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Commentary

"RADIO RESEARCH 1955" is the title of the report published by the Radio Research Board covering their activities over the past year.

It is a title which may conjure up visions of extensive and well equipped laboratories devoted to research and development of what we label "radio".

If "Propagation 1955" were substituted it would convey a more accurate meaning to the work of that very important branch of the Department of Scientific and Industrial Research.

The Radio Research Board can, of course, be fairly described as the main centre of propagation research in this country and it has co-operated closely with the two international organizations mainly concerned with propagation namely the International Scientific Radio Union (U.R.S.I.) and the International Radio Consultative Committee (C.C.I.R.). The former body concerns itself with the physics of radio while the latter deals with the technical problems involved in the most efficient use of the radio frequency spectrum throughout the world for communications, including sound and television broadcasting and navigation including radar.

The value therefore of the Radio Research Board's work—concerned as it is with what happens in the intervening space between the transmitting and receiving aerials—can hardly be overestimated.

The Radio Research Board is interested in the whole of the radio frequency spectrum. At one end of the scale the Board has started to investigate the possibilities of very low frequencies, about 15kc/s, for the operation of long distance navigational aids. As yet not much is known about propagation at these very low frequencies and it is anticipated that the investigations will have to be on a large scale with the co-operation of many organizations at home and abroad. One of the immediate difficulties the Board has encountered is the dearth of high power transmitters in this very low frequency band.

At the other end of the scale, because of the very high frequencies involved, the Board have found it necessary to introduce the little known symbol the gigacycle (Gc/s) in their studies of the propagation of surface waves on wires and plane sheets.

Another important side of the Board's work is on the behaviour of the ionosphere and the forecasting of radio transmission conditions and for this work observation laboratories are maintained not only at the Slough Headquarters but as far afield as Singapore and Port Stanley, Falkland Islands.

In the past few years considerable attention has been given to the possibilities of long distance transmission in the very high frequency range by means of scattering in the ionosphere and it is to be expected that the Board would take more than a passing interest in this new development in radio communications.

A considerable amount of practical information has been amassed and is now being analysed as a result of experiments carried out in co-operation with the Post Office and Admiralty on transmissions from the Shetland Islands.

Taking the analysis in conjunction with experience gained in the United States, it would appear that many of the earlier predictions were over optimistic, for it now seems that the frequency band from about 25 to 60Mc/s offers a means of regular communication for telegraphy only for distances up to about 1 300 miles. Telephony on a 24 hours/day basis now seems impracticable except by uneconomical "brute force" transmitters and the earlier forecast of a transatlantic television link using the scatter technique has been shown to be fundamentally unsound.

Much of this work may appear dull and lacking in glamour but without the Radio Research Board and similar organizations, world communications by the radio link would quickly come to a standstill. Let no one doubt the value of their efforts.

TRIDAC

A Research Flight Simulator

By J. J. Gait*, M.A., B.Sc., A.M.I.E.E., A.Inst.P., and J. C. Nutter†, M.A.

(Part 1)

The capabilities required of analogue computing machines are indicated, leading to a description of the large tri-dimensional machine at the Royal Aircraft Establishment, Farnborough. The basic computing elements and unique features of this equipment are discussed, followed by a detailed analysis and comparison of the three approaches to the axis transformation problem, with reference to the requirements of TRIDAC.

The three monitoring systems used are described and a speculation is made on the capabilities of future computers.

IN the past twenty years advances both in electronic techniques and components have been marked and electronic devices have been developed to meet a vast variety of needs. Important among these are computing machines of various types. Many of the present-day problems facing scientists and engineers are such that progress is materially speeded up, and in many cases is rendered practical, by the use of large complicated computing machines.

Of the various types of electronic computing aids which have been evolved that which has received the most publicity is the digital machine. Such machines, using coded streams of pulses to represent numbers and instructions, can, by speeding up the processes of numerical mathematics, solve complicated numerical problems with a combined accuracy and speed which outstrips the human calculator. Certain problems, however, call for a different type of machine—the analogue computer. Such machines are based on the use of physical quantities, such as length, angle or voltage, which in the computer vary continuously in proportion to the variables of the problem. The simplest and most familiar example of this type of computer is the slide rule in which lengths proportional to the logarithms of numbers are manipulated. Using modern electronic techniques, analogue computers capable of solving a variety of complicated problems with great speed are now possible. The two types of machines, the digital and the analogue, have fields of application in which each is the most appropriate type. The analogue machine is, at present, however, the only type capable of use as a " simulator " of certain problems. Thus in the fields of servomechanisms, autocontrols for aircraft and guided missiles, etc., it is important that there be computing machines capable of reproducing the dynamics of the system under study. It is also desirable that these machines operate on a real or 1 : 1 time scale, so that the machine reproduces events on the same time scale as would the real system, as this enables actual parts of the real or proposed system to be incorporated into the framework of the computing machine. This latter facility is required for two reasons:

(1) The exact mathematical description of the part in question may not be known, as in the case, for example, if the part is a complicated non-linear device or perhaps a human operator such as the pilot of an aircraft.

(2) When incorporated in the computing machine the part in question may be tested under conditions more realistic than in any other form of test short of a full-scale trial. Machines of this general type are referred to as simulators, although all such machines do not necessarily work on a real time scale. In effect, simulators provide working models of the real system under investigation. In many applications, for example those associated with the dynamics of high speed flight, the digital machines cannot at present operate fast enough to give real time scale solutions and the problem of incorporating parts of the real system into such machines is most difficult. On the other hand, electronic analogue machines are ideally suited to such applications, being capable of the necessary speed of operation and such that the inclusion of parts of the real system within the framework of the computer is relatively straightforward.

The basic electronic analogue computing techniques suitable for simulators are now well known[1,2] and a variety of such machines have been constructed. These machines range in size from a small rack of equipment up to TRIDAC (TRI-Dimensional Analogue Computer)[3], the largest and most ambitious machine of this type as yet built in this country. TRIDAC, which is located at the Royal Aircraft Establishment at Farnborough, was jointly designed by the Royal Aircraft Establishment and Elliott Brothers (London) Limited, for the purpose of simulating the flight, control and guidance of guided missiles and aircraft in a three-dimensional volume of space. The machine, which occupies some 6 000ft² of floor area, is partly electronic (8 000 valves) and partly mechanical (e.g. nine high performance hydraulic servomechanisms). The building housing the machine and its power generating plant (600kVA) is shown in Fig. 1.

Before considering some of the more interesting features of TRIDAC, a brief statement of the need for simulators in general and TRIDAC in particular is appropriate.

Theoretical and numerical work, directed towards studying the dynamics of systems to determine performance, stability, etc., is subject to a number of serious shortcomings. The classical methods of solving integro-differential equations are in general suited only to problems of analysis; that is the performance, stability, etc., can be determined for a given system, but should these be unacceptable then the method of analysis usually gives no guide as to how the system should be altered to yield a desired performance and stability. In a large number of present-day design tasks the problem is to construct systems which will meet specific performance and stability criteria; that is the problems are essentially ones of synthesis rather than analysis.

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† *Elliott Bros. (London) Ltd.*

The operational calculus, based on the Laplace transform, provides a powerful method for the synthesis, from simple elements, of complicated systems which will meet desired criteria. Operational methods are now in general use for work on servomechanisms etc., but they are applicable in a direct form only to linear systems—in particular to systems capable of representation by linear integro-differential equations with constant coefficients. If the system is not of this type then, in general, analysis of selected systems and trial and error methods rather than direct synthesis must be employed in the search for the system arrangement which meets the desired criteria. Simulators provide the means whereby we may by-pass some of the limitations of purely theoretical or numerical work. They can be used to extend operational synthesis into the non-linear domain and to provide large numbers of solutions if a trial and error exploration is necessary or if statistical properties are required. As discussed earlier a real time scale simulator can include parts of the real system which are difficult or impossible to represent mathematically. Finally, but by no means least, simulators, by virtue of their "working model" properties, give to the user a physical appreciation of what might otherwise be only a complicated mathematical formulation of his problem. In the field of guided weapons, the mathematics of the dynamics of flight of a missile, with six degrees of freedom and under automatic guidance and control, as it is directed towards a manoeuvring target are most involved. The coefficients in the equations are continuously changing because of their dependency on such functions as altitude, speed, mass, etc. The equations are non-linear because the kinematics of the relative motion of target and missile involve trigonometric functions. Discontinuities also occur due to physical limits on the motion of moving parts, to backlash, striction and so on. The cost and uncertainties of full-scale flight trials are so high that experimental trial and error methods cannot be considered and an elaborate simulator such as TRIDAC is the most practical method for detailed design studies prior to flight trials.

A General Description of TRIDAC

As a starting point it is to be noted that facilities are provided for the simulation of:

(1) A point-mass target which may be manoeuvred at will in a volume of space.

(2) The aerodynamic forces X, Y, Z along, and the aerodynamic moments L, M, N about, the missile principal axes. Each of these forces and moments can be considered to be of the form:

$$Y = (\partial Y/\partial u) \,.\, u + (\partial Y/\partial v) \,.\, v + (\partial Y/\partial w) \,.\, w + (\partial Y/p) \,.\, p + \text{etc.} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

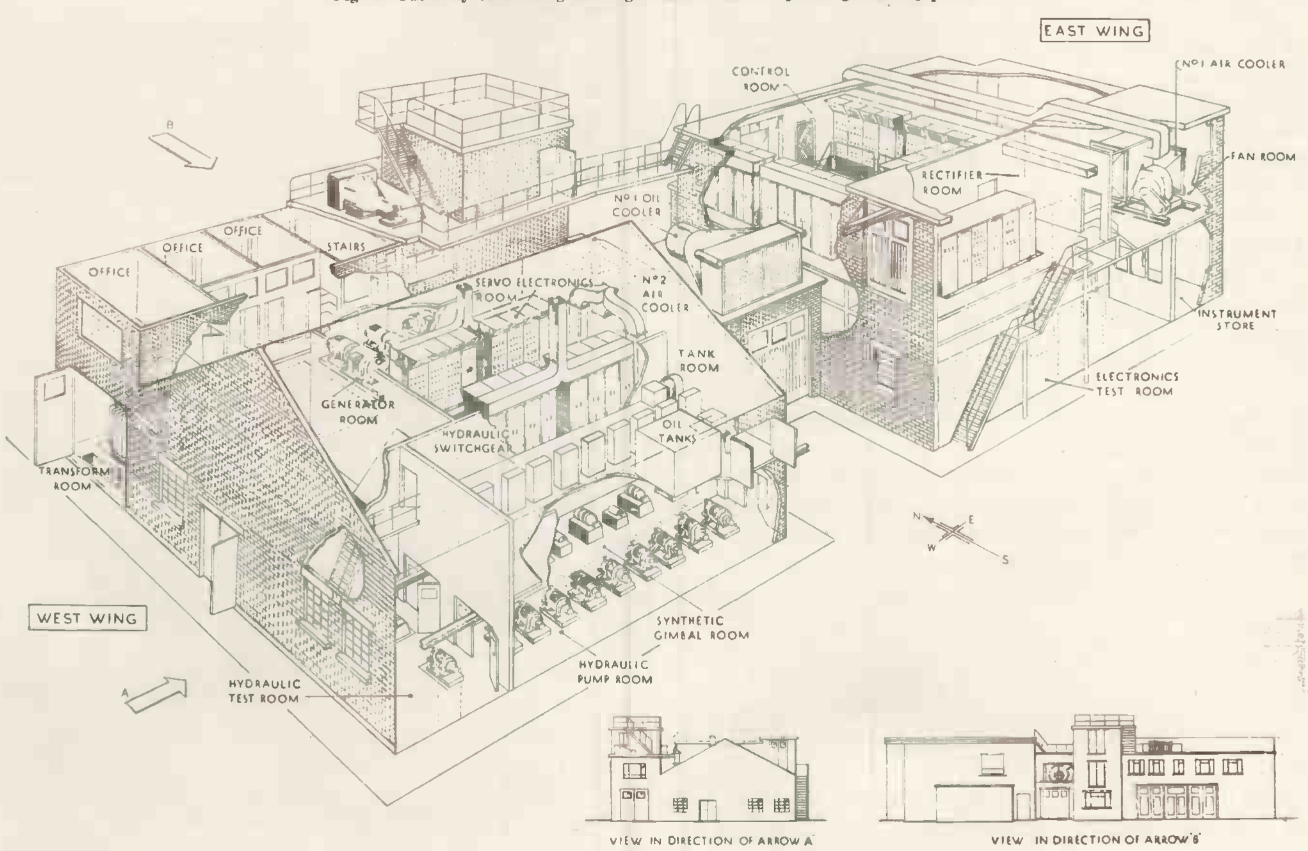
where u, v, w are the velocities along, and p, q, r are angular velocities about, the missile axes. The aerodynamic derivatives $\partial Y/\partial u$ etc. can be non-linear in the sense that they can be continuously varying as functions of incidence or sideslip, Mach number, altitude, etc.

(3) The equations of motion of the missile as derived from the forces along, and the moments about, the missile principal axes. These equations are of the form:

$$\left.\begin{aligned} u' &= (X/m) + (T/m) + g_x + rv - qw \\ v' &= (Y/m) + g_y + pw - ru \\ w' &= (Z/m) + g_z + qu - pv \end{aligned}\right\}$$

$$\left.\begin{aligned} p' &= (L/A) - [(C - B)/A)] \,.\, rq \\ q' &= (M/B) - [(A - C)/B] \,.\, pr \\ r' &= (N/C) - [(B - A)/C] \,.\, qp \end{aligned}\right\}$$

Fig. 1. Cut-away of building housing TRIDAC and its power generating plant



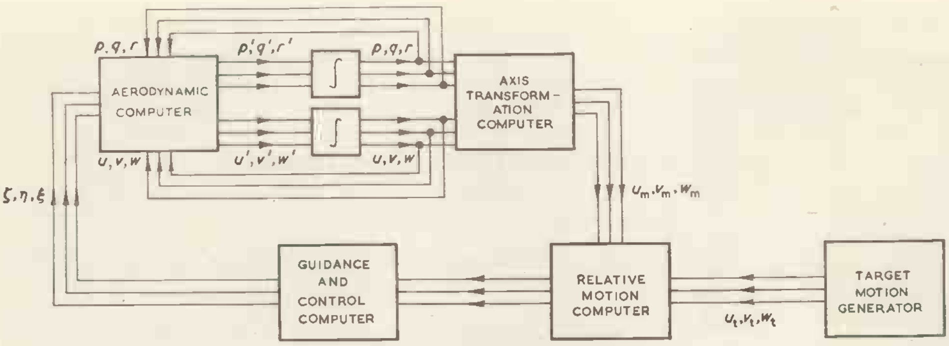


Fig. 2. Basic simulator layout

where $g_{x,y,z}$ are the gravitational acceleration components along the missile principal axes and T is the thrust along the missile X-axis. In the simulation of these equations the mass m, the moments of inertia A, B, C and the thrust T can be varied as functions of time.

(4) The control and guidance systems. These parts of TRIDAC are really flexible computers which may be set up to represent a wide variety of systems, including beam-riding and homing.

A major problem in the simulation of flight is that of axis transformation. The aerodynamic forces and moments are known only with reference to the missile axes, which are translated and rotated with the missile, while gravity forces, and in certain cases the target data, are known only with reference to axes fixed in the earth. To evaluate the motion of the guided missile, that is, to interrelate suitably the four major items outlined above, facilities must exist for making transformations of data between missile and earth axes. This is discussed in detail later, but for the present it may be noted that the process of axis transformation is essentially that of obtaining and then using the direction-cosines (l_1, m_1, n_1), (l_2, m_2, n_2), (l_3, m_3, n_3) of the fixed earth axes x_e, y_e, z_e relative to the moving missile axes x, y, z. The way in which the direction-cosines change during flight may be derived from a knowledge of the initial conditions for the flight together with the computed rates of turn p, q, r of the missile about its axes x, y, z.

The framework for the simulation of the flight of a guided missile is as shown in Fig. 2.

In the "aerodynamic computer"; knowing the three control surface angles (aileron ξ, elevator η, rudder ζ), the component velocities u, v, w and p, q, r, and with information on mass, moments of inertia, centre of gravity position, aerodynamic derivatives, altitude, gravity components etc.; the forces X, Y, Z and moments L, M, N acting on the missile may be computed and the force and moment equations solved to give the linear (u', v', w') and angular (p', q', r') accelerations as outputs. These accelerations are then integrated and fed back to satisfy the original assumption that the component velocities were available.

In the "axis transformation computer" the way in which the missile is turning about its own axes (given by p, q, r) is used to define the direction cosines linking the moving missile axes and a set of axes fixed in the earth. The linear motion of the missile relative to its own axes (u, v, w) may then be resolved into linear motion relative to the earth (u_m, v_m, w_m).

The "target motion generator" defines the flight path of a target aircraft in the earth axes (u_t, v_t, w_t) and the motions of missile and target are compared in the "relative motion computer". Some aspects of this relative motion are measured by the missile guidance system and used as instructions which pass via the missile control system to the servomechanisms which operate the missile control surfaces.

Although the basic problem for which TRIDAC was designed is that of simulating in the above way, in considerable detail, and in real time an attack on an aircraft target by a guided missile, it can be used for a variety of problems relating to the flight of missiles or aircraft. Within each of the main computing blocks of the machine, such as those shown in Fig. 2, a high degree of flexibility prevails, though the inter-connexions between such blocks are to a large extent fixed by the overall type of problem for which the machine was designed and consequently are more or less permanently fixed features of the machine. Outputs, that is records of the analogue voltage variations, may be taken from the machine in a variety of ways, ranging from conventional records against time of the variables to pictorial displays on cathode-ray oscilloscopes. In certain problems, plan and elevation plots of the target and missile trajectories are of interest.

The Basic Computing Elements

In detail the computing reduces to combinations of simple operations such as addition, subtraction, integration and differentiation with respect to time, multiplication, division, resolution by sine and cosine etc., and basic designs for electronic analogue units which will perform such operations are well known[1,2].

In TRIDAC the bulk of the computing elements are based on the use of high-gain, directly-coupled and drift-stabilized amplifiers having negligible input grid current. Such an amplifier when connected with input and feedback impedances so as to have an external gain which is low compared to its internal gain provides an output voltage

Fig. 3. Brick unit

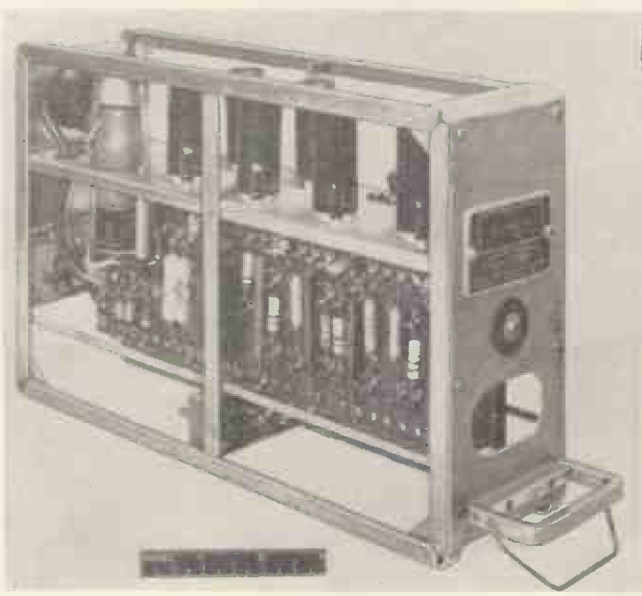


Fig. 4. Control room

which maintains the feedback current equal and opposite to the total input current. With various combinations of amplifiers and impedances it is therefore possible, due regard being paid to polarity, to carry out the operations of addition, subtraction, integration and differentiation of voltages with respect to time, and multiplication by constants. Certain non-linear relationships, such as $y = Kx^n$, can also be obtained using diode-resistor combinations in association with such amplifiers[4]. The identity $4xy = (x + y)^2 - (x - y)^2$ is used by electronic units which yield products of two variables rapidly and accurately. The aim throughout has been to keep the errors in any one computing unit below 0·1 per cent of the full-scale output of that unit. The entire electronic equipment of the machine is housed in some 2 000 small plug-in "brick units" like the one shown in Fig. 3. Part of the control room, which contains about two-thirds of the electronic equipment is shown in Fig. 4 and the control desk itself is shown in Fig. 5.

The electronic units of greatest general interest are the computing amplifiers, of which some 600 are available, and some details of these will now be given.

The accuracy requirement demands that the amplifier shall have a high gain over the frequency band from zero to several hundred cycles per second, besides very good zero stability to avoid integration errors and the necessity of constant manual resetting. To achieve this, a special amplifier is used consisting of a d.c. amplifier and a drift-correcting a.c. amplifier. The feedback drift error voltage at the input of the d.c. amplifier is modulated, amplified, rectified, smoothed and fed back to the d.c. amplifier to reduce the error. The drift of the d.c. amplifier, which has a good frequency response, is thus corrected by the a.c. amplifier, which has a low drift but a poor frequency response. The computational accuracy of the complete drift-corrected amplifier is largely determined by that of the resistors in the feedback and input circuits. These resistors are wire-wound and sealed in capsules to ensure both accuracy and stability. The allowed error in resistance is 0·1 per cent.

