarger, D. W. Concavity of Resistance Functions. *J.Appl. Phys. 27*, 42 (1956).

Yours faithfully,

H. M. Melvin,
Sanderson Engineering Laboratories,
University of Edinburgh.

A D.C. Coupled Circuit Using Voltage Stabilizing Valve

Dear Sir,—Further to Mr. Court's reply to my letter in the June issue, I do not wish to modify my earlier remarks but I should be happy to take a second opportunity to explain and drive home my point.

The introduction of neons as in Fig. 2 of the article (December 1955) overcompensates and reverses the klystron drift by associating the reflector with positive supply line drifts instead of negative ones. Clearly then a reduced number of neons can be chosen to effect a balance. The fact that in this case the supply lines do not drift certainly conceals this interesting property because it amounts to wearing two belts with our a.f.c. braces. I say this because neither compensation nor stabilized lines are necessary if the a.f.c. really does its job, although either (but surely not both!) are expedient if error handling capacity is limited. Stabilized lines are an unwelcome complication, and it may well be if they are not required for the local oscillator they are not required at all.

Mr. Court prefers to use a neon chain rather than a floating supply? Well—*chacun a son gout*. Incidentally, we would not stabilize this supply either, the t.r. cells do not require it.

Yours faithfully,

R. J. D. Reeves,
E. K. Cole Ltd.,
Malmsbury,
Wilts.

Diffraction Gratings

Dear Sir,—In the March issue of Electronic Engineering reference was made in an article "Digital Methods in Control Systems" to the development at the N.P.L. of Merton diffraction gratings and a method of measuring small displacements.

Our attention has now been drawn to a note under the heading "Errata" in your April issue, which we feel does not put our position in this work in proper perspective.

The facts are, that N.P.L. starting from the basic ideas of Sir Thomas Merton, has developed (a) gratings suitable for linear measurement and (b) the optical principles of using these gratings to produce moire fringes.

The gratings and the optical principles have been made freely available to Ferranti Ltd., Edinburgh, and to various potential users. Ferranti Ltd., have applied these Merton-N.P.L. gratings to their own system of machine tool control and have developed the technique of using them in their system. They have exclusive rights to the use of gratings for machine tool control to the extent to which they are entitled by their patents on their own system.

Yours faithfully,

R. Edmonds,
National Physical Laboratory,
Middlesex.

An A.C. Voltage Stabilizer

Dear Sir,—We have been reading through our article published in the June issue of Electronic Engineering and find that there is a small mistake in the text and that the constructional details of the transformers and the reactor were omitted.

The mistake occurs in paragraph 3 on p.262 where the words "reduction in input power" should read "reduction in maximum output power".

The details of the reactor and transformer are given below:—

SATURABLE CORE REACTOR

Core:— $2\frac{1}{4}$ in. Stack of Sankey 35A Stalloy Stampings

D.C. Windings:— 15 000 turns of 36 s.w.g. copper wire.

A.C. Windings:— 800 turns each of 22 s.w.g. copper wire.

AUTO-TRANSFORMER DESIGN (TRANSFORMER T_1)

Core:— $1\frac{1}{4}$ in stack of Sankey 36A Stalloy Stampings.

Reactor-End Winding:— 810 turns of s.w.g. copper wire.

Output-End Winding:— 390 turns of s.w.g. copper wire.

INTERNAL SUPPLIES TRANSFORMER DESIGN (TRANSFORMER T_2)

Core:— $1\frac{1}{4}$ in stack of Sankey 36A Stalloy Stampings.

Primary Winding:— 24 s.w.g. copper wire. If tapped at 10-0-200-220-240V the turns required are 0-38-783-860-935.

H.T. Windings:— 36 s.w.g. copper wire. For 350-0-350OV, 2 740 turns, centre-tapped. For 315-0-315V, 2 400 turns, centre-tapped.

L.T. Windings:— For 6·3V, 1A. (3 required) 25 turns of 20 s.w.g. copper wire. For 5V, 2A, 20 turns of 18 s.w.g. copper wire.