The circuit of the d.c. amplifier is shown in Fig. 6. It comprises three stages of amplification and a cathode-follower output stage giving an overall internal gain of 60 000 and a very low output impedance. The input stage is a cathode-coupled pair, the first valve of which is chosen to keep the grid current below 10^{-11}A. The drift-correcting voltage from the a.c. amplifier is applied to the grid of the second valve. By using either one or three valves in the cathode-follower output stage, maximum output currents of 4 or 12mA are available. The utilized output range of the amplifier is ± 30V.

Fig. 7 is a block schematic of an amplifier with its drift-corrector unit, in which the amplifier is shown with zero input signal and with the drift potential appearing as an injected potential V_d. The drift-corrector unit has a gain of β and the d.c. amplifier a gain α. With the corrector amplifier disconnected the drift error voltage, V_o, appearing at the output is:

$$V_o = \frac{(Z_1 + Z_2)\alpha}{Z_1 + Z_2(1 + \alpha)} \,.\, V_d \simeq \frac{Z_1 + Z_2}{Z_2} \,.\, V_d \text{ when } \alpha \text{ is large.}$$

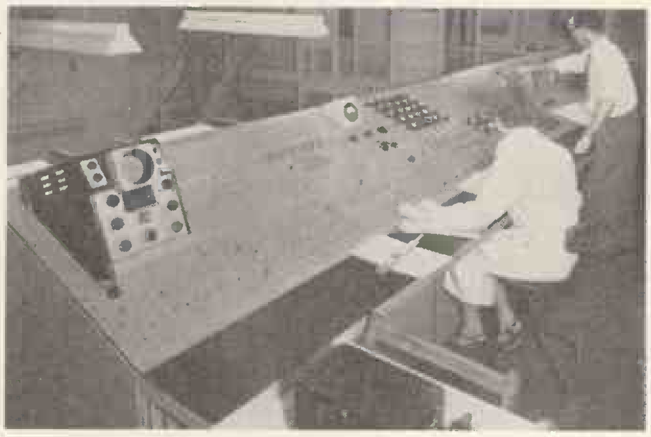
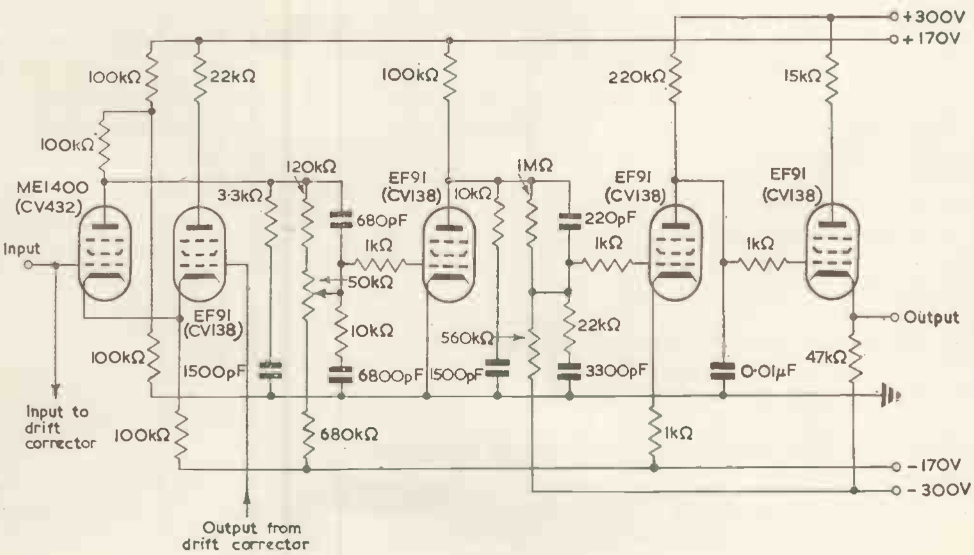


Fig. 5. Control desk

Fig. 6. D.C. amplifier circuit



When the amplifier gain is large the drift is therefore not reduced by a further increase of gain.

With the drift-corrector unit connected the drift error, V_o, is:

$$V_o = \frac{1}{1+\beta}\left(Z_1/Z_3 + \frac{Z_1+Z_2}{Z_2}\right) . V_d$$

In practice the decrease in drift achieved by the introduction of the $(1 + \beta)$ term more than offsets the increase in drift caused by the introduction of the input impedance,

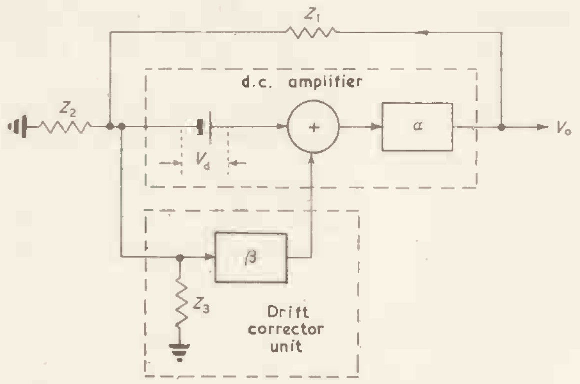


Fig. 7. Amplifier and drift corrector

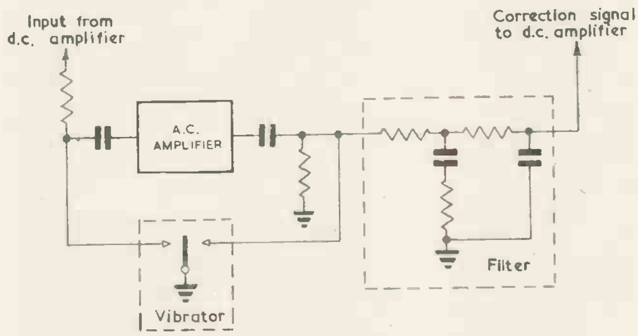


Fig. 8. Relay drift corrector

Z_3, of the corrector unit. A limit to the drift correction obtainable is, however, set by the drift errors introduced by the modulation and demodulation system.

The error voltage at the input to the d.c. amplifier may be converted to alternating current either by a relay chopper unit or a magnetic modulator unit. The latter gives poorer drift correction but requires less maintenance since it has no moving parts. The relay chopper unit is used at points where the drift must be kept as small as possible, e.g. in integrators and in reproducing exact geometrical relationships, whereas the magnetic modulator unit is used at less critical points, e.g. in reproducing aerodynamic relationships.

A diagram of the relay chopper a.c. amplifier is shown in Fig. 8. The relay vibrates at 400c/s, and modulates and demodulates the input and output respectively. The amplifier comprises two stages of amplification and a cathode-follower, and gives a low-frequency gain of about 1 000. The drift error at the output of the d.c. amplifier is held with this corrector to less than 1mV when the overall gain is unity.

The magnetic-modulator type of corrector is shown in Fig. 9.

The magnetic modulator is essentially a double iron-cored a.c.-excited unit of conventional design. The a.c. coils are wound one on each core, while the control winding covers both cores. The direct-voltage input is applied to the control winding, thus causing current at even harmonics of the excitation frequency to flow in that coil, the algebraic sum of the positive and negative even-harmonics being approximately proportional to the applied direct voltage. The output of the modulator appears across the resistor *r*. These even harmonics are then amplified in a valve amplifier and detected in a phase-sensitive peak rectifier so that the output is a direct voltage proportional to the input, and as such it may be applied to the d.c. amplifier as a drift-correcting signal. This unit maintains the drift error at the output of the d.c. amplifier within 2mV when the overall gain is unity.

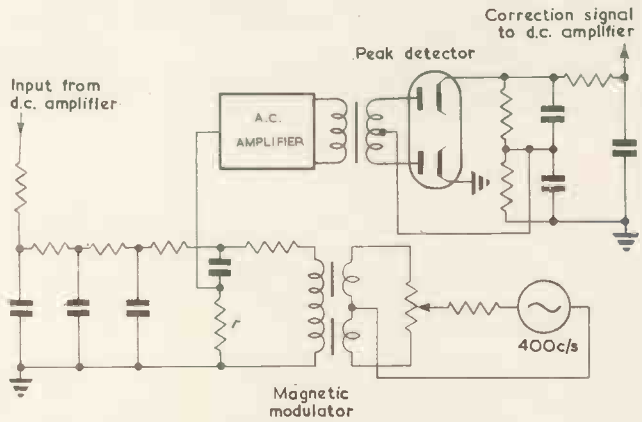


Fig. 9. Magnetic modulator drift corrector

Unique Features of TRIDAC

The features which made TRIDAC unique among simulators are the following:

(1) Its problem capacity and accuracy. The machine has of the order of ten times the capacity and is about ten times the physical size of any other analogue machine in this country and great care has been taken to achieve accurate computation throughout.

(2) Facilities are provided whereby many of the parametric variations, such as varying mass, and non-linearities, such as the variation of aerodynamic properties with missile altitude, of real systems can be reproduced.

(3) TRIDAC alone of simulators in this country can perform accurately and rapidly the calculations appropriate to resolutions between two sets of three axes in relative motion.

(4) The large size of the machine has necessitated the use of elaborate monitoring and fault-finding systems not necessary to the functioning of smaller machines.

The problems of varying parameters, axis transformation and monitoring as they apply to TRIDAC will be considered in Part 2.

REFERENCES

1. Wass, C. A. A. Introduction to Electronic Analogue Computers. (Pergamon Press Ltd., 1955).
2. Korn, G. A., Korn, T. M. Electronic Analogue Computers. (McGraw-Hill, 1952).
3. Spearman, F. R. J., Gait, J. J., Hemingway, A. V., Hynes, R. W. TRIDAC, A Large Analogue Computing Machine. *Proc. Instn. Elect. Engrs.*, *103*, Pt. B, 375 (1956).
4. Burt, E. G. C., Lange, O. H. Function Generators Based on Linear Interpolation with Applications to Analogue Computer. *Instn. Elect. Engrs. Monograph* 137M, (103 C.) (1955).

(To be continued)

The Lead Sulphide Photo-Conductive Cell

By M. Smollett*, Ph.D. and J. A. Jenkins*, M.A.

A theoretical and practical description of the lead sulphide photocell is given together with a number of possible applications.

SINCE the nineteenth century it has been known that certain materials have the power of changing their resistance on exposure to light. Such materials are known as photo-conductors. Lead sulphide possesses this property and when suitably prepared it will respond to light radiation in the wavelength region $0{\cdot}3\mu$ to $3{\cdot}5\mu$.

It is not until recently that lead sulphide has been developed as an infra-red detector and much of the initial work was done in Germany during the last war for military applications. However, research has now produced a highly efficient detector of the near infra-red for industrial purposes. A typical lead sulphide photocell is shown in Fig. 1.

Fig. 1. A typical lead sulphide photocell

The Theory of Photo-Conductive Cells

A lead sulphide photocell consists essentially of a thin layer of lead sulphide about 1μ thick, deposited on a glass substrate. The resistance of such layers is about $2\text{M}\Omega$ or more.

The properties of a photo-conductive layer may be described with reasonable accuracy by the simple energy band theory of semi-conductors. According to this the electrons in a crystal are packed into discrete bands of energy separated by gaps which are entirely empty of electron levels in the pure material, but which may contain energy levels due to foreign atoms or defects if the crystal is imperfect.

The top-most filled band is called the valence band because the electrons in it provide the valence bonding of the crystal; the next band after this is the conduction band which in semi-conductors generally contains few electrons, but these electrons are free to move and conduct charge. The energy required to bridge the gap between these two bands is called the activation energy of the solid (see Fig. 2).

When a quantum of light with an energy greater than the activation energy falls on a solid and is absorbed, the most likely process to occur is the raising of an electron from the valence to the conduction band, leaving behind an electron vacancy or "hole" in the valence band. The electron conducts charge as in metals and it may be shown that the hole conducts charge as if it were a positive electron. The result of the absorption of light is therefore to create new charge carriers which augment the conductivity. The new carriers will not exist indefinitely in the valence and conduction bands; they will have a certain probability of recombining. When the illumination is removed the increased conductivity will begin to decay. The time taken for the increase in conductivity to fall to $1/e$ of its equilibrium value is called the time-constant of the photo-conductor and is denoted by τ.

This picture of photo-conductivity is the simplest one that has been advanced and it sometimes called "single crystal" or "numbers" theory of photo-conductivity. Other ideas have been suggested to account for photo-conductivity in thin evaporated layers[2] such as a lowering of potential barriers between micro-crystals due to the space charge of trapped electrons and holes which have been produced by the light. It appears, however, that the simple theory is sufficient to account for the observed photo-conductive phenomena.

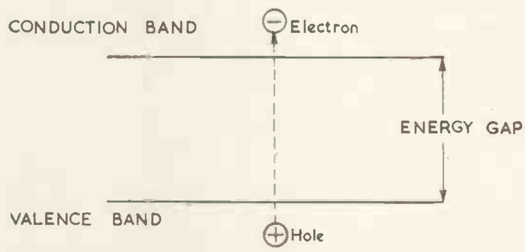


Fig. 2. Photon-induced transition in a semi-conductor, creating a hole-electron pair

The "numbers" theory of photo-conductivity may be shown to lead to the following approximate result for the change in resistance of a layer when it is irradiated

$$\Delta R_c = -2\,R_c^2\,N_o\,e\,\mu\,\tau \quad \ldots\ldots\ldots\ldots \quad (1)$$

R_c is the cell resistance; ΔR_c is the change in resistance caused by light; N_o is the number of hole-electron pairs generated in the layer per second; μ is the mobility of a charge carrier; τ is the time-constant; e is the electronic charge.

The above equation will provide an understanding of many of the phenomena observed in lead sulphide cells.

Operating Characteristics of Lead Sulphide Cells

SIGNAL

The lead sulphide cell is often used in a circuit similar to that shown in Fig. 3. A steady voltage of up to 250V is applied across the cell and a $1\text{M}\Omega$ load. The incident illumination is interrupted at a frequency of a few hundred cycles per second by a rotating, holed disk. The alternating voltage produced across the load is fed into an amplifier. The interruption of the light is a convenient way of separating the signal current from the standing current in the cell and load.

Small changes ΔV in the voltage across the load caused by a change ΔR_c in the resistance of the cell can be readily calculated by differentiating Ohm's law for the circuit giving

$$\Delta V = -\frac{V\,R_L\,\Delta R_c}{(R_c + R_L)^2} \quad \ldots\ldots\ldots\ldots \quad (2)$$

* *Mullard Ltd.*

where V is the voltage across the cell and load and R_L is the load impedance. On substituting for ΔR_c from equation (1)

$$\Delta V = \frac{2\, V_0\, R_L\, N_0\, e\, \mu\, \tau}{(1 + (R_L/R_c))^2} \quad \ldots\ldots\ldots\ldots (3)$$

This equation governs the behaviour of the cell in the circuit shown in Fig. 3.

The sensitivity of a lead sulphide cell is normally tested

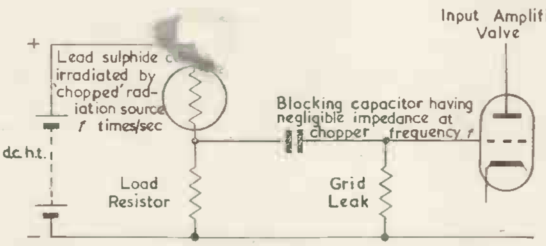


Fig. 3. Lead sulphide cell with chopped radiation source and a.c. amplifier

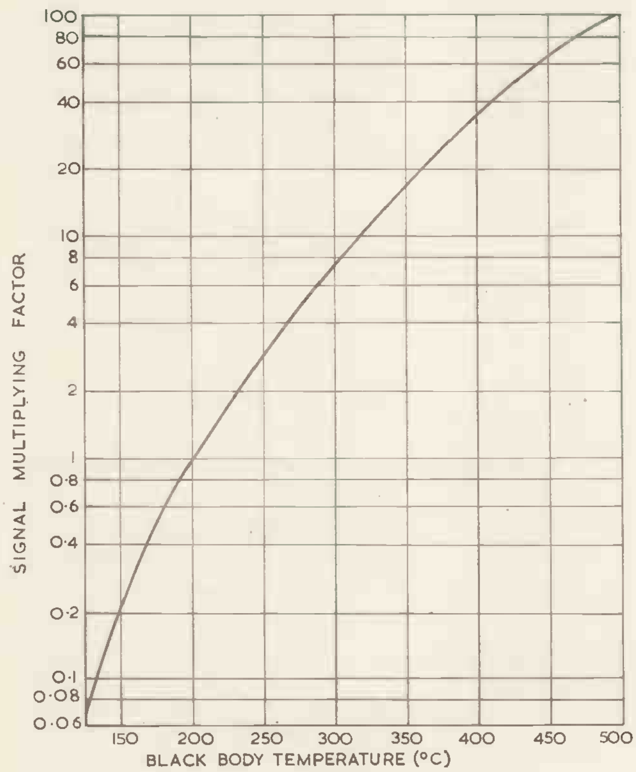


Fig. 4. Signal conversion factor with respect to 200°C

under two standard conditions: a black body test and a tungsten lamp test. In the black-body test the cell receives radiation from a 3·0mm diameter aperture in a 200°C black body situated 20cm from the cell. This system delivers 4·9μW of radiation power to the cell area. Under these conditions and at room temperature, about a millivolt of signal is produced across a 1MΩ load. The signal increases rapidly for black body sources at higher temperatures. For a 500°C black body the signal is increased a hundredfold over that at 200°C. A curve giving the signal conversion factor for various temperatures of the black body is shown in Fig. 4.

In the tungsten lamp test, 0·05 lumens from a lamp at the colour temperature of 2 700°K fall on the cell with 250V applied across the cell with no load in the circuit. The signal under these conditions is 3mA/1m change in the light intensity. In this test the light is not interrupted and the current is measured by a milliammeter placed in the circuit. The signal obtained from a cell varies with the wavelength of the light used. A spectral response curve illustrating this is shown in Fig. 5. The cell is only useful for wavelengths between 0·3μ and 3·5μ.

Lead sulphide cells have a response which is linear with increasing illumination and voltage, as long as too high a level of voltage or intensity is not reached. This linearity is to be expected from equation (3). Too great an intensity of light will heat the cell and affect its time-constant. A similar result is found for high voltages which also lead to a diminution in the signal.

The variation of the signal from a cell when the temperature is altered is shown in Fig. 6. The reason for the maximum at lower temperatures can be seen from equation (3). When the cell layer is cooled its resistance increases in

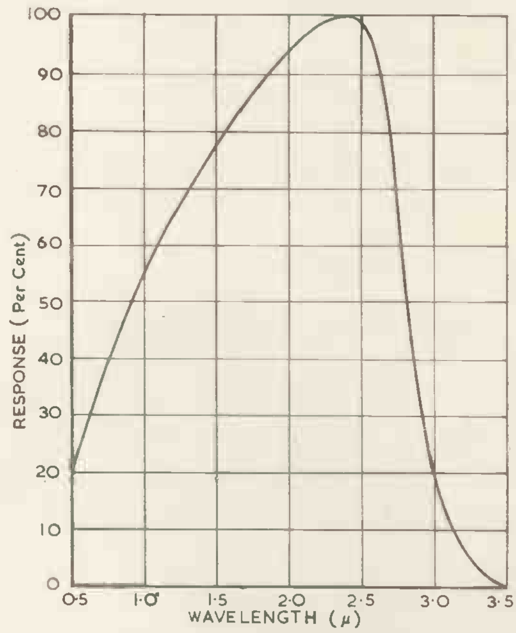


Fig. 5. Spectral response curve

common with all semi-conductors. Consequently $R_L/R_c \rightarrow 0$ and the denominator diminishes to a limiting value of 1. This makes the signal increase. In the numerator the only variables are μ and τ. Experiments show that in general after an initial increase the product $\mu\,\tau$ decreases as the temperature is lowered. This will cause the signal to fall. As a result of these two variations one gets the typical curve shown in Fig. 6.

The frequency of interruption has an effect on the signal given by a cell. This is because if the period of interruption is of the same order as the time-constant of the cell, the signal will not be able to build up to its maximum value. The relation of signal to interruption frequency may be shown theoretically to be given by

$$\Delta V_f = \frac{B}{(1 + 4\pi^2 f^2 \tau^2)^{\frac{1}{2}}} \quad \ldots\ldots\ldots\ldots (4)$$

where ΔV_f is the signal at a frequency f; τ is the time-constant and B is a constant for a particular cell. A typical curve of signal against frequency is shown in Fig. 7.

NOISE

In common with other semi-conducting materials lead sulphide layers produce a noise voltage when a current is passed, which is in excess of Johnson noise. This noise voltage, unlike Johnson noise, is frequency dependent. The mean square noise voltage per unit bandwidth $V_N{}^2$ of a cell may be represented by the following formula

$$V_N^2 = Ai^l R_o^m f^n$$

where A is a constant, i is the current flowing through the cell, R is the cell resistance and f the frequency at which the noise is observed. Experiment shows that the indices l, m, n have the average values 2, 2 and -1 respectively, giving for the noise voltage

$$V_N^2 = Ai^2 R^2 f^{-1} \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

These indices may vary widely from cell to cell.

When the voltage across the cell and 1MΩ load is 200V

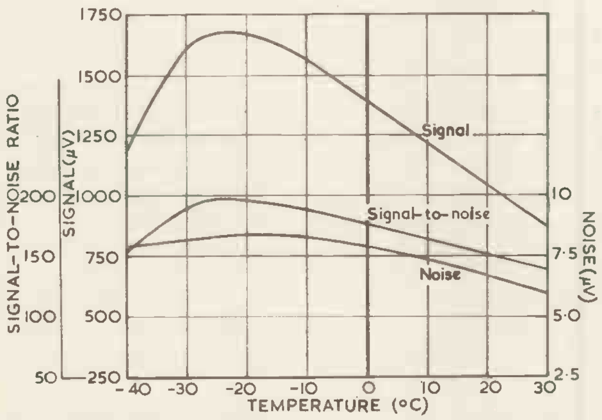


Fig. 6. Variation of noise, signal and signal-to-noise ratio with temperature

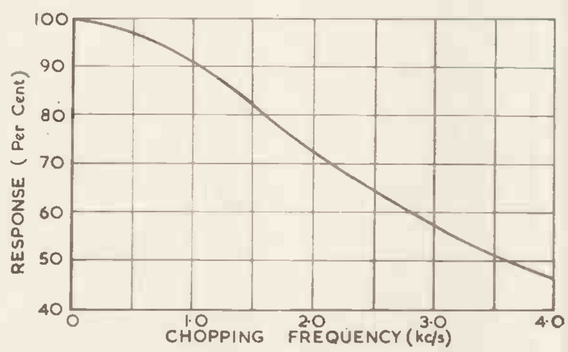


Fig. 7. Variation of signal with chopping frequency

and the root mean square noise voltage is measured at 800c/s in a bandwidth of 50c/s this noise voltage is found to amount to about 5μV.

For applications such as spectroscopy in which a good signal-to-noise voltage ratio is required it is evident from equations (4) and (5) that one cannot necessarily maximize the signal-to-noise ratio by reducing the frequency indefinitely, as the noise voltage increases faster than the signal voltage as the frequency is reduced. Equation (5) also shows that compromises must be made concerning the optimum applied voltage for a maximum signal-to-noise ratio, because the noise is voltage dependent. Curves showing signal, noise and signal-to-noise variation with applied voltage are shown in Fig. 8.

A useful measure of the performance of a cell with respect to noise is the minimum detectable power. This is the radiation power (of given spectral frequency distribution) which when falling on the cell will give a signal voltage equal to the noise voltage appearing across the cell. This power corresponds to the detection limit of the cell; a lower incident power would be lost in the noise voltage. For Mullard photocells this power is $3{\cdot}0 \times 10^{-9}$W (from a 200°C black body source). Some idea of this high sensitivity may be judged from the fact that, without any optical system, the cell will detect radiation from an ordinary soldering iron (temperature approximately 350°C) placed at a distance of 100 yards away.

When a cell is cooled the noise increases to a maximum. This can be seen from Fig. 6 where noise, signal-to-noise and signal are plotted against the temperature.

Applications of the Lead Sulphide Cell

The lead sulphide cell described here with its high sensitivity and fast response time has a great number of uses in industry and research. In radiation pyrometry very small temperature variations can be detected in temperature sources which are as low as 100°C and which may be some 100 yards away. The cell can therefore be used

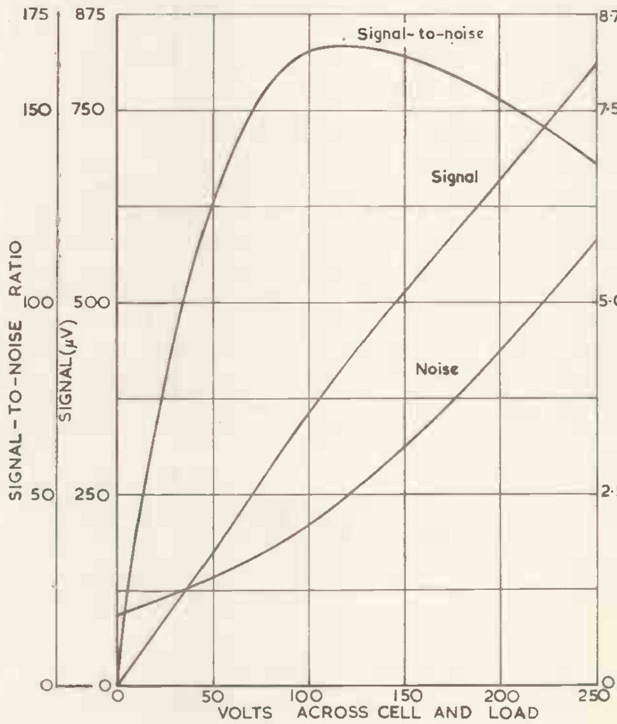


Fig. 8. Variation of signal, noise and signal-to-noise ratio with applied voltage

in measuring and monitoring temperature and temperature changes without actual contact with the work. One use in this field is the measurement of temperatures resulting from severe braking in moving vehicles and for detecting hot axle bearings. Gas, oil-fired and pulverized fuel furnaces may be monitored and the cell used to follow temperature fluctuations rather than the luminosity of the flame.

Any industrial control and protective system where visible light must be excluded may be conveniently worked with infra-red radiation and cells. Other important applications of the lead sulphide cell include intruder alarms, infra-red telephony and industrial or astronomical spectroscopy. The absorption of infra-red radiation by various gases opens possibilities for rapid gas analysis and humidity measurements.

These examples indicate the usefulness of the lead sulphide cell and show some of the profitable applications to which it may be turned in industry and research.

Acknowledgment

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REFERENCES

1. Elliot, A., *Electronics*, p. 97. (Pilot Press Ltd, London, 1947).
2. Smith, R. A. *Advances in Physics*. 2, 321 (1953).

Waveguide Components with Non-Reciprocal Properties

(Part 2)

The Behaviour of Ferrites in Waveguides

By J. Brown*, Ph.D., A.M.I.E.E., and P. J. B. Clarricoats†, B.Sc.(Eng), A.C.G.I.

The phenomenon of ferromagnetic resonance and the Faraday effect in an infinite medium have been described in Part 1; this article describes the behaviour of ferrites in waveguides and the operation of various waveguide components employing ferrites.

Circular Waveguides with Axial Ferrite Rods

As the theory given in Part 1 refers to propagation in an infinite ferrite medium a first step in designing waveguide components using ferrites is to examine the extent to which this theory applies in the waveguide case. It has been previously explained that when a plane electromagnetic wave is transmitted through a longitudinally magnetized ferrite medium, the plane of polarization of the wave is rotated. This is the Faraday rotation effect which arises from the different velocities of propagation of circularly polarized waves of opposite rotational senses. In a circular waveguide containing a longitudinally magnetized

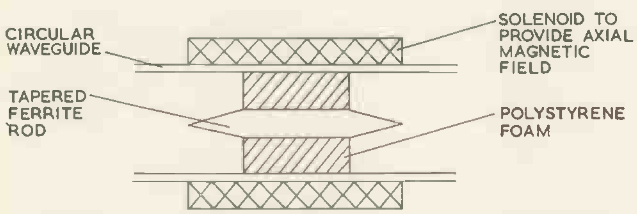

Fig. 6. Waveguide arrangement for the study of the Faraday rotation effect

ferrite rod (as shown in Fig. 6) the Faraday rotation effect is again observed, for although the propagating wave is no longer a plane wave it is nevertheless linearly polarized near the centre of the waveguide and may be decomposed into circularly polarized waves of opposite rotational senses which have different phase velocities. Owing to the high value of the dielectric constant which most ferrites possess at microwave frequencies, a small quantity of material introduced into a waveguide considerably perturbs the normal field pattern and propagation constant, even in the absence of applied field. One of the assumptions made in deriving the expression for Faraday rotation (see equation (23)) is that the r.f. magnetic field is transverse to the static magnetic field which is applied along the direction of propagation. In Fig. 7 the field pattern of the dominant (H_{11}) mode of circular waveguide is shown: near the guide centre the magnetic field is nearly transverse. If a rod of ferrite is placed coaxially in the waveguide and the diameter of the rod is sufficiently small compared with the waveguide diameter, then the ferrite does not perturb the field pattern appreciably, and it is reasonable to expect the Faraday rotation formula:

$$\theta' = \sqrt{(\epsilon/\mu_o)}\,(\gamma M_o/2) \quad \ldots\ldots\ldots\ (24)$$

to hold at least qualitatively.

Experimental evidence as well as more detailed theory show that the above statement is approximately correct provided that the ratio rod diameter/waveguide diameter is less than about 0·1. As the ratio rod diameter/waveguide diameter increases more energy passes through the rod and less through the surrounding air in the guide. This causes an increase in Faraday rotation together with an increase in transmission loss (assuming that the ferrite has some loss). In general the most desirable property of a Faraday rotation element for use in waveguide components of the types described in a later section, is that there should be considerable Faraday rotation with negligible transmission loss. Thus the parameter which is usually important

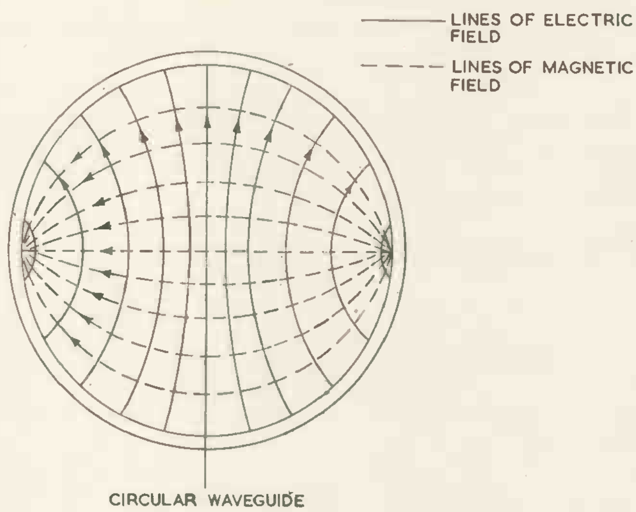

Fig. 7. Transverse field pattern of the H_{11} mode in circular waveguide

in describing the performance of a particular Faraday rotation component is the ratio rotation/loss. This parameter is normally independent of rod length but dependent quite critically on rod diameter. Experiment shows that the ratio rotation/loss reaches a maximum when the ratio rod diameter/waveguide diameter is equal to about 0·2 (for ferrites whose dielectric constant is about 10). One interpretation of this result is that above the maximum diameter the longitudinal component of the r.f. magnetic field in the ferrite increases more than the transverse component, causing the loss to increase without appreciably increasing the Faraday rotation.

An additional complication, which may arise in the waveguide case, is the excitation of higher order propagating modes. It is usual practice to choose the diameter of an air-filled waveguide so that only one mode (the dominant mode) can propagate. With the introduction of a ferrite into the guide, the effective diameter of the guide is increased and if there is sufficient ferrite, one or more

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higher order modes (in addition to the dominant mode) may propagate. If this happens the amplitude and rotation of the wave transmitted through the ferrite have a very irregular dependence on length, diameter and field strength.

Irregularities similar to those caused by the presence of higher order modes may occur due to multiple reflections of the dominant mode within the ferrite, unless special care is taken to avoid an impedance mismatch between the section of waveguide containing the ferrite and the empty waveguide. This matching problem may be solved in a number of ways, such as tapering the ferrite rod or using a quarter-wave matching transformer.

The theory which Polder[4] evolved to account for wave propagation in an infinite ferrite medium strictly applied to one which is magnetically saturated. In practice most of the change in Faraday rotation occurs before the ferrite is saturated and in fact the rotation depends on applied field in a similar way to the magnetization. Various authors have examined the case of propagation in an unsaturated ferrite and they show that the relation between B and H is still governed by a tensor permeability, in which the components of the tensor are now different from those quoted in Part 1 for the saturated ferrite.

The main differences between propagation in an infinite ferrite medium and propagation in a waveguide containing ferrite can be summarized as follows:

(a) The finite size of the ferrite causes the propagation constant to depend on the shape, size (particularly diameter) and the position of the ferrite in the waveguide.

(b) Complicated effects due to the propagation of higher order modes and multiple reflections may occur, respectively, if the ferrite cross-section is not small compared with that of the waveguide, and if the waveguide containing the ferrite is not matched to an air-filled waveguide.

(c) If the ferrite is operated in an unsaturated state the components of the tensor permeability relating B and H are not the same as those for the saturated ferrite.

Ferrites in Rectangular Waveguides

Consider the plan view of the magnetic field configuration of the dominant H_{10} mode of rectangular waveguide shown in Fig. 8. It can be seen that the transverse field is a maximum at the waveguide centre and is zero at the waveguide walls; conversely the longitudinal field is a maximum at the walls and zero at the centre. Along the planes cutting the figure in the lines AA and BB the longitudinal and transverse fields are equal in amplitude and 90° out of phase. Thus if the wave which propagates along the z-axis is viewed in the x-direction, it is circularly polarized at all points in the planes AA and BB. The senses of rotation are opposite for these two planes. If the static magnetic field is applied along the x-direction then a thin sheet of ferrite placed parallel with the plane AA presents a permeability $(\mu - k)$ to a wave propagating in the positive z-direction. Conversely in the plane BB the magnetic field configuration is rotating in an opposite direction and a ferrite strip placed there would present a permeability $(\mu + k)$ to a wave propagating in the positive z-direction. The permeabilities presented by the ferrite strips for a wave propagating in the opposite direction are the reverse of those quoted above since, viewed in the direction of the applied magnetic field, the r.f. vectors are now rotating in opposite directions. This forms the basis of many of the non-reciprocal components which employ transversely magnetized ferrites. It is important to note that the non-reciprocity in the transverse field case arises from the asymmetrical positioning of the ferrite, whereas the non-reciprocal properties of the longitudinally magnetized ferrite arise through the reversal of the direction of propagation with respect to the magnetic field and hence do not require any geometrical asymmetry in the waveguide.

A few remarks are again necessary regarding the validity of the above arguments in the practical case. Although the differences between the infinite medium and waveguide cases, previously summarized refer to the longitudinally magnetized ferrite in circular waveguide, they are for the most part applicable to transversely magnetized ferrite in either rectangular or circular waveguide. In practice the magnetic vectors are only circularly polarized on a plane which is nearly halfway between the centre and wall of the waveguide, in the absence of the ferrite. With a ferrite slab of quite small thickness (e.g. about 1/10th of the guide width), the fields in the ferrite are drastically modified and the polarization is generally elliptical. This causes the plane for maximum non-reciprocity to move nearer the walls as the slab thickness increases.

The Gyrator

Now that the behaviour of ferrites in waveguides has been briefly examined, various applications can be discussed. There are three principal waveguide components

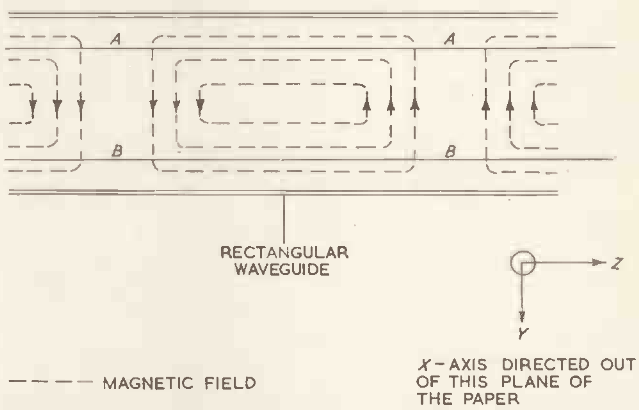


Fig. 8. Magnetic field configuration for the H_{10} mode in a rectangular waveguide at a plane which is normal to the direction of the applied static magnetic field

employing ferrites: the gyrator, the isolator, and the circulator. Each can be made to work either with a longitudinally magnetized ferrite in circular waveguide or a transversely magnetized ferrite in rectangular waveguide. In order to contrast the differences between the two methods, where they exist, the components are described in turn treating longitudinal field methods first.

The gyrator is defined as a component which causes a phase shift of 180° for one direction of propagation through it and zero phase shift for the reverse direction of propagation. In designing a gyrator it is sufficient to ensure that the phase shifts for propagation in the two directions differ by 180°: an extra length of empty guide can then be used to make the total phase shifts even and odd multiples of 180° for the two directions respectively, this being equivalent to the defined gyrator property. The first microwave gyrator to be developed employed the longitudinal field Faraday effect. In Fig. 9 a schematic view of such a circular waveguide gyrator is shown. The input wave is a linearly polarized dominant H_{11} mode in circular waveguide, the ferrite length and magnetization being arranged to cause 90° rotation of the plane of polarization of the forward travelling wave. A backward travelling wave, having the same phase at some plane on the output side of the ferrite as the forward travelling wave, is rotated by 90° in the same direction as before, when passing the ferrite, and thus arrives at the input with the same plane of polarization but differing by 180° in phase with respect to the forward travelling wave. The attenuating cards A_1 and A_2 absorb any energy which reaches the input or output with the wrong polarization.

A gyrator in circular waveguide is infrequently required but with suitable transitions from rectangular to circular waveguide the component may be used in a rectangular waveguide system. In this form the overall length of an X-band gyrator would be about 6in.

A more compact gyrator in rectangular waveguide can be made using the property, associated with an asymmetrically placed transverse magnetized slab of ferrite, that the phase constants are different for the two directions of propagation. A device of this type is shown in Fig. 10. In this case the length and magnetization of the ferrite are chosen so as to cause a differential phase shift of 180° between a forward travelling wave and a backward travelling wave.

It is not possible at the present time to make a theoretical comparison between the performance of the circular waveguide and the rectangular waveguide type of gyrator but in practice their performance at X-band appears comparable. The first gyrators were made in circular waveguide since ferrite rods were more easily obtainable than ferrite plates. Now that suitable ferrite plates are becoming available the rectangular waveguide type gyrator will usually be preferred in a system which uses rectangular waveguide since waveguide transitions will not be required in this case.

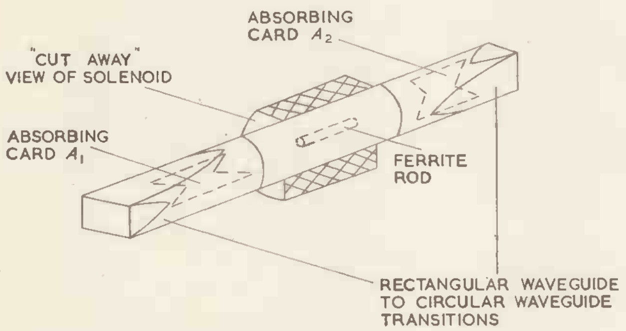


Fig. 9. Circular waveguide gyrator

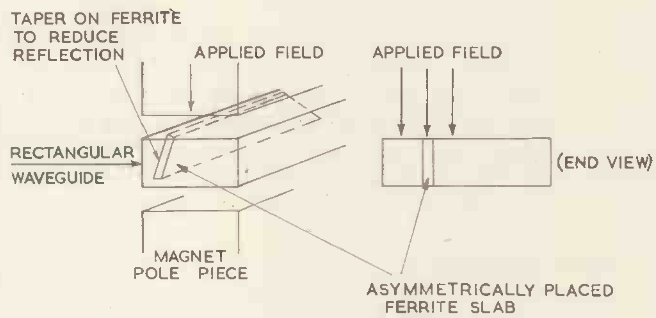


Fig. 10. Arrangement for rectangular waveguide attenuator or phase shifter employing a transversely magnetized ferrite slab

Isolators

An isolator is defined as a component which allows a wave travelling in one direction to pass with little or no attenuation but strongly attenuates a wave travelling in the opposite direction. The most obvious application for an isolator is as a pad between an oscillator and a load, which is not well matched to the waveguide. The forward wave travels from the oscillator to the load with little attenuation but the reflected wave from the load is attenuated before it arrives back at the oscillator, with a consequent reduction in any pulling effects. The wave reflected from the load can only be absorbed by the isolator: it cannot be reflected from the isolator, as may be proved by invoking the Second Law of Thermodynamics.

The circular waveguide version of the isolator is similar to the gyrator in Fig. 9, except that the planes of polarization of the input and output linearly polarized H_{11} modes now differ by only 45°. If rectangular to circular transitions are used, they also must be oriented to have corresponding directions differing by 45°. The input H_{11} mode in the circular waveguide has its electric field normal to the attenuating vane, A_1; the length and magnetization of the ferrite rod are selected to cause a rotation of 45° anti-clockwise in the plane of polarization of the forward wave, which thus reaches the attenuating vane, A_2, with the electric field normal to the vane, and is not attenuated by it. A wave in the opposite direction passes the vane, A_2, with its electric field normal to the vane and the 45° anti-clockwise rotation brings the electric field parallel to the plane of vane, A_1, in which the wave is therefore absorbed. The attenuating vane A_2 is necessary only when the circular waveguide isolator is used in conjunction with transitions into rectangular waveguide when the isolation may be reduced by multiple reflection effects (if this attenuating vane were omitted). Typical performance figures for an X-band isolator of this form are:—

Forward attenuation	0·5dB
Backward attenuation	30dB
Overall length	6in

In this type of isolator the backward wave power is absorbed in the attenuating vane and not in the ferrite in contrast to the absorption type isolator to be described next. In an isolator operating at high power levels it would be possible to use a sign-post junction at the input and in this way the backward wave power can be absorbed in a suitably cooled high power matched load. Such an arrangement will be mentioned again when the circulator is described.