Yours faithfully,

F. A. BENSON.
M. S. SEAMAN.
Department of Electrical Engineering,
University of Sheffield.

Tolerance Limits in Matching

DEAR SIR,—By an unfortunate choice of implied definition for mismatch, Dr. Alexander arrives at the conclusion[1] that for a given mismatch the load should be chosen higher rather than lower the generator impedance.

This conclusion has been reached by assuming that the percentage error in the load resistance is the criterion of matching. For small errors (less than 20 per cent. say) such an assumption is harmless, but by considering errors of 100 per cent and greater, it becomes obvious that this is an unsatisfactory parameter to use. Thus for errors of 100 per cent the load resistor is zero or twice the correct value, while for greater percentage errors the load becomes a generator (negative resistance) for the lower value.

Some other criterion of matching must be sought and equation (9) in the article may be modified to suggest one; thus:

$$\frac{\text{Power}}{\text{Matched Power}} = f = \frac{4q}{(q+1)^2} \qquad (9)$$

where $q = \dfrac{\text{Load resistance}}{\text{Generator resistance}}$

Equation (9) may be rewritten:

$$f = \frac{4}{q + (1/q) + 2} \qquad (9a)$$

which shows that a given degree of mismatching occurs for ratios $q = n$ and $q = (1/n)$ (e.g. $f = 0{\cdot}75$ for $q = 3$ or $\frac{1}{3}$ as in Table 1 of the article).

Thus mismatch should be considered relative to the ratio of actual load to ideal load rather than percentage departure of load from ideal.

That this is also the case for most applications of resistors has been recognized in the choice of the standard values of resistors, which have been chosen with equal ratio steps:

(e.g., in the 20 per cent tolerance range the decade has a range of 10/1 and 6 steps so that if the ratio of each step is r:

$$\log r = 1/6 \log 10$$
$$r = 1{\cdot}4677$$

and the values 10, 15, 22, 33, 68, 100 have been obtained by rounding off to the nearest two figures).

Functions dependent on ratios are best plotted to a logarithmic base and when this is done the unsymmetrical curve of Fig. 1 in the article becomes symmetrical about $q = 1$ as in Fig. A. Fig. 2 could also be made more symmetrical by this method of plotting.

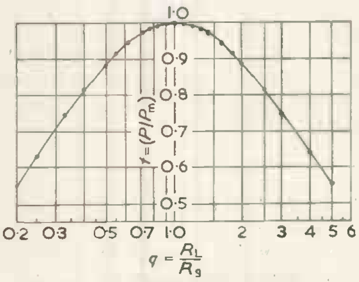


Fig. A. Logarithmic plot of Fig. 1.

To summarize, a percentage based graph distorts most curves since there are few parameters where -100 per cent is just as small a departure from the desired value as $+100$ per cent. Conclusions based on such curves can be misleading as in this instance. Dr. Alexander's conclusions still hold providing one reads "percentage departure from matched load" for his "a given mismatch" but since matching (and most other usages of resistors) depends on the ratio, actual value/desired value, the percentage base distorts the picture (a resistor of 110Ω is 10 per cent above 100Ω but the 100Ω resistor is only 9·09 per cent down from 110Ω).

Yours faithfully,

W. H. P. LESLIE.
Mechanical Research Laboratory, D.S.I.R.

REFERENCE

1. ALEXANDER, W. Tolerance Limits in Matching. *Electronic Eng.*, *28*, 162, (.956).

The author replies:

DEAR SIR,—Mr. Leslie's remarks form an interesting extension to the discussion of this topic.

That the conclusions are misleading is not true, however, when considered within the context of the article.

I should like to take this opportunity of pointing out a typographical error in Table 1, page 163. All the values of power transfer are negative and not "±" as shown.

Yours faithfully,

W. ALEXANDER,
The University,
Nottingham.

Comment on the Saturated Thermionic Diode

DEAR SIR.—Some recent discussions of the saturated thermionic diode[1,2] have related the anode current I_a to the filament voltage V_f by a solution:

$$I_a = K V_f^n \qquad (1)$$

where K is a constant. It is found, however, that n is not really constant, but decreases slowly with increasing V_f.