An isolator in rectangular waveguide employs the non-reciprocal transverse field absorption effect associated with an asymmetrically placed slab of ferrite. The arrangement is very similar to that of the transverse field gyrator shown in Fig. 10, only the position of the slab for optimum backward to forward loss will not be in general the same as that for optimum gyrator performance. The principle of operation depends on the ferromagnetic resonance absorption of the wave which propagates in the direction for which the magnetic field is circularly polarized in a positive sense when viewed in the direction of the static magnetic field, applied transversely to the ferrite. At X-band the field strength for resonance is about $2{\cdot}5 \times 10^5$AT/m (3 000 oersteds) which can be quite conveniently produced by a permanent magnet. Typical performance figures for an isolator of this form are:

Forward isolation	0·5dB
Backward isolation	30dB
Overall length	2 to 3in

The saving in length compared with the Faraday effect isolator, is off-set at high power levels by the requirement for forced cooling of the ferrite which in this arrangement absorbs all the power from the backward wave.

Circulators

A third type of non-reciprocal component is called a circulator and can best be explained by the simplest example, a circuit with three pairs of terminals as in Fig. 11. This circuit is a three-arm circulator if it has the following properties:

(1) A signal input to arm 1 is completely transmitted to arm 2, there being no output from 3.

(2) A signal input to arm 2 is completely transmitted to arm 3.

(3) A signal input to arm 3 is completely transmitted to arm 1.

A possible circuit to give these properties is shown in Fig. 11, the sections *AB* and *BC* each being waveguides of length equal to a quarter of the guide wavelength and *CA* being a gyrator with zero phase shift for the direction *AC* and 180° phase shift for the other direction. If a signal is fed to arm 1, the signals arriving at *C* by the two paths are in anti-phase and hence cancel giving no output at arm 3. Similar arguments establish the properties (2) and (3).

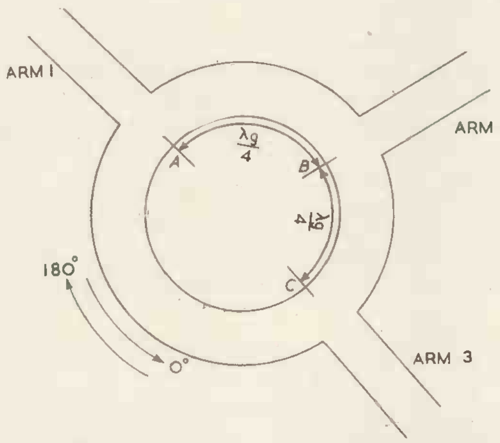


Fig. 11. Schematic diagram of three arm circulator using a gyrator

Two obvious applications for such a circuit are as a duplexer, and as an isolator. In the first of these, the transmitter, aerial and receiver are connected to arms 1, 2 and 3 respectively: theoretically this gives perfect isolation between the transmitter and receiver but in practice small phase errors in the sections would result in too large a signal reaching the receiver. A further limit on the use of any passive duplexer using non-reciprocal circuit elements is that the power reflected from the aerial always reaches the receiver. For an aerial with a v.s.w.r. of 0·8, a fairly typical value in radar installations, the signal level at the receiver resulting from reflection at the aerial would only be about 20dB below the transmitter power, considerably above the level which can be tolerated.

In the second application of the three-arm circulator the signal source is connected to arm 1, the termination being supplied to arm 2 and a matched load to arm 3. Any reflection from the termination is therefore dissipated in the matched load, so that the termination is isolated from the oscillator. This circuit is suitable for use as a high-power isolator since a high-power load can be used on arm 3.

In the gyrators described, reversing the direction of the magnetic field interchanges the directions for zero and 180° phase shifts. If the magnetic field is reversed in the circuit of Fig. 11, the sequence $1 \rightarrow 2 \rightarrow 3$ is changed to $1 \rightarrow 3 \rightarrow 2$. Accordingly, this circuit can be used as a waveguide switch, the signal fed to arm 1 leaving by arm 2 or arm 3 depending on the direction of the steady magnetic field applied to the gyrator.

Circulators with any number of arms can be made and a circuit for a four-arm circulator, designed by Hogan, is shown in Fig. 12. The operation of this component is as follows.

An H_{10} mode incident in the rectangular waveguide arm *A*, of signpost junction S_1, excites an H_{11} linearly polarized wave in the circular waveguide, whose plane is rotated by 45° in passing the ferrite. This wave excites an H_{10} wave in arm *B* of sign-post junction S_2 but ideally does not couple with arm *D*. A wave in arm *B* excites an H_{11} wave in a circular waveguide which is rotated by 45° in the same direction as before and couples into arm *C*. Similarly a wave in arm *C* couples into arm *D* and a wave in arm *D* couples into arm *A*.

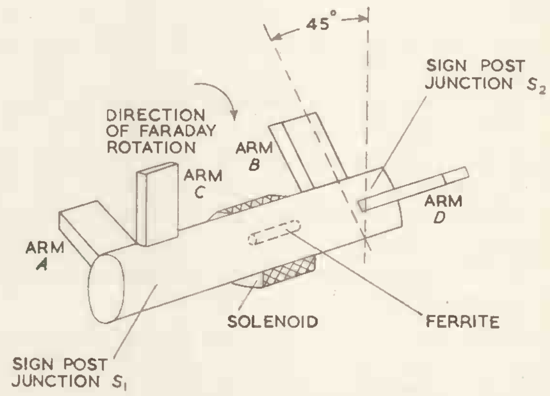


Fig. 12. Four arm circulator (after Hogan) utilizing the Faraday rotation effect

Other Components

In conclusion, brief mention can be made of a number of other important applications of waveguide components containing ferrite. The longitudinal field gyrator can be used directly as a controlled attenuator if a variable instead of static magnetic field is applied to the ferrite. If this field is periodically changed then amplitude modulation of a microwave carrier can simply be obtained. Similarly alternating magnetic fields can be applied to ferrite phase shifters which can be used to cause frequency modulation of those klystrons or magnetrons which can normally be tuned by mechanically adjustable short-circuits. "Modulated" ferrite phase shifters can also be used to produce electronic scanning of multiple feed microwave antennas.

REFERENCES

6. Hogan, C. L. The Ferromagnetic Faraday Effect at Microwave Frequencies and the Application of the Microwave Gyrator. *Bell Syst. Tech. J. 31*, 1 (1952).
7. Suhl, H., Walker, L. R. Topics in Guided Wave Propagation through Gyromagnetic Media. *Bell Syst. Tech. J. 33*, 579, 939, 1133 (1954).
8. Lax, B., Button, K. J., Roth, L. M. Ferrite Phase Shifter in Rectangular Waveguide. *J. Appl. Phys. 25*, 1413 (1954).
9. Fox, A. G., Miller, S. E., Weiss, M. T. Behaviour and Applications of Ferrites in the Microwave Region. *Bell Syst. Tech. J. 34*, 5 (1955).

Design of Computer Circuits for Reliability

By W. Renwick*, M.A., B.Sc.

The steps which can be taken to increase the reliability of a digital computer during the circuit design stage are indicated. Factors influencing the choice of components and mechanical design are discussed. The advantages of marginal checking are pointed out and the ways in which circuit design can facilitate this are discussed.

IN a large scale computing machine reliability is of first importance. A measure of the reliability of a particular machine may be got by comparing the hours of faultless work done with the total number of hours during which the machine was operating or being serviced. When the reliabilities of different machines are compared, account must be taken of the different operating speeds, as reliability should be related to the amount of useful work completed[1].

Among the steps which can be taken to increase reliability are the following:

(a) Use of test and diagnostic programmes to check the machine and to facilitate fault-finding.

(b) Preventative maintenance including marginal checking which enables an imminent fault to be found before it causes an actual failure of the machine.

(c) Precautions taken during initial circuit design including the mechanical construction and the choice of components.

This article is concerned only with (c) and it is hoped to indicate the philosophy behind the circuit design of EDSAC II, a large high-speed digital computing machine now being built at the University Mathematical Laboratory, Cambridge.

Failures in electronic apparatus may be divided into two main classes, permanent and intermittent. Permanent failures are usually caused by catastrophic component failures which are fairly easily traced and quickly rectified, for example, an open-circuit valve heater, a short-circuited capacitor or even a broken wire will, in general, all cause permanent failures. On the other hand, the intermittent fault is usually much more difficult to trace and might be caused by, among other possible troubles, a dry-soldered joint, a component or components drifting to the edge of their operating margin, variation in the voltage supplies or by circuits being sensitive to actual pulse pattern.

Choice of Components

The ideal component for use in a large computing machine, or of course in any electronic equipment, should satisfy the following conditions:

(a) Very long life, say greater than 100 000 hours.

(b) Characteristics defined within close limits.

(c) No drift in characteristics with temperature, operating conditions or time.

Secondary characteristics which, although desirable, are not so important in large computers, are:

(d) Small physical size.

(e) Low power consumption.

(f) Robustness.

No component at present available satisfies all these requirements, although the rectangular hysteresis loop ferromagnetic core and the semi-conducting devices offer attractive possibilities. The point-contact germanium diode is already well-tried and has been found reliable. The rectangular loop core and transistor have upper frequency limits and are confined to use in slower machines, although the advent of rectangular loop ferrites has greatly extended the use of ferromagnetic cores. While the semi-conducting devices have fairly low maximum operating temperatures, because of their low power consumption there is no difficult cooling problem involved.

For the fast large-scale computer there is no device which replaces completely the thermionic valve, although the valves used are far from having the ideal characteristics required for switching applications and seem to contradict all the conditions laid down above. The special quality valves now becoming available offer hope for some improvement in valve reliability[2]. A figure of merit for a valve to be used in direct-coupled switching circuits when the resistor and supply voltage tolerances are zero, or in capacitance or transformer coupled circuits, can be taken as

$$\frac{i_a}{E_c\,(C_{ae}+C_{ge})}$$

where i_a is the anode current flowing, E_c is the change in grid voltage required to cut off i_a, and C_{ae} and C_{ge} are the anode-earth and grid-earth capacitances respectively. This compares with the figure of merit

$$\frac{g_m}{C_{ae}+C_{ge}}$$

for a valve to be used in a wide-band amplifier. When the resistor and voltage tolerances are not zero, the performance of the valve depends on the anode voltage[3] and the figure of merit can be taken as

$$\frac{i_a}{(E_c+V_a)\,(C_{ae}+C_{ge})}$$

where V_a is the anode voltage required to maintain the anode current i_a. This means that in addition to requirements (a), (b) and (c), the valve should have high anode current for low anode voltage at the operating point, which may conveniently be taken at V_g equal to zero. It should also have a sharp cut-off and low inter-electrode capacitances. In addition a high heater to cathode insulation enables valves to be connected at different voltage levels without separate heater supplies.

There are three main types of resistor available, namely, the solid carbon composition grade 2 resistor, the cracked carbon high stability grade 1 resistor and the wire wound resistor. The solid carbon type is prone to severe drift in service, the Radio Components Standardization Committee specification allowing for 5 per cent drift during one year's shelf life and up to 15 per cent in operation, and is thus only of use in places where the value is absolutely non-critical, for example as grid or anode stoppers. The wire-wound resistor although it is extremely stable, drift from all causes

* *The University Mathematical Laboratory, Cambridge.*

being in general within 1 per cent, cannot be used in high speed applications, because of its inherent reactance. This leaves the grade 1 resistor as the only suitable component for this application. The R.C.S.C. specification for grade 1 resistors over $\frac{1}{2}$ watt rating and under 500kΩ is 2 per cent drift during one year's shelf life and $1\frac{1}{2}$ per cent drift during operation. It should be stressed that the drift in all types of resistors is independent of initial selection tolerance so that a nominally 1 per cent resistor can be supplied legitimately outside the specified tolerance. Since all cracked carbon resistors tend to drift to a higher value with time, it is possible to use this fact to compensate for drift in designing circuits. Resistors with values in the range from 1kΩ to 100kΩ being more stable should be used, if possible, and values greater than 1MΩ should be avoided. The boron carbon resistor promises a considerable improvement in stability over the standard grade 1 type. Variations in resistance are due to changes in temperature, humidity and load, and stability can be increased by derating a resistor and avoiding large temperature changes in the equipment.

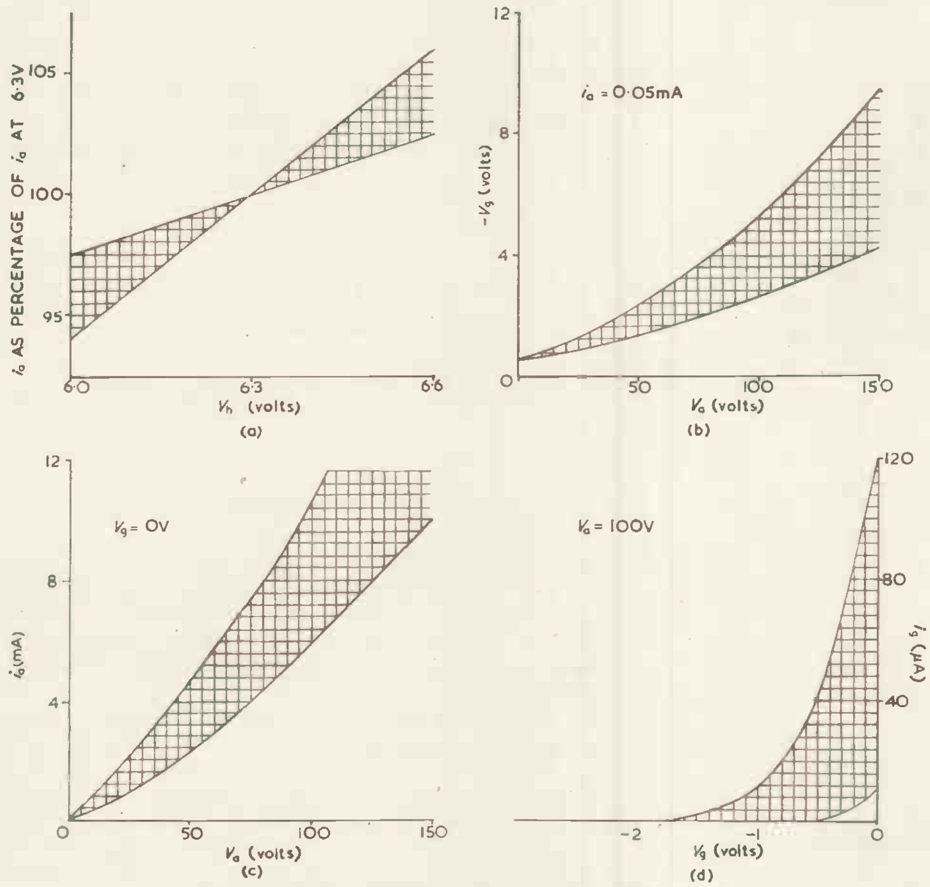


Fig. 1. Variation of characteristics of batch of 100 12AT7's

Notes on the Use of Valves

In the previous section the characteristics required by a valve for use in switching circuits were described. Here it is proposed to draw attention to the ways in which it is possible to increase the probability of a valve giving satisfactory service for a reasonable life-time. Most of the points which will be mentioned follow the manufacturers recommendations for the use of valves[4] but it is important that they should be followed, especially when long life is required and no apologies are offered for repeating them here.

Valves should be mounted in accordance with any restrictions imposed by the manufacturer, and it is especially important in the case of valves with close electrode spacing, for example valves with high mutual conductance should have the major axis of the grid vertical, to avoid electrodes sagging causing inter-electrode shorts. Care should also be taken that air is allowed to circulate freely and that the way in which valves are mounted does not obstruct their ventilation. It is a good rule that valves should not be mounted base upwards unless force cooled.

In general valves should not be operated without a d.c. connexion between each electrode and the cathode because of the possibility of secondary emission from the electrode concerned. In particular the resistance between control grid and cathode should be kept as low as possible and should never exceed the maximum value given in the published data for the valve.

The heater voltage (or current for series operated valves) should be kept as close to the nominal value as possible and should not vary by more than ± 5 per cent in any case. It is just as important not to under-run as over-run heaters as valve life is adversely affected in both cases. In addition low heater voltage results in reduced performance. The effect of varying heater voltage on the performance of a 12AT7 is shown in Fig. 1(a). Because of the thermal stresses in the heater during the warming-up period, it is advisable to switch the heater supply on in several steps. Another point to be noted is that it is undesirable to run valve heaters in series unless they are designed for constant current operation.

It should be remembered that the characteristics given in the published data are nominal only, and that the actual characteristics may depart from these by as much as ± 40 per cent. The histogram in Fig. 2 shows the distribution of valves with tolerance on anode current measured with zero grid bias and 100V on the anode, for 100 12AT7's In general it has been found that about half the number of valves of all types are initially more than 10 per cent from nominal. Thus to allow for initial tolerance and deterioration during life, design should be based if possible on a value of half the nominal. The variation of zero grid bias and 100V on the anode, for 100 12AT7's. in Fig. 1. Some characteristics depend only on the geometry of the valve structure, for example, cut-off voltage as a function of anode voltage for a triode, and not on the history of the valve, and in these cases allowance for initial tolerance only need be made.

Circuit Design

Apart from the choice of type, rating and siting of components, and in the case of valves following the recommendations of the previous section, there is nothing that can be done to reduce the chance of a permanent failure. On the other hand the intermittent fault offers scope for remedy at the initial circuit design stage. The design should be such that allowance is made for variations in component characteristics and voltage supplies and also that marginal checking can be easily and fruitfully applied. If marginal checking is to be of full advantage it should indicate an imminent fault due to the drifting of any component in the circuit, and it also may be of great help in tracing a fault to a particular unit or circuit. Both of these points should be borne in mind during the design stage. The other main aim in the choice of circuit design is the elimination of failures due to pulse-pattern or duty-cycle sensitive circuits, as these failures are usually among the most difficult to trace.

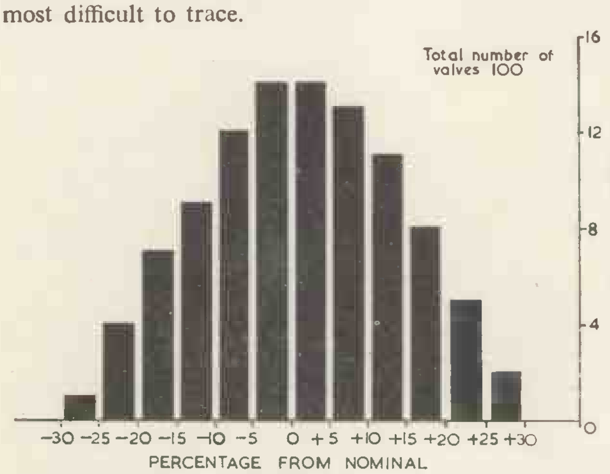


Fig. 2. Distribution of 12AT7's with tolerance on i_a at $V_g=0$, $V_a=100V$

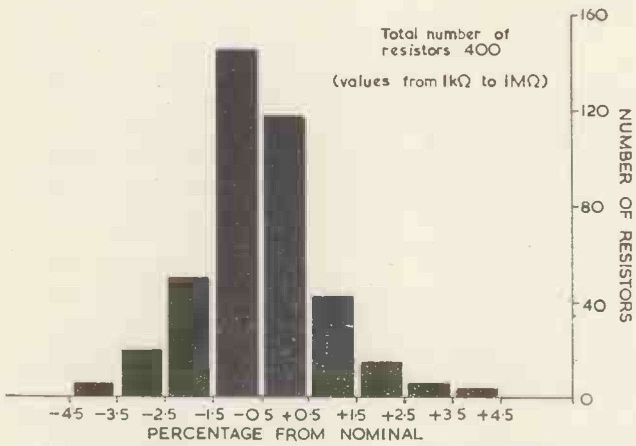


Fig. 3. Distribution of grade 1, 5 per cent resistors, with tolerance

Duty-cycle faults are caused by a time-constant somewhere in the circuit and may be due to a.c. couplings or to shunt capacitances. With a.c. coupling the error is due to the inability of the a.c. circuit to transmit the d.c. component of the pulse train or by the low-frequency cut-off of the system. The shunt capacitance causes circuits to operate too slowly, the error then being due to the high-frequency cut-off. The only sure cure for the low-frequency trouble is to use direct-coupled circuits throughout. Nothing can be done about the high-frequency cut-off except to keep the operating frequency low enough to eliminate this trouble. A useful marginal check which may be applied is to increase the transfer frequency (or decrease the transfer pulse width) and ensure that circuits are still operating correctly.

The use of direct-coupled circuits introduces some problems into the design which are best tackled by the method described in a previously published paper[3] where the design of a direct-coupled flip-flop is discussed, or by one of the other design methods already described[5,6]. Circuits which do not lend themselves to these or similar design methods are best avoided. It should be pointed out that circuit margins are only increased at the expense of speed of operation, the relationship is similar to the gain—bandwidth limitation for wide-band amplifiers. If the distribution of components inside the tolerance range is known, it is possible to decrease the initial design tolerances while still keeping very low the probability of getting an unsatisfactory circuit initially. The measured distribution of a group of 400 5 per cent grade 1 resistors in the range from 1kΩ to 1MΩ is shown in Fig. 3 and this shows that no resistors were found outside the tolerance range and that the chance of finding a resistor more than 3½ per cent from nominal is 1 in 50. If we consider the flip-flop circuit of Fig. 4 for example, the worst case for the tolerances on one side is obviously when one of the resistors R_2 and R_3 is high and the other low. The probability of finding two resistors, one whose value is more than −3½ per cent from nominal and the other more than +3½ per cent from nominal is over 1 in 10 000. Even under these conditions

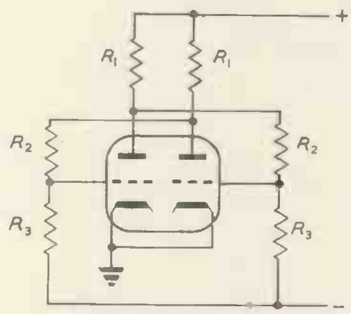


Fig. 4. Flip-flop circuit

the circuit will be stable unless the valve happens to be a bad one, and the probability, from Fig. 2 of the valve being worse than −25 per cent from nominal is 1 in 100. Thus with a circuit designed for these tolerances on resistors and valves the total probability of getting an unstable flip-flop is 1 in 10^6.

The reduction in design tolerances will obviously reduce the range over which components can drift without being replaced, but since most grade 1 resistors tend to drift towards a high value with time, the resistor drift will be self-compensating to a certain extent. Since resistors with a selection tolerance of 1 per cent may be supplied outside the tolerance limits and exhibit the same drift characteristics as those of greater selection tolerance, no long-term gain can be achieved by their use.

Marginal Checking

As has already been stated, marginal checking should give an indication that any circuit is operating near the edge of its margin and also it should, if possible, be of some help in locating a fault. The choice of circuit is limited to those whose behaviour, when subjected to the proposed marginal checking procedure, can be analysed. As an example consider again the flip-flop of Fig. 4.

The operation of this circuit is more readily understood if we take the current/voltage characteristic obtained when one anode resistor is removed and the current flowing into the circuit is measured as the voltage applied is varied, as shown in Fig. 5. When the voltage is zero the current will be $E_3/(R_2+R_3)$, the current flowing in the resistor chain only. As the voltage is increased V_1 will conduct (if the circuit has two stable states) and the current will increase being the sum of the anode current of V_1 and the current $(V+E_3)/(R_2+R_3)$ through the chain. When the voltage is equal to $[E_3(R_2/R_3)+E_C(1+R_2/R_3)]$ where E_C is the cut-off voltage of V_2, V_2 will start to conduct and a further increase in voltage will cause the current to decrease, as the drop in anode voltage of V_2 cuts off V_1, until V_1 is completely cut off. The current will again be given by $(V+E_3)/(R_2+R_3)$ until V is equal to $(E_3R_2)/R_3$ when V_2 will draw grid current and the current will then be equal to V/R_2. This characteristic is shown in Fig. 6 with the load line representing normal operation drawn in, giving the two stable operating points P and Q. Deterioration of V_1 will have the effect on the characteristic shown by the broken line in Fig. 6. Variation in R_2 and R_3 will move the point where V_2 starts to conduct as shown by the dotted lines in Fig. 6. The operating margin can be determined by reducing E_3 until the circuit has only one stable state at Q, and by increasing E_3 until the only stable state is at P. If in the

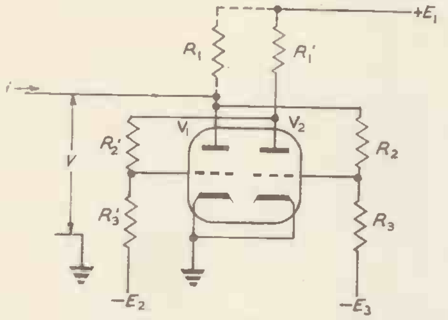


Fig. 5. Flip-flop circuit arranged to obtain current/voltage characteristic

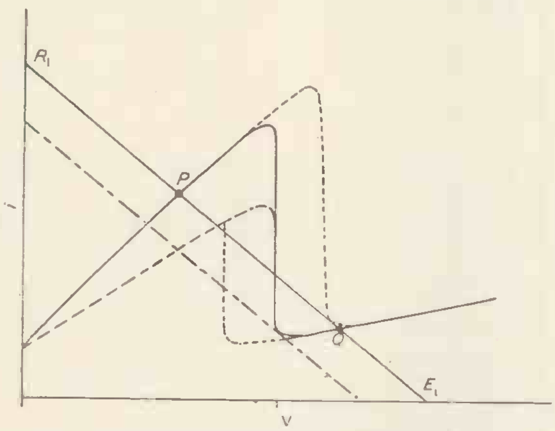


Fig. 6. i/V characteristic and load line

same way E_2 is varied the margins for the other half of the flip-flop can be checked. A useful marginal check for the flip-flop circuit of Fig. 4 is therefore to vary the negative return voltage of each chain independently or simultaneously in opposite directions.

Another point which must be watched is that circuits designed with wide margins should not feed circuits with narrow limits on input as this effectively cuts down the margin. For example consider the circuit of Fig. 7 which transfers the contents of one flip-flop V_1 to another flip-flop, V_4. For satisfactory operation of the gate formed by V_2 and V_3, the input swing required is at least twice the cut-off voltage of V_2. The flip-flop is quite stable with less than this voltage swing on its grid but components would have to be changed when marginal checking indicated that the transfer was not taking place. In this case the difficulty is easily overcome by taking the output from each grid of V_1 to the corresponding grid of V_2.

Not only does the marginal checking procedure for the flip-flop indicated above, check the d.c. stability of the circuit, but it also checks the trigger sensitivity margins.

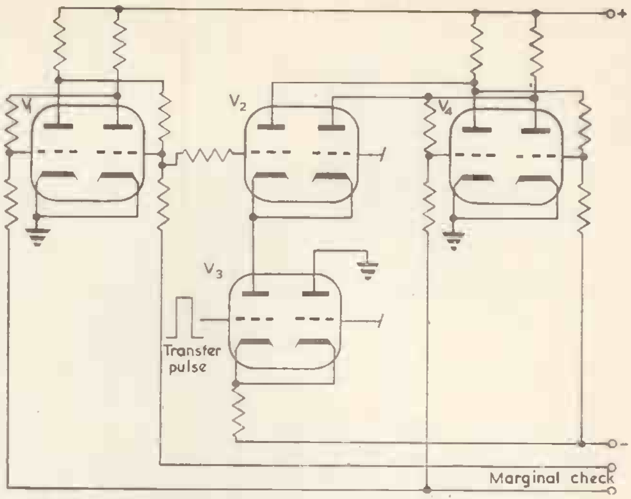


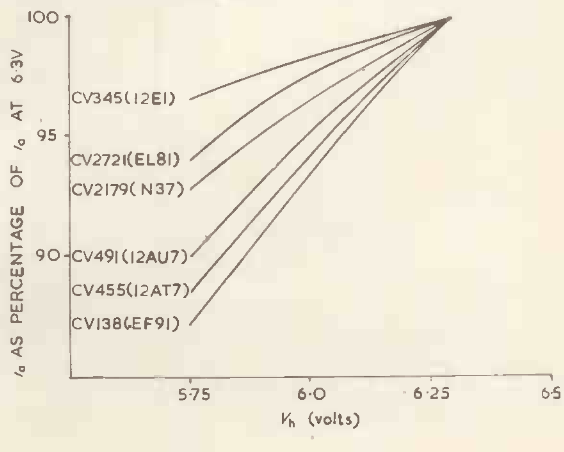
Fig. 7. Flip-flop V_1 feeding flip-flop V_4 via gate valves V_2 and V_3

Reducing E_2 and E_3 obviously makes a flip-flop triggered by the method shown in Fig. 7 more difficult to set or reset. To illustrate this consider the characteristic of Fig. 6 when the operating point is at Q. When current is drawn by the trigger valve from the anode, this has the same effect as reducing the h.t. voltage and if this reduction is great enough the load line will intersect the characteristic only once as indicated on Fig. 6. When the trigger current is removed the operating point will move to P and the circuit will have changed state. Thus the marginal check applied checks both the flip-flop and its associated trigger valves.

A marginal check which is sometimes applied is reducing the heater voltage but this is not to be recommended, unless the circuit is one to which other marginal checking techniques cannot be applied, not only because of the adverse effect it may have on valve life but also because a given percentage reduction in heater voltage has widely different effects on valve performance from type to type. This can be seen clearly from Fig. 8 where anode current as a percentage of anode current at the nominal heater voltage is plotted against heater voltage, for several common valve types.

Since very little data is available about the rate at which component characteristics drift, it is impossible to forecast the probable frequency of failures due to this cause, and to estimate how often marginal checking should be applied. Obviously the greater the amplitude of the margins applied, the less frequently need the machine be checked, but more often will components have to be

Fig. 8. Effect of reducing heater voltage for various valve types



replaced. Both the frequency with which marginal checking is applied to the machine and the amplitude of the margins with which it is tested will be determined empirically when the machine is built.

Mechanical Construction

Since the most important aim in the design of a large-scale computing machine is reliability, any steps which can be taken to reduce the time that a machine is being repaired are to be welcomed. There are two contradictory schools of thought on this point, one supporting the idea of plug-in units which can quickly be replaced when necessary by a spare, and the other supporting the view that faults should be repaired *in situ* and that to this end all components should be easily accessible to the maintenance engineer. The difficulty with the latter point of view would appear to be that even when a fault has been located to a particular unit, it may take some considerable time, which could be useful running time, to find the actual component or components which have to be replaced. When a plug-in construction is used there is still the choice to be made of the size of the replaceable unit. The advantages of as large a unit as possible would appear to be twofold. First, the pin-pointing of a fault need not be so exact and secondly, the number of plug connexions is reduced. When plugs and sockets are used care must be taken to ensure alignment and good contact.

It is important that the cooling of all components should be adequate and the mechanical design of the units should facilitate this. The siting of components is also important so that temperature sensitive components are kept as far away as practicable from components producing heat. A final point to be kept in mind is the layout of the wiring to reduce the number of soldered joints and stray capacitances to a minimum. The experiment of making all soldered joints butt joints so that a simple mechanical test can be used to discover dry-soldered connexions, is being tried on EDSAC II.

Conclusion

In this article an attempt has been made to point out the things which should be kept in mind during the initial design of a computing machine. It is by no means an exhaustive treatment of the subject but it is hoped that it will have given some idea of the philosophy behind the design of EDSAC II. Whether these ideas have been successfully put into practice or not, only the reliability record of the machine itself can tell.

REFERENCES

1. Elmore, W. B. Reliability in Electronic Data Processors. *Computers and Automation. 4*, 5 (1955).
2. Rowe, E. G., Welch, P., Wright, W. W. Thermionic Valves of Improved Quality for Government and Industrial Purposes. *Proc. Instn. Elect. Engrs. 102*, Pt. B, 3 (1955).
3. Renwick, W., Phister, M. A Design Method for Direct-Coupled Flip-Flops. *Electronic Engng. 27*, 246 (1955).
4. Radio Valve Practice. (B.V.A. 1948).
5. Rubinoff, M. Further Data on the Design of Eccles-Jordan Flip-Flops. *Elect. Engng. N.Y. 71*, 10 (1952).
6. Paivinen, J. O., Auerbach, I. L. Design of Triode Flip-Flops for Long-Term Stability. *Trans. Inst. Radio Engrs. Prof. Group on Electronic Comput. EC2*, 14 (1953).
7. Inter-Service Standards for Radio Components. RCG110, RCS111, RCS 112. (Radio Component Standardization Committee).
8. High-Speed Computing Devices. Ch. 17 (McGraw-Hill).
9. Symposium : Diagnostic Programs and Marginal Checking for Large Scale Digital Computers. *Convention Record Inst. Radio Engrs*. Pt. 7 (1953).
10. Pollard, B. W., Lonsdale, K. The Construction and Operation of the Manchester University Computer. *Proc. Instn. Elect. Engrs. 100*, Pt. 2, 501 (1953).

Ultrasonic Gauging in Shipyards

Ships are subject to corrosion by saltwater and on oil tankers the rate of wastage is aggravated by the action of crude oil and refined products, many of which are highly corrosive, carried within the holds. The thickness of ships' plating must be regularly measured and corroded plates replaced.

The normal method of checking for corrosion on ships is to drill holes to enable mechanical gauges to be inserted, after which the holes must be closed again by riveting or welding. The disadvantages of this process are obvious and the problem has recently been overcome by the introduction of an ultrasonic gauge, known as the Type 1101 Ultrasonic Thickness Gauge, made by Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5, which operates on the resonance principle. A variable oscillator within the instrument generates an ultrasonic signal, which is injected through a crystal probe into the plate, reflected by the far, inaccessible side of the plate and picked up again by the probe. If the outgoing and reflected signals are in phase, they will reinforce each other and resonance occurs. This condition is easily recognized by a sudden rise of the hum in the earphones, and by a sharp deflexion on the meter on the instrument.

Resonance can only occur if the wall thickness is a direct multiple of the wavelength of the signal. In operation, the probe is applied to the plate and the wavelength is slowly changed by the control knob, until resonance is achieved, when the thickness of the plate is read off a scale graduated directly in inches, opposite another scale showing the resonant frequency. The gauge is suitable for measurements from 0·06in to 12in in steel, and for a similar range in other metals.

Shell Tankers Ltd. decided some time ago to give this method a trial.

Several difficulties had to be overcome, the first being to devise a satisfactory method of cleaning a spot on the plate to be measured. The probe of the gauge must be applied directly to the metal and paint, scale and dirt must therefore be removed. Heavily pitted surfaces must be smoothed, to provide a sufficiently large area of contact between the probe and the plate, an air-driven grinder providing the most convenient solution.

Another difficulty arose over closely adhering scale on the far side of the plate which absorbed the signal, so that it became impossible to obtain an accurate reading. It was found, however, that a few vigorous hammer blows sufficiently loosened the scale to eliminate its damping effect.

The cost of ultrasonic gauging depends largely on the skill and speed of the gauge team, but there can be no doubt that by dispensing with staging, economies can be effected, two operators being able to gauge the hold of a tanker without any staging whatever, by taking advantage of the 50ft lead for the probe.

The cost of grinding and gauging should be appreciably less than that of drilling and making good and there is also a saving in time. With some experience, the complete ultrasonic gauging of a spot can be completed in about two minutes.

In the course of their tests, Shell Tankers Ltd carried out many random checks on ultrasonically gauged locations by drilling and direct mechanical gauging and showed that it is possible easily to obtain readings with an accuracy of $\pm 0{\cdot}005$in for plate thicknesses lying between 0·25 and 0·5in.

A Fast-Acting Phase Conscious Indicator

With Good Rejection of 2nd and 3rd Harmonics

By D. L. Davies*, B.Sc., A.M.I.E.E.

A phase-conscious rectifier circuit of the gated type is described. Its particular features are high rejection of 2nd, 3rd, and multiples of these harmonics, plus a high rejection of quadrature signals. A practical circuit is described in detail, and a short mathematical analysis of the mode of operation is included.

THE circuit to be described was designed to meet the need for a fact-acting phase sensitive indicator, with excellent rejection of at least 2nd and 3rd harmonic frequencies. Other requirements were that the instrument should ignore any quadrature signal in the presence of a reference phase signal, and should also ignore any spurious frequencies.

It was known that an instrument of the thermocouple wattmeter type would have an excellent performance as regards rejection of quadrature and harmonic signals, but the slow movement of this type of instrument was prohibitive for the application under consideration. Hence, a system of gated rectification was employed, being comparatively "fast-acting".

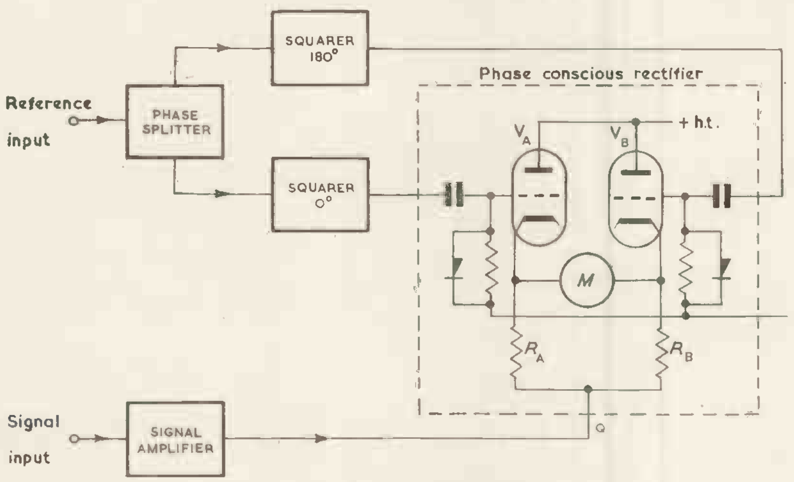


Fig. 1. Principle of operation

Principle of Operation

Referring to the block diagram in Fig. 1, a sinusoidal reference voltage is applied to the phase splitter stage, providing "push-pull" output signals. Each of these two outputs is fed into a squaring stage of adjustable mark-to-space ratio. In each case, these are adjusted to give output waveforms, as shown in Fig. 2(b) and (c). The two outputs are identical in shape, but displaced in phase by 180°.

These two gating waveforms are d.c. restored, so that they always move negatively with respect to a fixed d.c. potential P. They are then applied to the grids of two cathode-followers, V_A and V_B. The indicating meter is connected between the two cathodes.

The signal is applied to the common junction of the two equal cathode resistors, R_A and R_B, via a signal amplifier, which should have no phase shift at the frequency of operation. The junction of R_A and R_B should be at a d.c. potential Q which is lower than P by at least the peak signal voltage applied to the phase conscious rectifier.

When the grids of both V_A and V_B are in the "up" position, i.e. at voltage P, then both cathodes are held at equal potentials, and hence no current flows through the meter M. If, however, V_A is cut off by its grid moving negatively with respect to P, and the grid of V_B is still held at P, then signal current will flow through R_A and M to the cathode of V_B. Alternatively, if the grid of V_A is held at P and V_B is cut off, then signal current will flow through R_B and M to the cathode of V_A; i.e. in the opposite direction to the previous set of conditions. Thus, the signal is gated through M in opposite directions alternately.

N.B.—In addition to the signal current gated through M, there will also be a d.c. component due to the p.d. between P and Q. However, this is equal and opposite for each gating period and averages out to zero, the meter movement acting as the integrator.

Consider a sine wave signal of reference phase and frequency applied to the junction of R_A and R_B. From 2(d), it can be seen that, when the grid of V_A is held at P and V_B is cut off, then current flows through R_B and M during the portion with vertical shading. Current next flows when the grid of V_B is held at P and V_A is cut off; as shown by the portion with horizontal shading. It can be seen that the signal voltage for this second current flow is of opposite polarity to that of the signal during the first current flow. However, the direction of current flow through M has also reversed, so that the two currents add. The integrated current throughout a cycle is then read by the meter.

If a sine wave signal of reference frequency, but of quadrature phase, is considered, as in Fig. 2(e), it can be seen that the integrated current through M becomes zero. Hence, there is no response to any quadrature component of the input signal waveform.

Similar reasoning can be applied to signal waveforms consisting of 2nd or 3rd harmonic frequencies of any phase. It can be seen from Fig. 2 (f.g.h.i.) that the mean current through M is zero in each case. Thus, there is no response to any 2nd or 3rd harmonic components of the input signal waveform. The same can be said for all multiples of these harmonics. Harmonics which are not multiples of either 2nd or 3rd, i.e. 5th, 7th, 11th, etc., are reduced by a factor proportional to their order, e.g. 5th harmonic is reduced to 20 per cent of its value, 7th to 14 per cent, and so on. (See Appendix).

Circuit Description

The reference voltage is applied to a concertina type of phase splitter V_{1a}, the two outputs being fed into two

* *Solartron Research & Development Ltd.*

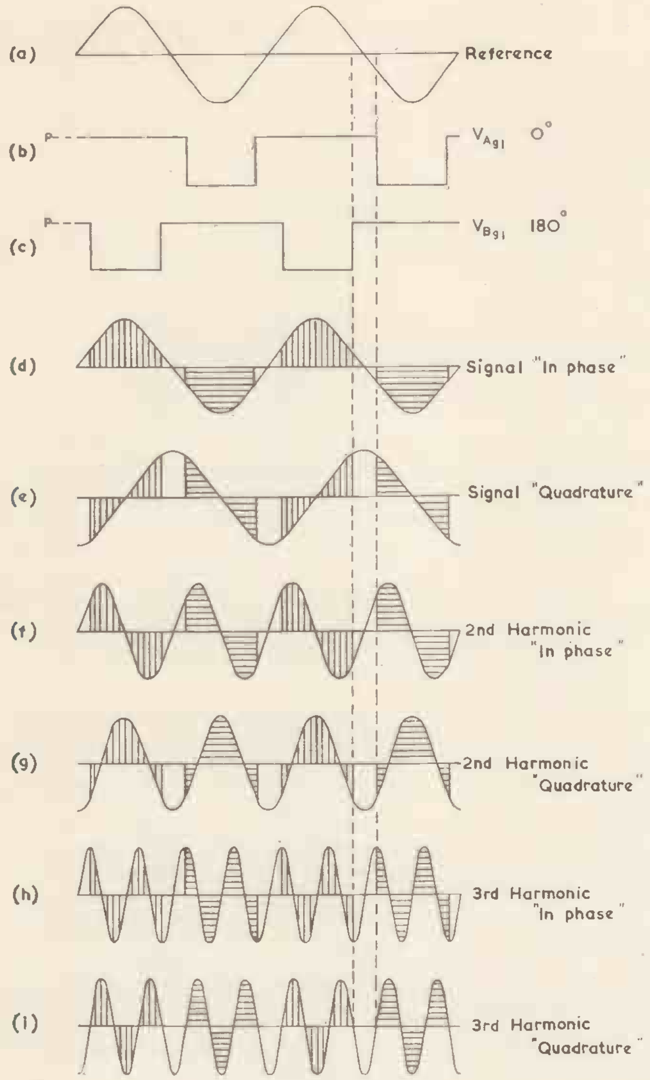


Fig. 2. Waveforms

identical cathode-coupled limiters V_2 and V_3. By means of VR_1 and VR_2, the output waveforms from these two squarers are adjusted to be as in Fig. 2(b) and (c) respectively, i.e. " up " for $\frac{2}{3}$rds of a cycle, and " down " for $\frac{1}{3}$rd; but displaced from each other by 180°.