It has been proposed[3] that the rational relation between I_a and V_f is:

$$I_a = aV_f^{4/5}\exp(-bV_f^{-2/5}) \qquad (2)$$

a and b being constants. It is true, as pointed out in the answer to Reference 3, that practical diodes will rarely, if ever, follow equation (2) exactly. Nevertheless, that relation, having at least some rational basis, might be applicable over a wider range than would any purely empirical equation.

Fig. 1 shows $\ln(I_a/V_f^{4/5})$ versus $V_f^{-2/5}$ for some results taken from Reference 2. The plots are seen to fit fairly well to straight lines, as equation (2) would predict, and the slope gives a value of b around 40.

If equation (1) were correct, one would have $d(\ln I_a)d(\ln V_f) = n$, i.e., a plot of $\ln I_a$ versus $\ln V_f$ would give a straight

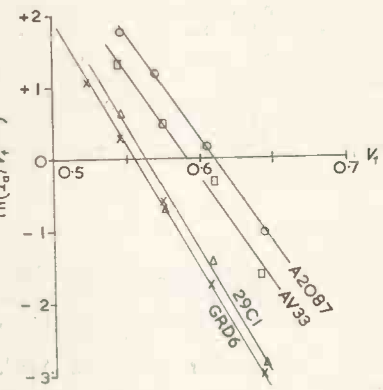


Fig. 1.

line, of slope n. For equation (2) one finds:

$$\frac{d(\ln I_a)}{d(\ln V_f)} = V_f\frac{d(\ln I_a)}{dV_f} = 4/5 + \frac{2b}{5V_f^{2/5}} \qquad (3)$$

Thus, if equation (2) is to be approximated over some region by equation (1), n should be equal to $4/5 + 2b/5V_f^{2/5}$. The appropriate value of n will then decrease with increasing V_f, as already mentioned. Also, $b = 40$ would make n around 8 or 9 for applicable values of V_f, in agreement with the references.

REFERENCES

1. BENSON, F. A., SEAMAN, M.S. Saturated Diodes. *Electronic Engng. 27*, 360 (1955).
2. FRASER, H. J., ANTHONY, V.C. Saturated Diodes. *Electronic Engng. 28*, 41 (1956).
3. ARMSTRONG, H. L. Saturated-Diode Operation. *Electronic Engng. 25*, 216 (1953).

yours faithfully,

H. L. ARMSTRONG,
Radio and Electrical
Engineering Division,
National Research Council of Canada
Ottawa.

Short News Items

The Institution of Electronics (Northern Division) will hold their Eleventh Annual Electronics Exhibition at the College of Technology, Manchester, from 12 to 18 July. In addition to the exhibition lecture and film show programmes have been arranged. Admission tickets (free) and further details may be obtained from the Honorary Organizing Secretary, Mr. H. Birtwistle, 78 Shaw Road, Rochdale, Lancashire.

The Third International Congress on High Speed Photography will be held in London from 10-15 September. More than forty papers dealing with the techniques and applications of high speed photography will be presented. Many of the techniques and applications featured in the Congress will be demonstrated at an International Exhibition which will be held in the Congress building. A number of films dealing with high speed photography will also be shown. Further information is obtainable from the Congress Secretariat, Third International Congress on High-Speed Photography, Department of Scientific and Industrial Research, Charles House, 5-11 Regent Street, London, S.W.1.

The 14th annual British Radio Component Show will be held in the Great Hall, Grosvenor House, Park Lane, London, W.1, from 9 to 11 April 1957 with a possible extension to 12 April. There will be a full day's preview on Monday, 8 April, for specially invited visitors instead of the half-day preview this year. Admission is by ticket only, obtainable from the Radio and Electronic Component Manufacturers' Federation, 21 Tothill Street, London, S.W.1.