These two gating waveforms are applied to the grids of V_4, being d.c. restored by two germanium diodes to move negatively with respect to point P, which is at about 60V positive.

Signal voltages are fed via a signal amplifier to a cathode-follower V_{1b}, whose output drives the return point of the cathode resistors of V_4. The quiescent voltage at this point is about 40V positive. The signal amplitude at this point is always sufficiently low so that positive peak signals will not lift the cathodes of V_4 above the potentials to which they are clamped when the gating waveforms are in the " up " state. Likewise, the gating waveform in the " down " state must be sufficiently low so that it keeps the valve which is being gated, off; even when peak negative signal is applied to the cathode return point.

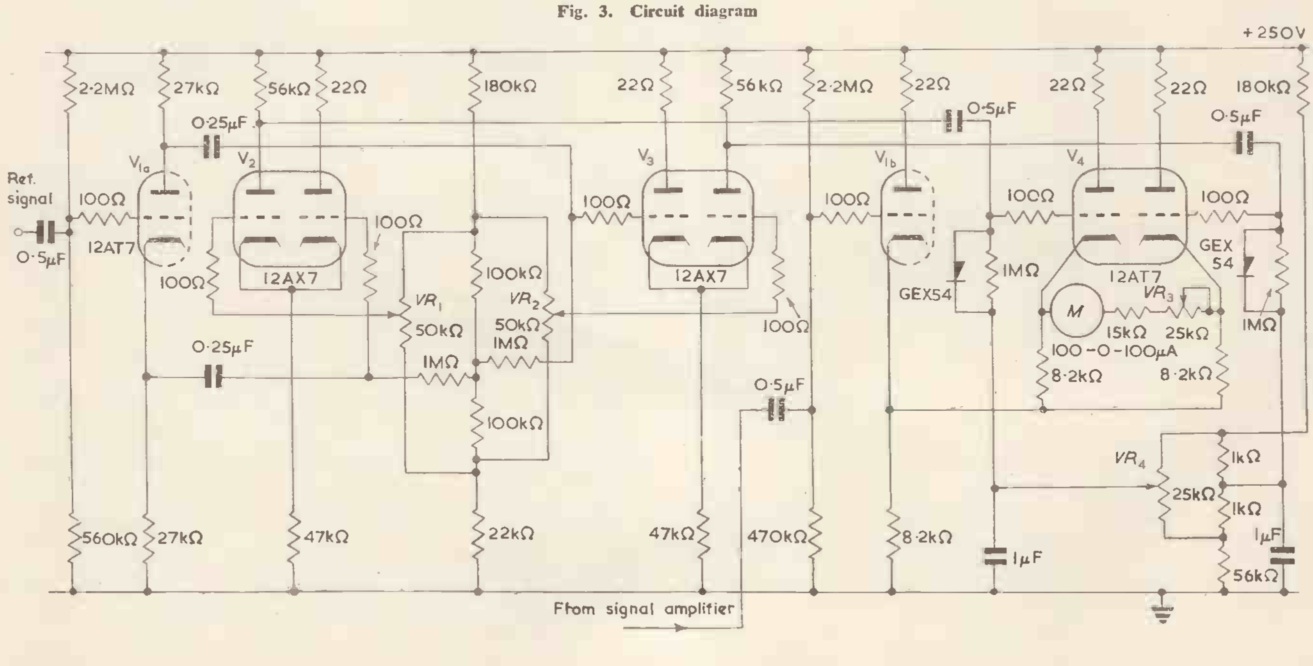
VR_3 is a variable resistor in series with the meter M, and is used to adjust the sensitivity of the indication. This resistor is common to both current paths, i.e. via the cathode resistors of V_{4a} or V_{4b}, so that it affects both current contributions equally.

In order to balance out the different quiescent levels at the cathodes of V_{4a} and V_{4b} due to their possibly different contact potentials, VR_4 is provided to adjust the return point of one grid, either positively or negatively with respect to the other. In the description of the operation of the circuit, this small difference was ignored, and the potential to which both grids were returned was referred to as point P, whereas in practice there is a slight difference between the two grid return points.

Performance

The circuit described was designed to operate over a reference frequency range from 60c/s to 2 400c/s. Its measured rejection of 2nd, 3rd and multiples of these harmonics was at least 100:1. 5th, 7th, 11th harmonics, etc. decreased in proportion to their order, as indicated earlier. Rejection of quadrature signal could not be measured accurately, but was better than 100:1.

Fig. 3. Circuit diagram



Appendix

The gating waveform which eventually samples the signal is made up of Fig. 2(b) and (c), with say (c) reversed. The phase conscious rectifier, consisting of V_4, can then be thought of as a multiplier between the gating waveform and the signal under test. Any harmonic rejection properties of the multiplier may then be examined in terms of the harmonic content of the gating waveform. The gating waveform is as in Fig. 4. Let it be considered as a Fourier series of the form:

$$f(t) = A_o + \sum_{n=1}^{n=\infty} [B_n \sin (n\omega t) + A_n \cos (n\omega t)]$$

By inspection, this waveform is symmetrical about the base line—hence there is no steady or d.c. term.

Again, by inspection, the function is an even one, i.e. is symmetrical about $t = 0$. Thus, there can be no sinusoidal terms in its Fourier series—sine terms being odd functions.

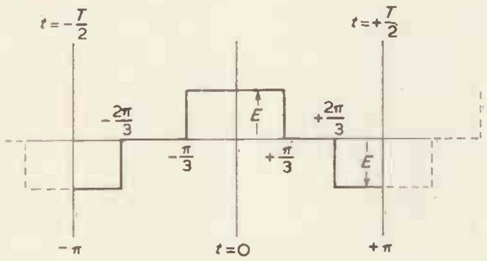


Fig. 4. Gating waveform $f(t)$

If we denote the coefficient of any cosine terms by A_n, then it can be shown that for any periodic function of time $f(t)$, then:

$$A_n = (1/\pi) \int_{-\pi}^{+\pi} f(t) \cos n(\omega t) \, d(\omega t)$$

where $\omega = 2\pi f = (2\pi/T)$

T being the periodic time.

Thus $\omega t = \pi$ when $t = (T/2)$ (see Fig. 4).

Since the function under consideration is discontinuous, we can consider the above integral as composed of several parts.

Thus:

$$A_n = (1/\pi) \int_{-\pi/3}^{+\pi/3} E \cos n(\omega t)\, d(\omega t) + (1/\pi) \int_{+2\pi/3}^{+\pi} (-E) \cos n(\omega t)\, d(\omega t)$$

$$+ (1/\pi) \int_{-\pi}^{-2\pi/3} (-E) \cos n(\omega t)\, d(\omega t)$$

$$= (E/n\pi)[\sin n(\omega t)]_{-\pi/3}^{+\pi/3} - (E/n\pi)[\sin n(\omega t)]_{+2\pi/3}^{+\pi} -$$

$$(E/n\pi)[\sin n(\omega t)]_{-\pi}^{-2\pi/3}$$

$$= (E/n\pi)[\sin(n\pi/3) - \sin(-n\pi/3) - \sin(n\pi) +$$

$$\sin(2n\pi/3) - \sin(-2n\pi/3) + \sin(-n\pi)]$$

But $\sin x = -\sin(-x)$

Thus:

$$A_n = (2E/n\pi)[\sin(n\pi/3) - \sin(n\pi) + \sin(2n\pi/3)]$$

$$= (2E/n\pi)[\sin(n\pi/3) + \sin(2n\pi/3)]$$

For increasing values of n, we find the following values of $(A_n\pi/2E)$:

n	1	2	3	4	5	6	7	8	9	10	11	12
$\frac{A_n\pi}{2E}$	$+\sqrt{3}$	0	0	0	$\frac{-\sqrt{3}}{5}$	0	$\frac{+\sqrt{3}}{7}$	0	0	0	$\frac{-\sqrt{3}}{11}$	0

Thus the function $f(t)$ can be expressed as:

$$f(t) = (2\sqrt{3}E/\pi)[\cos(\omega t) - (1/5)\cos(5\omega t) +$$

$$(1/7)\cos(7\omega t) - (1/11)\cos(11\omega t) + \ldots\ldots]$$

Let the signal waveform be represented by, say, $f'(t)$

$$\text{where } f'(t) = a_o + \sum_{n=1}^{n=\infty} [b_n \sin(n\omega t) + c_n \cos(n\omega t)]$$

Multiplying $f(t)$ and $f'(t)$ together, we see immediately that a large number of terms will become zero, due to the following relationships:

$$\int_0^{2\pi} \sin nx \ \sin mx \, dx = 0 \text{ for } m \neq n$$

$$\int_0^{2\pi} \cos nx \ \sin mx \, dx = 0 \text{ for } m \neq n$$

$$\int_0^{2\pi} \cos nx \ \cos mx \, dx = 0 \text{ for } m \neq n$$

Thus, the only terms that can exist in the product are those due to the 5th, 7th, 11th, etc. harmonics of the signal waveform, since they are the only ones that will give rise to finite integrals.

Furthermore, the product terms due to these harmonics will decrease in direct proportion to the order of the harmonic concerned.

Finally, it is worth noting that the terms which disappear, due to the following relationships, are those due to quadrature signals, i.e. there will be no quadrature response.

$$\int_0^{2\pi} \sin nx \cdot \cos mx \, dx = 0 \text{ for } m = n$$

As has been noted, the above analysis is substantiated in the practical circuit described.

Acknowledgments

The author wishes to express his thanks to the directors of Solartron Research & Development Ltd. for permission to publish these notes.

Opening Ceremony by Television Link

What is claimed to be Britain's longest closed television link was set up recently in connexion with the opening of the British Oxygen Company's "tonnage" oxygen plant at Margam in South Wales.

This oxygen plant, which is the first of its kind in Great Britain, has an output of 100 tons of oxygen per day and supplies oxygen in bulk to the nearby Abbey Works of the Steel Company of Wales.

It will be followed very shortly by a companion plant with a capacity of some 200 tons/day (one ton of oxygen is equivalent to approximately 26 000ft^3 at normal temperature and pressure).

The opening ceremony was performed by the Home Secretary and Minister for Welsh Affairs, Major the Rt. Hon. Gwilym Lloyd-George, M.P., from the Board Room of the British Oxygen Company Ltd. at Bridgewater House, London, after which a "tour" of the oxygen plant and steelworks was carried out by means of a closed television link. The television link involved a distance of some 200 miles.

At Margam, three Marconi standard image orthicon cameras were used, their signals being relayed through normal control equipment to an s.h.f. radio link at Pyle, from whence they were fed into a G.P.O. coaxial cable routed to Wenvoe and then by five video repeater links to the Museum Telephone Exchange in London for connexion to Bridgewater House.

A Circuit for Analogue Formation of xy/Z

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

A quarter squares multiplier, using a triangle carrier waveform in the squaring circuits is extended to give division simultaneously with multiplication. This is achieved by controlling the slope of the triangle carrier waveform so as to be proportional to the divisor Z.

THE scheme described herein embodies the difference of squares method for formation of the product, using triangle waves for the generation of squares[1]. Division of the product is effected by control of the slope[2] of the triangle wave input to the squarers, maintaining a constant height of triangle and constant repetition rate.

Basic Circuit (Fig. 1) (Waveforms Fig. 2)

Trigger pulses of constant repetition rate, generated by the oscillator trigger the triangle wave generator. The slope of the triangle is proportional to the voltage Z, which is the input to a Miller integrator triangle wave generator. It is essential that the flyback time of the triangle be minimized to ensure accurate division. Formation of the squares by clipping off from the input triangle heights $|x + y|$ and $|x - y|$ gives outputs proportional to $(x + y)^2/Z$ and $(x - y)^2/Z$ respectively. The difference of these yields an output proportional to xy/Z.

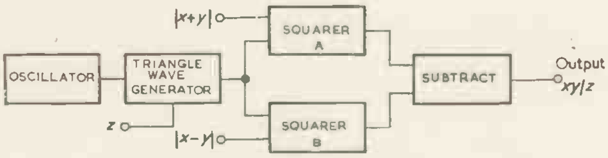


Fig. 1. The multiplier/divider

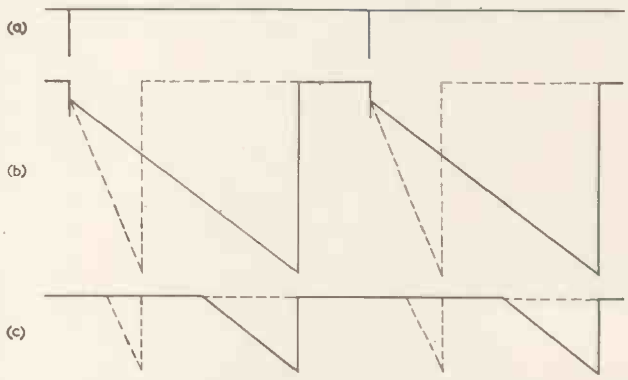


Fig. 2. Waveforms
(a) Trigger pulse from oscillator, (b) triangle wave outputs for two values of Z, (c) corresponding squarer outputs

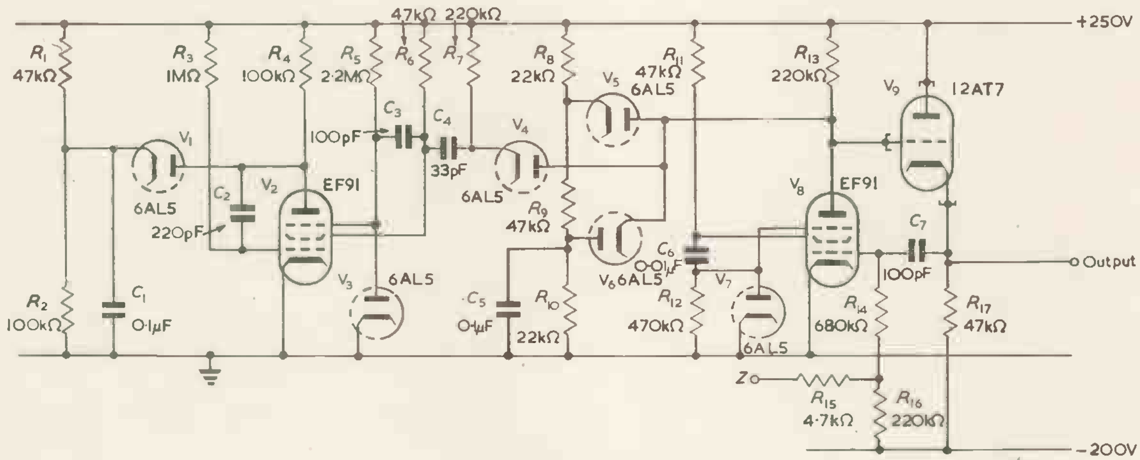


Fig. 3. Variable slope triangle wave generator

Variable Slope Triangle Wave Generator (Fig. 3)

An a.c. coupled Miller transitron (V_2) generates at its screen grid a square wave having regenerative negative-going steps, and a repetition rate of 5kc/s.

V_1 stabilizes the positive level of the timing waveform at V_2 anode to give a stable frequency of oscillation for the circuit. Differentiation of the square wave at V_2 screen, by C_4R_7, provides a negative pulse which triggers the variable slope triangle wave generator (V_8) via V_4. The run-down rate of this Miller integrator is proportional to the input Z.

Rapid recharging of the integrating capacitor C_7 is obtained by recharging it from the output cathode-follower V_9. To obtain the highest ratio of charging to discharging time for C_7, its value must be a minimum compatible with linear triangle generation. V_8 is effectively a d.c. coupled Miller transitron, since the time-constant of the screen grid to suppressor coupling (C_6R_{12}) is very long compared with the repetition rate. Diodes V_5 and V_6 provide stabilization of the triangle wave d.c. levels.

Squarers

Fig. 4 shows the circuit of squarer A, in which the inputs $+(x + y)$ and $-(x + y)$ are added to the triangle wave on adding chains $R_{18}R_{19}R_{20}$ and $R_{22}R_{23}R_{24}$ respectively. VR_1 and VR_2 provide adjustment of the input levels to V_{11} and V_{10} so that when both x and y are zero the outputs from V_{12} and V_{13} are both just zero. The effect of stray capacitance at the output point is minimized by returning R_{30} to h.t.+, using V_{14} to stabilize the d.c. output level. This ensures that the fast flyback is maintained at the output. Squarer B is identical to squarer A, except that the inputs to the adding chains are $+(x - y)$ and $-(x - y)$.

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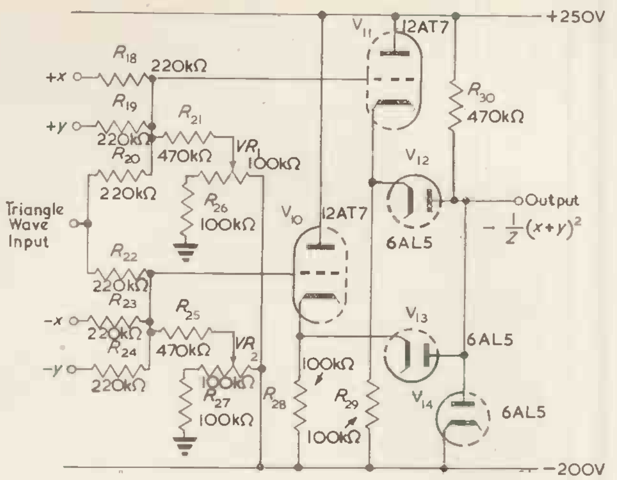

Fig. 4. Squarer

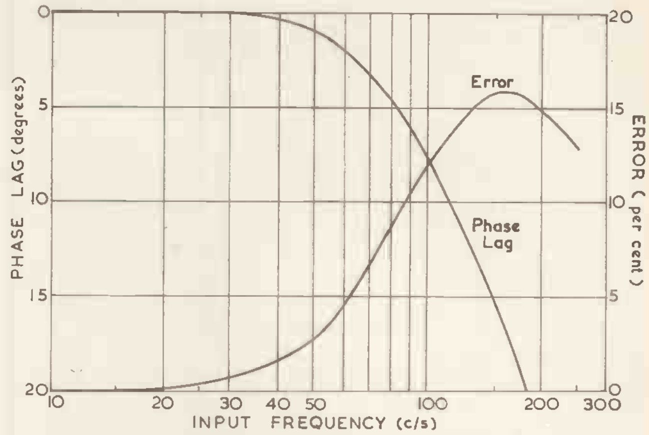

Fig. 6. Multiplier phase shift and error for sinusoidal x or y input
Z and x (or y) constant. Oscillator frequency 5kc/s

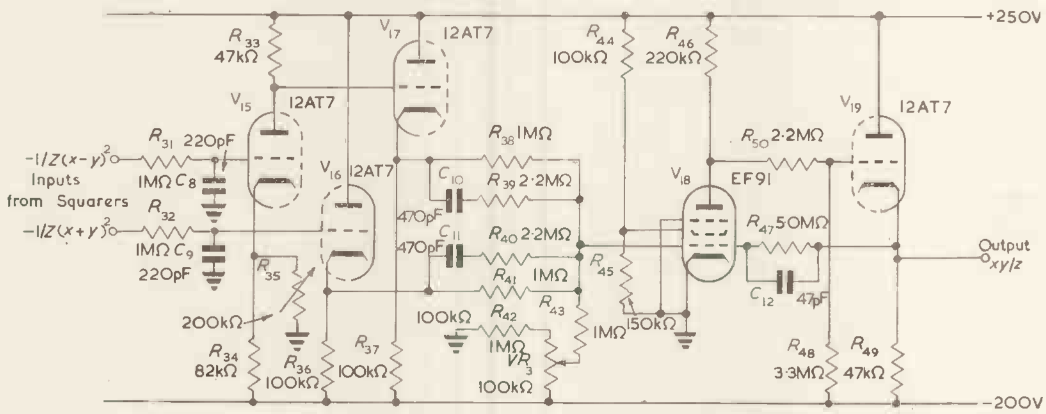

Fig. 5. Subtraction of squarer outputs

Subtraction of Squarer Outputs (Fig. 5)

Accurate reversal by V_{15} of the input $-(1/Z)(x - y)^2$ is adjusted by R_{35}. V_{16} and V_{17} cathode-follow the signals $-(1/Z)(x + y)^2$ and $+(1/Z)(x - y)^2$, which are added and reversed by the anode-follower V_{18}, to give an output proportional to xy/Z. Smoothing out of the carrier frequency is by $C_8R_{31}(=C_9R_{32})$ and $C_{12}R_{47}$. A limited amount of low frequency phase advance is provided by $C_{10}R_{39}(=C_{11}R_{40})$ to compensate for the signal phase lag due to the circuit which provides smoothing of the carrier frequency. For correct compensation at low frequencies:

$$C_{10}R_{38} = C_{11}R_{41} = C_8R_{31} + C_{12}R_{47} = C_9R_{32} + C_{12}R_{47}$$

Fig. 6 shows experimental curves of phase shift and change in output amplitude with frequency, for sinusoidal x or y input, when y (or x) and Z are constant.