H.R.H. The Duke of Edinburgh's Study Conference on the Human Problems of Industrial Communities within the Commonwealth and Empire will take place at Oxford from 9-27 July. The Duke of Edinburgh will give his Presidential Address at the opening ceremony in the Sheldonian Theatre on the afternoon of 9 July. The Study Conference will bring together some 300 people in manager/employer and trade union/operative roles of industry from a wide cross-section of races, industries and occupations. There will be 90 members from the United Kingdom and 138 from other countries of the Commonwealth and 52 from Colonial territories, in addition to 20 Group Chairmen. 29 countries and territories will be presented.

The Institution of Post Office Electrical Engineers will celebrate the 50th anniversary of its foundation during the 1956-57 session. The first Council meeting of the Institution was held on 6 June 1906, and the first local centre meeting on 8 October 1906. The October 1956 issue of The Post Office Electrical Engineers' Journal will be a special issue celebrating the Institution's anniversary and will be devoted to articles reviewing the development and growth of the British Post Office telecommunications services and of the mechanization of postal services. Those interested are advised to place an order before 1 August with the publishers of The Post Office Electrical Engineers' Journal, Messrs. Birch and Whittington, 49 Upper High Street, Epsom, Surrey. This special issue will be sold at the usual price of 2s. 6d.

The BBC announces that a site has been chosen for the projected medium power television station in Cumberland at Sandale, 1 200ft above sea level and some 14 miles south-west of Carlisle. The new station is expected to be ready for service before the end of 1957; it will bring the BBC's television service to the greater part of Cumberland and Westmorland as well as to the coastal areas of south-west Scotland.

The Physical Society Exhibition in 1957 will be held in both the Old and New Halls of the Royal Horticultural Society, Westminster, London, S.W.1. The dates of the Exhibition are the 25th to 28th March inclusive.

The Radio Industry Council will hold a Scottish Radio and Television Exhibition at the Kelvin Hall, Glasgow, in mid-May 1957. It is expected that all the leading manufacturers of radio and television receivers will exhibit.

Marconi's Wireless Telegraph Co Ltd are to supply the U.S.S.R. with a large quantity of television equipment. Among the main items in the order are included two 3-Camera Television Outside Broadcast Vehicles, two mobile petrol-electric Trailer Power Units and a comprehensive supply of spares.

Britain's Master Metre—a copy of the international standard of linear measurement in the metric system—is being repolished and reruled by the Societé Genevoise D'Instruments Physique in Switzerland. The bar weighs eight pounds and is made of 90 per cent platinum and 10 per cent iridium. It is valued at £3 000.

The Telegraph Condenser Co. Ltd have recently celebrated their Golden Jubilee, having been established in 1906.

The Birmingham Branch Office of Midland Silicones Ltd has moved to larger and more central premises at Union Chambers, 63 Temple Row, Birmingham 2.

Sir Harold Bishop, Director of Technical Services of the BBC, will in future be known as Director of Engineering.

RCA Great Britain Ltd is the new name for RCA Photophone Ltd. The change of name does not indicate any alteration in the policy or constitution of the company.

Mr. W. M. York has been elected to the board of directors of E. K. Cole Ltd.

Mr. Clive Barwell, F.I.A.M.A., General Publicity Manager, Mullard Ltd, has been re-elected President of the Incorporated Advertising Managers' Association for a further year.

The headquarters address of the Export Credits Guarantee Department is now 59 Gresham Street, London, E.C.2 (telephone number Monarch 6699).

Farnell Instruments Ltd are moving to new premises on 1 July at Hereford House, North Court, Vicar Lane, Leeds 2. The telephone number remains unchanged, Leeds 32958.

Telcon-Magnetic Cores Ltd, of Chapelhall, Airdrie, Lanarkshire, announce the appointment of Mr. Brian D. Jenkins, B.Sc., A.M.I.E.E., as Technical Representative for the company, covering London and the South of England.

Erratum. In the description of the Multiple Pulse Generator, manufactured by A. E. Cawkell, on page 269 of the June issue, the rise-time for the output pulse should be $0{\cdot}1\mu$sec and not 1μsec as stated.