Conclusion

The circuit described is accurate to ± 2 per cent of maximum output for four-quadrant values of x and y between plus and minus 50V, and for all values of Z between about 40 and 200V (Z_{min} to $5Z_{min}$). Greater accuracy could be obtained by using a lower oscillator frequency. The multiplier must be supplied from stabilized positive and negative lines, when the settings of squarer zeros and the subtractor zero (VR_3) are adequately stable. Output ripple is dependent upon the value of Z, and at $Z = Z_{min}$ the r.m.s. ripple is 2 per cent, increasing to about 8 per cent of output voltage at $Z = 5Z_{min}$. Inputs $\pm x$ and $\pm y$ are assumed to be available, but can be generated from x and y by anode-followers.

REFERENCES

1. NORSWORTHY, K. H. A Simple Electronic Multiplier. *Electronic Engng. 26*, 72 (1954).
2. DOUCE, J. L. A Simple Analogue Divider. *Electronic Engng. 26*, 155 (1954).

A New Van de Graaff Installation

A 2MV Van de Graaff electrostatic particle accelerator has recently been installed by Standard Telecommunication Laboratories Ltd. for industrial research. The accelerator has a 4ft belt 10in wide which travels at 4 000 rev/min, the initial charge on the belt being 35kV. The output of the generator can be either positive or negative and both the particle source and the target can be changed to give various types of output. The electron beam power is 100W and electrons can be accelerated to 98 per cent the velocity of light.

The research programme in the immediate future will be mainly into the effects of irradiating various materials and in particular insulators, as for example polythene, the melting point of which can be raised considerably by irradiating it in an electron beam. The various effects of irradiating semiconductor materials will also be investigated.

The Van de Graaff generator and the control equipment have been supplied by the High Voltage Engineering Corporation of America.

Wide-Range Voltage Stabilizers

By F. A. Benson*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E.,
and L. J. Bental*, B.Sc., Dipl. Ing., M.Eng.

A brief review is first given of methods which have been developed to provide a stabilized supply with a wide voltage range and whose output can be adjusted down to a very low voltage. Details of one form of such stabilizer, which has been constructed by the authors, are then given with some information about its performance and limitations. The supply, which is best suited for laboratory work, has an output voltage adjustable within the range from zero to 300V. The maximum load current is 180mA for output voltages under 50V, but at higher output voltages it is 200mA.

ALTHOUGH the literature on thermionic-valve stabilizers is very extensive, the majority of arrangements have the disadvantage that the output voltage cannot be reduced below a certain minimum value of about 100 to 200V. To obtain voltage control below this minimum, considerable elaboration of the usual arrangement is necessary and this may reduce the degree of stabilization. Methods of obtaining a stabilized supply with a wide voltage range whose output can be adjusted to a very low voltage are discussed here. The authors have constructed one form of such a stabilizer in order to obtain detailed information about its performance and limitations. A description of this equipment is included.

Most regulated power supplies employ a stabilizing unit based on the series-parallel valve circuit[1] (Fig. 1). In this circuit the control valve V_1 is in series with the load and the total load current flows through it. V_2 is a d.c. voltage amplifier which amplifies variations in the output voltage V_o. Glow-discharge tube V_3 provides the reference voltage; it maintains the cathode of V_2 at a constant potential with respect to the negative terminal. If there is a deviation ΔV_o in the output voltage, part of it is applied by means of potentiometer R_1 to the grid of V_2 where it is amplified and then fed directly to the grid of control valve V_1. In this way the change of V_o is counteracted. The output voltage can be varied by adjusting R_1.

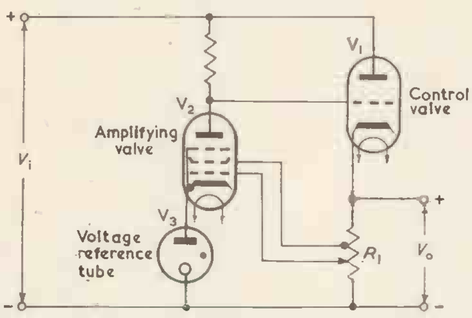

Fig. 1. Series-parallel valve stabilizing circuit

It can be shown that in the series-parallel valve circuit just described both the stabilization ratio and the internal resistance depend on the amplification factors of the two valves V_1 and V_2. In practical units a power pentode connected as a triode is frequently used as the control valve. In this case the amplification factor of the control valve does not usually exceed 20 and the contribution of V_1 towards the regulation is only small. For good regulation a supply for feeding the screen grid of the control valve should be provided, thus taking full advantage of the high amplification factor of the pentode. The d.c. amplifier stage commonly employs a pentode in view of the high gain obtainable. To counteract input voltage variations the screen grid supply for V_2 is sometimes taken from the input side of the unit instead of from the output (see Fig. 2). Further improvement may be achieved by stabilizing the screen grid voltage of the control valve and the anode supply for the amplifier valve by means of glow-discharge tubes.

The output voltage of the circuit of Fig. 1 cannot be reduced to zero. In fact V_o must always be greater than the sum of the voltage drop across the amplifier valve (usually more than 75V), the grid bias of the control valve (approximately 20V) and the reference voltage. If, for example, an 85A2 tube is used to provide a reference voltage of about 85V, the minimum output voltage obtainable is near 180V. This minimum can, of course, be reduced by connecting a potential divider across the tube V_3, thus reducing the reference voltage to about 20V or, if convenient, by using a battery to provide the reference voltage. Special sub-miniature valves with a working anode-cathode voltage of only about 25V can be employed as voltage amplifiers. These modifications enable adjustment of the output voltage down to about 65V. In practice, however, it is often desirable to have a stabilized power supply, the output voltage of which can be varied continuously over the range from a few hundred volts down to zero. Circuits meeting this requirement have been developed by Working[2], Garratt[3], White and Mansford[4], Scroggie[5], Green[6], Houle[7,8], May and Skalnik[9] and Admiraal[10].

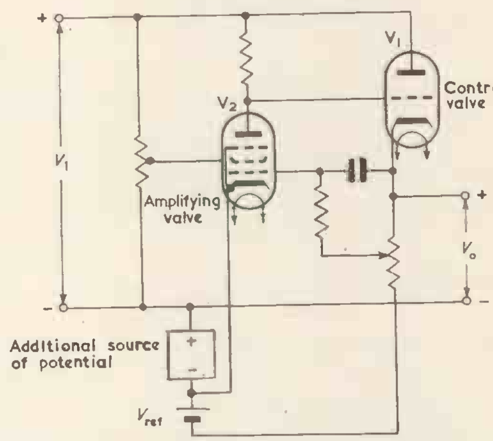

Fig. 2. Modified stabilizing circuit to obtain low output voltage

Working[2] employed two stabilized supplies connected in parallel, one supply having a fixed output voltage of approximately 200V, the other having an output voltage variable over the range 200 to 400V. By connecting the

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negative terminals of the two units together, any voltage from zero to 200V could be obtained between the two positive terminals.

Garratt[3] developed a simpler circuit by using an additional voltage supply which was stabilized with a VR150 glow-discharge tube. Low output voltage could thus be obtained. This circuit is very similar to one developed previously by White and Mansford[4] which is shown in Fig. 2. The fundamental principle is to use an auxiliary source of potential to keep the cathode of the amplifying valve at a negative potential with respect to the negative terminal. If this potential is so large that the anode of the amplifying valve can be made sufficiently negative to completely cut off the control valve, then it is possible to reduce the output voltage to zero. The basic circuit of White and Mansford has frequently been employed, the negative potential for the d.c. amplifier usually being obtained from separate transformer windings through a rectifier unit and stabilized with glow-discharge tubes. The circuit has the advantage of being easily adaptable to various requirements. If a large current is desired, for example, several control valves can be used in parallel. The voltage range can be widened by employing a transformer which gives the required supply voltage. It should be remembered, however, that the control valve is in series with the load and there is a large voltage drop across the valve, therefore, at low output voltages. This drop should not exceed the specified maximum anode voltage. In order not to overrate the valve it is customary to provide the h.t. transformer with tappings. When low output voltages are selected by means of the range switch, the anode supply voltage for the control valve is reduced accordingly.

Full details of a regulated supply with a continuously variable output voltage from zero to 500V at currents up to 100mA have been given by Scroggie[5]. A circuit diagram of his unit is shown in Fig. 3. Two EL37 power pentodes are employed as control valves and the d.c. amplifier is an EF42 high slope pentode. A resistor is incorporated in the screen grid circuit of V_2 in addition to the glow-discharge tube V_5 to counteract input-voltage variations. Metal rectifiers are used in the auxiliary negative supply circuit, the latter being stabilized with 85A1 glow-discharge tubes. This negative supply may be used to provide additional negative output voltages (terminals A, B and C). No appreciable current should be drawn from the negative output to avoid overloading the glow-discharge tubes. Scroggie emphasizes the necessity of smoothing the auxiliary supply very thoroughly to avoid fluctuations in the reference provided by V_3. A supply similar to Scroggie's has been described by Green[6]. In this case four 807 valves in parallel provide the control element and a single 6AC7 valve, operated below earth, serves as a d.c. amplifier. The output voltage of Green's supply is adjustable within the range 3·5 to 500V at currents up to 300mA. The unit has an internal impedance of 8 to 12Ω.

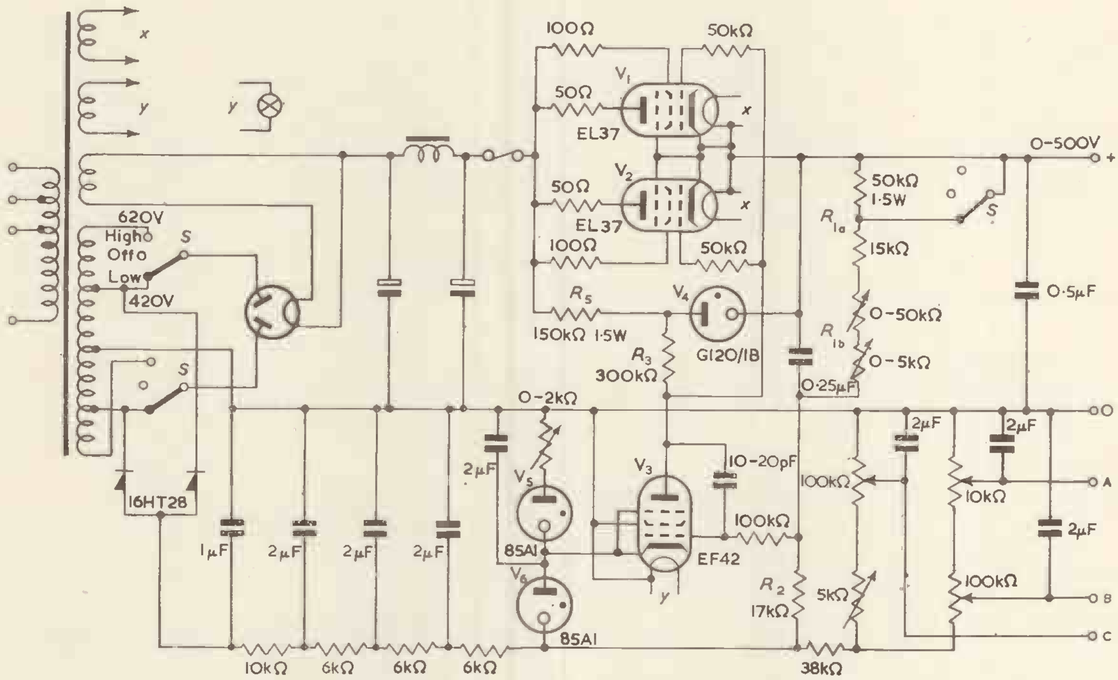


Fig. 3. 0 to 500V regulated power supply

Houle[7,8] criticizes stabilizers which are based on the above circuit because of the many separate heater transformer windings required and the excessive power dissipated in the auxiliary circuits. He has described an alternative circuit for a supply with a stabilized output voltage variable from zero to 300V (Fig. 4). The usual d.c. amplifier has been exchanged by one based on a carrier-voltage principle. The source of carrier voltage is an oscillator with a frequency of 150kc/s and the modulator comprises a 1N38 crystal diode. The d.c. control voltage (i.e. the difference between part of the output voltage and the reference voltage) varies the conduction of this diode. Amplification of the controlled carrier voltage is effected by a conventional tuned amplifier and the d.c. signal is recovered, after amplification, by a 6H6 diode and fed to the grid of the control valve. The internal impedance of the equipment is fairly high, being about 40Ω. Although the various circuit elements in this unit are simple and small, it seems that the principle has not generally been adopted.

May and Skalnik[9] have developed a stabilizer, the output voltage of which is not only adjustable to zero, but which maintains regulation while carrying reverse currents. This is of importance for bias supplies for class-C operated amplifiers. A beam power valve is connected in parallel with the load and serves as an electronic bleeder which draws an approximately constant current regardless of the output voltage. Thus the supply can draw a negative current equal to the bleeder current (about 25mA) before losing its stabilizing ability.

All the above arrangements possess the disadvantages that the amplifier valve is controlled by only part of the

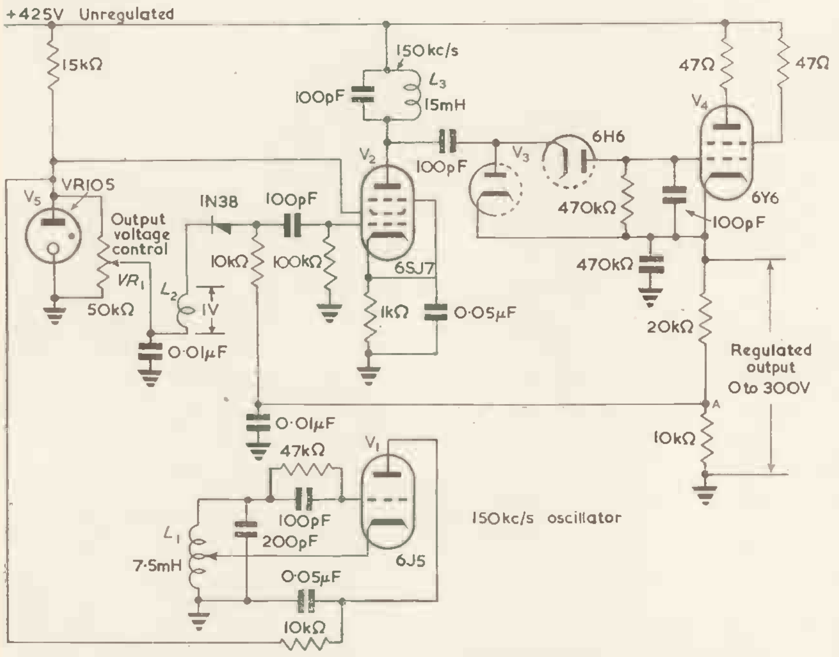


Fig. 4. 0 to 300V stabilizing circuit employing a carrier-type control amplifier

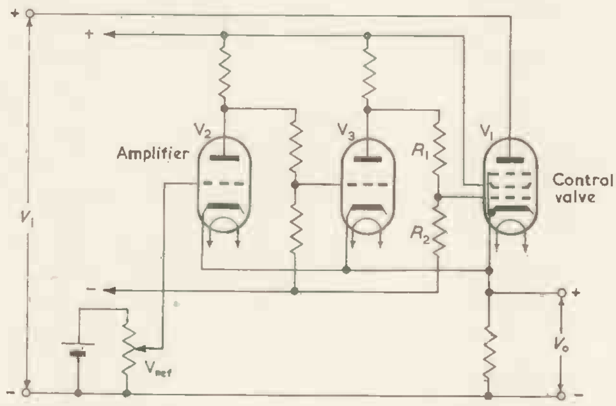


Fig. 5. Stabilizing circuit in which the variations in output voltage are applied in their entirety to the amplifier

output voltage variation and that this part depends on the desired output voltage and varies usually between one-half and one-sixth. Admiraal[10] has designed a circuit in which the total variation of output voltage is always applied to the amplifier. This circuit, which is shown in Fig. 5 is, therefore, useful when a very high degree of stabilization is required. The control valve V_1 is a pentode. Two stages of amplification are necessary so that the voltage variations are applied to the grid of the control valve with the correct phase. (If only one voltage amplifying valve is used, the reference voltage source must be connected to the cathode of it. The cathode current of the amplifying valve must then flow through the reference source and because of variations in this current the stabilization might be rather poor.) The anode voltage for the amplifier valves is derived from auxiliary transformer windings required for feeding the screen grid of the control valve. The negative terminal of this auxiliary supply is negative with respect to the cathode of the control valve and the required potential for the control grid of this valve can easily be obtained by adjusting the potential divider R_1 and R_2. The auxiliary supply should be pre-stabilized so that the screen grid of the control valve and the anode supply for the amplifier valves are kept at constant potentials independent of the input voltage. In this circuit the output voltage is almost equal to the reference voltage and it can be adjusted by varying the latter. (It has been shown elsewhere[1] that the larger the reference voltage the greater the stabilization ratio.)

The authors have constructed a stabilizer very similar to that of Admiraal. Details of this unit will now be given with some information about its performance and limitations. Several changes were introduced into Admiraal's circuit to conform with various requirements, but the fundamental principles have not been altered. Thus:

(a) The unit is based on the circuit shown in Fig. 5.

(b) The control valves are power-pentodes with a separate pre-stabilized supply feeding their screen grids.

(c) The amplifier employs sub-miniature pentodes.

The heater current of the sub-miniature valves is very small (15mA) and can be derived, therefore, from the stabilized auxiliary supply without overloading the glow-discharge tubes. This results in a considerable improvement in the performance of the equipment.

The supply has an output voltage adjustable within the range from zero to 300V. At output voltages under 50V the load current should not exceed 180mA to avoid exceeding the maximum anode dissipation of the control valves. At higher output voltages the maximum load current is 200mA.

The complete circuit diagram of the unit is shown in Fig. 6. Four KT66 beam-power tetrodes are employed as control valves with resistors in the anode and grid leads to eliminate oscillations. Several control valves are required since it is the total anode dissipation that limits the output current of the unit. The screen grid voltage is stabilized at 210V with two QS1206 glow-discharge tubes. A 200V, 15W incandescent lamp is used as a series resistor to avoid large current variations through the glow-discharge tubes when the input voltage varies. Stabilization of the screen grid voltage limits the current through each control valve and also limits the number of valves which may be connected in parallel to obtain large output current. At maximum output voltage and current (300V and 200mA respectively) and under unfavourable conditions (input voltage below nominal value) the four KT66 tetrodes draw a considerable screen grid current from the QS1206 stabilizer tubes. If the current is increased further, or if another control valve is connected, the current through the QS1206 tubes may drop below the extinction value. For similar reasons the input voltage should not drop more than 12 per cent below the nominal value.

The amplifier consists of two direct-coupled stages employing DL66 sub-miniature pentodes. As pointed out

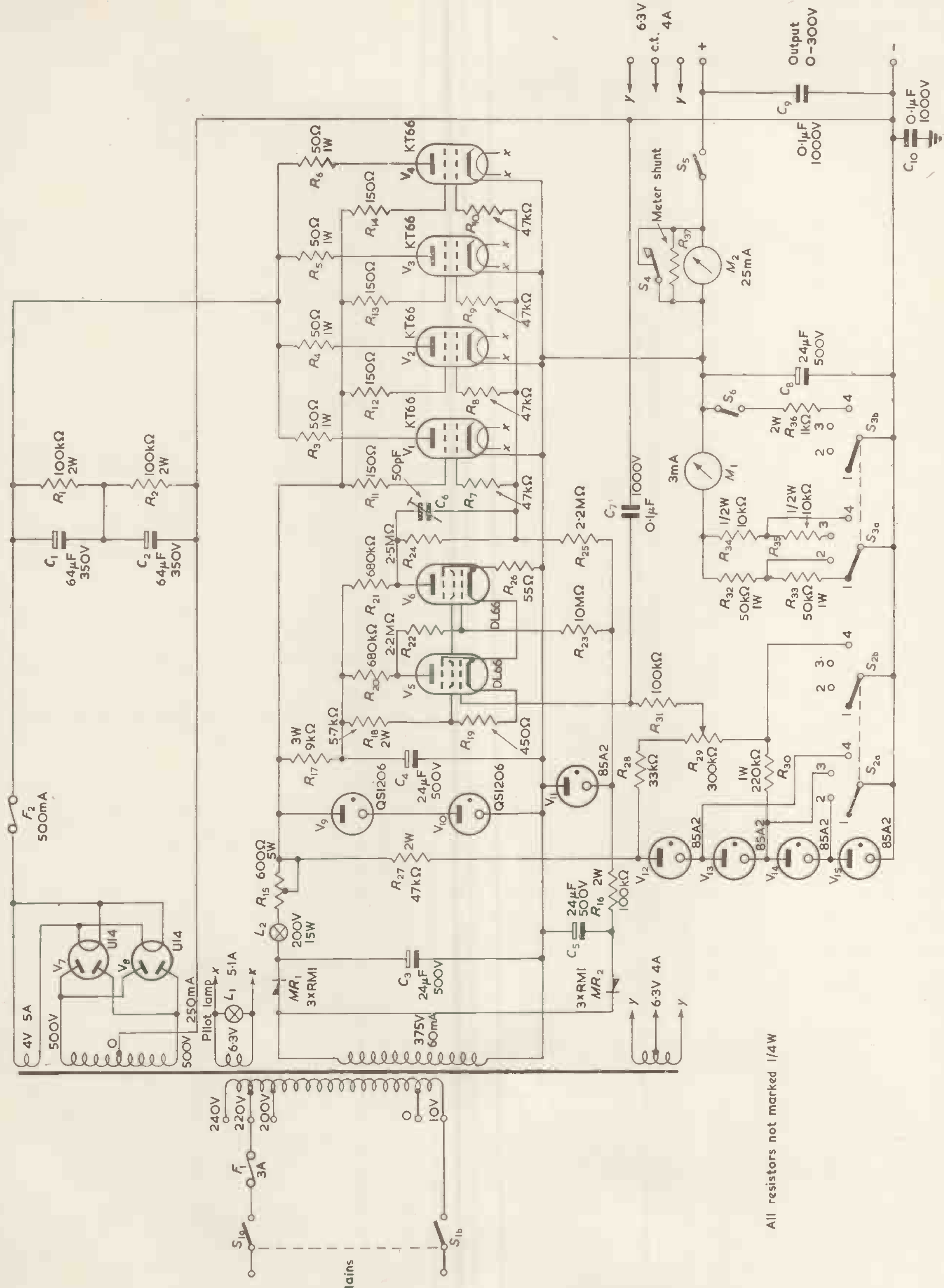


Fig. 6. Wide voltage-range stabilized power supply

by Admiraal, the anode supply and the anode resistors for the DL66 valves should be increased to 75V and 680kΩ, respectively, to reduce the effect of the load current on the relative anode current variations. Both anode and heater supplies for the amplifier valves are derived from the stabilized auxiliary supply. The glow-discharge tube V_{11} provides a constant negative potential with respect to the cathodes of the KT66 valves. The grid bias for the control valves can be set at its correct value by means of the potential divider R_{24} and R_{25}. The requirements that

(a) the output voltage V_o may be reduced to zero,

(b) the regulation at maximum output voltage (V_o = 300) is effective,

limit the values of the resistors to close tolerances.

Parasitic oscillations were encountered as a result of

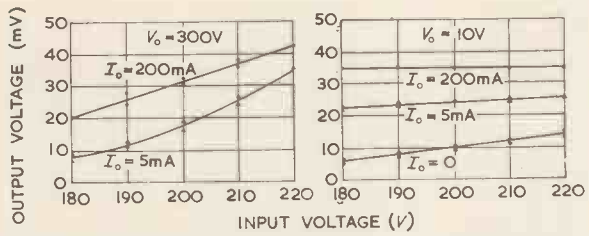


Fig. 7. Output voltage variations as function of input voltage for various load currents

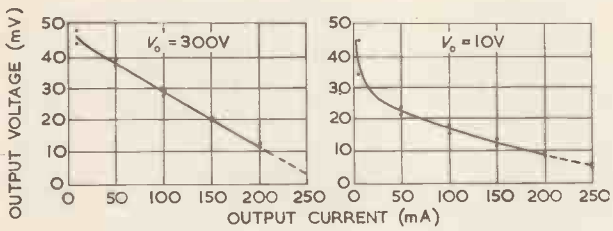


Fig. 8. Output voltage variations as function of load current

the very high gain in the feedback loop. To eliminate these, a capacitor, C_6 was connected across the resistor R_{24}. The value of C_6 was found to be rather critical (about 20pF) and a pre-set capacitor was therefore used. Further, the d.c. amplifier and all leads to the grids of the control valves should be screened.

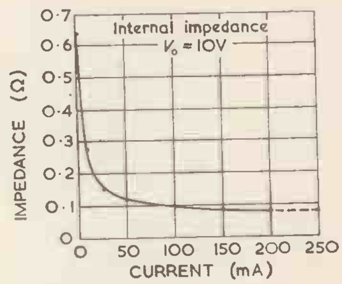
The reference voltage is taken from four 85A2 tubes connected in series and fed from the auxiliary supply. The supply voltage is pre-stabilized and the series resistor is of large value so that the reference-voltage tubes are operated at a very nearly constant current. Care should also be taken to avoid any running-voltage variations in these tubes due to temperature changes or vibration. By means of the range switch S_2, each 85A2 tube can be successively short-circuited. This reduces the reference voltage (and also the output voltage) in steps of about 85V.

The meters M_1 and M_2 read the output voltage and current, respectively. (Switch S_3 enables full-scale deflexions of M_1 at 30, 60, 150 or 300V to be obtained). The shunt R_{37} for the milliammeter M_2 adds considerably to the very small internal resistance of the equipment. A switch S_4 is therefore provided, which short-circuits this meter when not in use.

The performance of the stabilizer was determined for input voltage fluctuations and for load-current variations. Fig. 7 shows the output voltage as a function of the input voltage at output voltages of 10V and 300V. The curve for $V_o = 10$V and $I_o = 0$ shows that for very low output voltages the bleeder current is insufficient. The additional bleeder R_{36} should, therefore, be switched in when a current of less than 10mA is drawn from the supply at low output voltages. The stabilization ratio $(\Delta V_i/V_i)/(\Delta V_o/V_o)$ is approximately 2000. The effect of load current on the output voltage is also extremely small, as shown by Fig. 8. The rapidly increasing part of the $V_o = 10$V curve for currents under 10mA can be eliminated by the above mentioned bleeder. It is then found that the internal impedance of the equipment (i.e. $-\Delta V_o/\Delta I_o$) is about 0·1 to 0·2Ω for all voltage ranges except at very low currents. Fig. 9 shows the internal impedance as a function of load-current for an output voltage of 10V.

The auxiliary circuits of the equipment tend to become rather complicated and the power consumption is fairly high so that the supply is best suited for laboratory work where a very high degree of stability is required. At no load and an output voltage of 300V the input power is about 110W and at full load it is approximately 260W. To dissipate a relatively large amount of power in the auxiliary

Fig. 9. Internal impedance of equipment at an output voltage of 10V. (Additional bleeder switch on)



circuits is considered justified in order to obtain the excellent performance of the unit. Even though the stability of this unit is extremely good, it is inferior to that obtained by Admiraal[10]. This circuit is difficult to adjust, however, and if one component is changed the whole circuit needs some modification. Further, normal laboratory components have been used throughout and not special high-stability ones.

Acknowledgments

The work recorded here has been carried out in the Department of Electrical Engineering at the University of Sheffield. The authors wish to thank Mr. O. I. Butler, M.Sc., M.I.E.E., A.M.I.Mech.E. for facilities afforded in the laboratories of this Department.

REFERENCES

1. Benson, F. A. Voltage Stabilizers. Electronic Engineering Monograph (1950)
2. Working, E. B. An Electronic Voltage Regulator. *Indusr. Engng. Chem. 30* 435 (1938).
3. Garratt, P. M. Wide-Range Voltage-Regulated Power Supply. G.E. (America) Data Folder 46243, *C.R.B.* Ref. 43/1701 (1943).
4. White, E. L. C., Mansford, H. L. Improvements in Voltage Stabilizing Devices for Electric Supply Systems. British Patent 515,435 (1939) and U.S. Patent, 2, 268, 790 (1942).
5. Scroggie, M. G. Stabilized Power Supplies. *Wireless World 54*, 453 (1948).
6. Green, C. E. Wide-Range Voltage Regulator. *Electronics 23*, 192 (March, 1950).
7. Houle, J. Wide-Range Voltage Regulator. *Electronics 24*, 202 (August, 1951).
8. Houle, J. Carrier-Type Regulated Power Supply. *Radio-Electronic Engng. 16*, 14 (November, 1951).
9. May, J. C., Skalnik, J. G. Regulated Bias Supply. *Electronics 24*, 136 (December, 1951).
10. Admiraal, D. J. H. A Stabilized D.C. Supply Unit for Output Voltages of 0 to 300V. *Philips Electronic Application Bulletin 14*, 61 (May, 1953).

A Method of Tuning Resonant Cavistie

By W. M. Haywood*, B.Sc.

A method of tuning resonant microwave cavities over a bandwidth of approximately one per cent is described. The inductive tuning rod is enclosed in thin wall glass tubing, all movement therefore taking place outside the vacuum envelope. The effects of the glass tube on cavity properties are described and the performance of the system under conditions existing during valve processing examined. An assessment is made of the merits of the tuning system.

IN the construction of klystrons it is usually necessary to incorporate in the valve a mechanism allowing the tuning of the resonant cavities. This is necessary, to allow for manufacturing variations in the construction of cavities, resulting in nominally similar cavities having different resonant frequencies; and also to allow adjustment of bandwidth and examination of valve performance as one, or more, of the cavities are detuned.

A normal method of cavity-tuning employed is an inductive one, utilizing a rod penetrating the cavity radially from the perimeter, in the region of high magnetic field. Until now this method has necessitated the use of vacuum-tight flexible bellows in a form of construction such as that shown in Fig. 1.

The use of these bellows involves numerous disadvantages. Firstly, they may limit the baking temperature of the valve. Thus, when brass bellows have been used the baking temperature has been limited to 350°C by dezincification of the brass, a temperature which is too low to ensure adequate outgassing of valve components. Secondly, bellows are prone to leaks which are difficult to seal effectively when they occur within a convolution. Thirdly, when brazing the bellows in position, the temperature is such that near the brazes the bellows may lose their flexibility necessitating the use of a greater length of bellows than would otherwise be the case. Also, the use of a bellows assembly involves the use of a considerable number of small components, and the mechanism can become rather large and unwieldy.

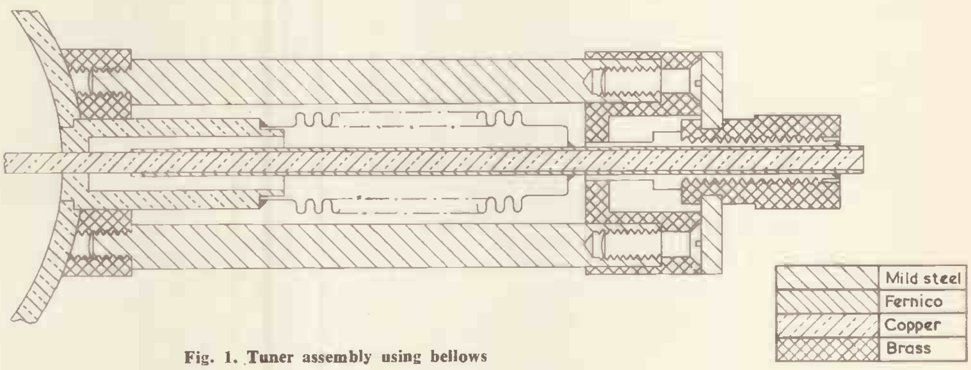


Fig. 1. Tuner assembly using bellows

In the work to be described the use of bellows has been avoided entirely. This has been achieved by mounting the tuning rod outside the vacuum envelope, a thin glass tube penetrating the cavity wall in which the copper tuning rod can be moved, forming the vacuum envelope. This is shown in Fig. 2. While this method avoids the practical limitations of bellows detailed above, it has several disadvantages which affect mainly the electrical performance of the valve in which it is used. The presence of glass within the cavity results in a reduction in cavity shunt impedance, R_{sh}, which, in a klystron operating at a d.c. level where cavity losses are important, will result in diminished valve performance.

Measurements have been made on the changes in R_{sh} and other cavity properties arising from the use of this "glass dome" tuner in an X-band cavity, and, by processing a cavity in which such a tuner was employed an assessment made of the likely performance of the mechanism when mounted in a valve.

* *Mullard Ltd.*

Initial Measurements

To investigate the effect of introducing a glass tube in to the cavity on the cavity properties, measurements were made of the change of shunt impedance of a cavity as a "Kodial" glass tube of wall thickness 0·5mm, and closed at one end, was introduced from the side of the cavity. The cavity terminated a waveguide line, the Q being measured by Lawson's method from measurements of the v.s.w.r. and phase on the line in the region of cavity resonance[1]. From this, R_{sh} was found by loading the cavity nose with a dielectric rod and measuring the change in resonant frequency[2].

The results of these measurements, shown in Fig. 3, indicate that the reduction in R_{sh} was sufficiently small (10 to 20 per cent for adequate tuning range) to warrant further investigation of the method.

Physical Construction

A diagram of the construction of the tuning mechanism is shown in Fig. 2.

The glass tubing employed was precision-bore tubing of wall thickness 0·5mm closed at one end and of internal diameter 3·1mm, the tuner rod being 3mm in diameter. The required penetration into the cavity being known from considerations of the desired tuning range of the cavity, the requisite length of glass tube was sealed by means of an outside seal on to a Fernico cylinder. The concentricity of the glass tube and Fernico cylinder is obviously essential, the copper tuning rod passing down both, and this was achieved using a carbon mandrel through both, while the seal was made. A ledge machined on the inside of the main Fernico piece which is brazed into the cavity block, located the glass rod in the cavity, the position of the ledge being finally determined by machining after the glass seal had been made so that the penetration of the glass rod was accurately determined. The glass tube and Fernico assembly was then brazed to the main Fernico piece around the outer rim, as shown in Fig. 2, this forming the vacuum seal.

A standard micrometer thread, 40 t.p.i., together with a graduated scale, on the outside of the main Fernico piece held the tuner rod in position, allowing the movement of the copper tuning rod in and out of the cavity.

Measurements

Two mechanisms of the form described above have been made and the measurements carried out on each are described below.

Tuner 1

This was constructed to examine the electrical properties of a cavity with a "glass dome" tuner without any intention of testing the vacuum properties of the dome and associated seal and braze. The wall thickness of the glass dome was 0·5mm and set to permit a tuning rod penetration of 3mm into the cavity. No difficulty was experienced with the sealing or brazing.

Three sets of measurements were carried out on the effect of the glass dome on:

(a) Tuning range of the copper rod.

(b) Cavity Q.

(c) Cavity shunt impedance R_{sh}.

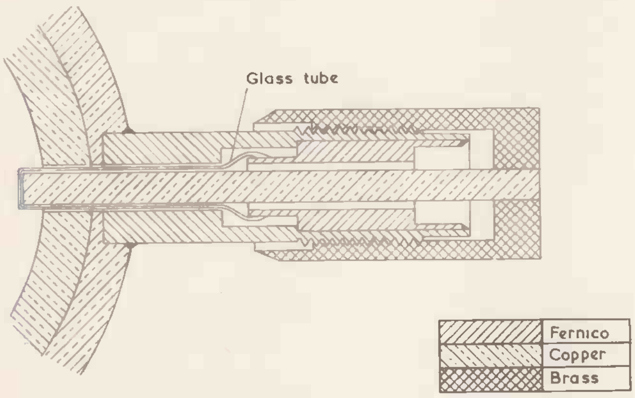


Fig. 2. Tuner assembly with tuning rod outside envelope
(Scale 1/16in = 1mm)

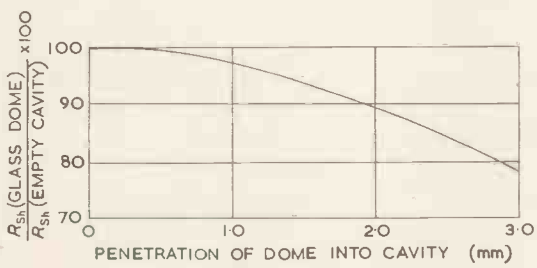


Fig. 3. Variation of R_{sh} with penetration of glass dome

The results of measurements with and without the glass dome present are shown in Table 1. The methods used for measurement of R_{sh} and Q are described elsewhere[2,3].

The results indicate that the tuning range is not appreciably affected by the presence of the glass dome, the effect of the glass dome being to increase the resonant frequency of the cavity between 10 to 20Mc/s depending upon the tuner position. This is shown in the tuning curves Fig. 4.

There is a reduction in R_{sh} due to the presence of the glass by approximately 20 per cent compared with the empty cavity. When the tuner rod is in position however the glass rod produces a reduction of only 5 per cent in R_{sh}. In fact R_{sh} is increased by the insertion of the tuner rod when the glass dome is present.

This apparent change may be a result of the limits of experimental accuracy achieved; the probable error in the Q measurements is within ± 5 per cent and the accuracy of measurement of R_{sh} probably ± 10 per cent. However, as a similar effect is observed with tuner 2 described later, it is thought that the increase in R_{sh} caused by the insertion

TABLE 1

Experimental results for tuner 1

CAVITY CONDITION	Q	R_{sh}
No tuner or dome	1900	30 000
Glass dome in position ..	1550	24 500
Tuner only (fully in)	1700	26 500
Glass dome in position and tuner fully in	1350	25 500
Tuning Range of Copper Tuner; Penetration 3mm		
Without glass dome		140Mc/s
With glass dome		136Mc/s

of the tuner rod when the glass dome is present is due to a distortion of the E field by the tuner rod. This results in some of the E field passing transversely through the glass rod with resultant reduction in glass losses.

Tuner 2

In order to test the vacuum properties of the glass dome a tube was constructed which could be evacuated, allowing the tube to be baked as during the processing of a valve.

The measurements made were similar to those made with

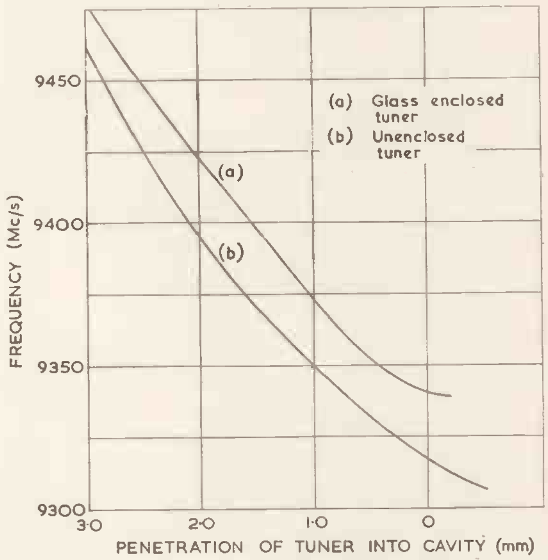


Fig. 4. Tuner 1 tuning range

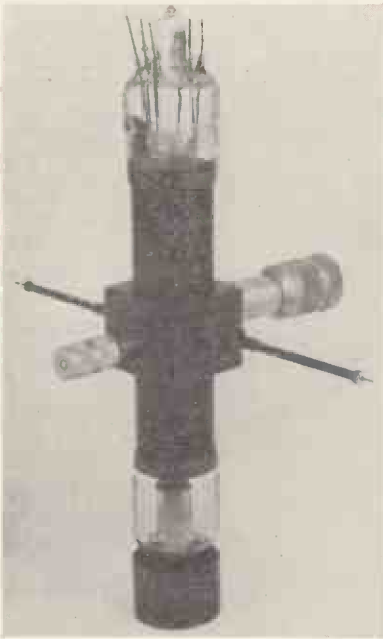
the first tuner, probes being sealed into the cavity in such a position to allow the unloaded Q of the cavity to be measured. R_{sh} was also measured as previously. After completion of measurement the tube was evacuated and a pressure of 1×10^{-6} mm Hg obtained. The tube was then baked, the rate of temperature rise being about 60° per hour and maintained at a temperature of about 360° C for 12 hours. Subsequently, the tube was taken through a second baking cycle, a temperature of 450° C being maintained for two days, a final pressure, at room temperature, of 5×10^{-7} mm being achieved. The glass dome withstood this treatment satisfactorily, no leaks or sucking in of the glass dome being experienced. The tuning range, Q and R_{sh} were again measured after this processing, no changes being observed. The results are shown in Table 2.

In this case the tuning range was less than that of the first cavity, the rod penetrating 2mm into the cavity instead of 3mm as previously. Within the limits of experimental error, discussed previously, the results confirm measurements made with the first cavity.

TABLE 2

Experimental results for tuner 2

CAVITY CONDITION	Q	R_{sh}
No tuner or dome	3200	60 500
Glass dome in position	2400	46 000
Tuner only (fully in)	2850	56 500
Glass dome in position and tuner fully in	2450	51 000
Tuning Range of Copper Tuner; Penetration 2mm		
Without glass dome		62Mc/s
With glass dome		57Mc/s



An experimental two-cavity amplifier klystron with a glass dome tuner in the output cavity. The larger tuner on the input cavity is of the bellows type.

Conclusions

In the tuning mechanism described, the use of a flexible bellows system has been avoided by enclosing the inner rod in a thin glass envelope. While this has the advantage of eliminating the difficulties encountered with a bellows mechanism the reduction in cavity shunt impedance caused by the presence of glass in the cavity may partially offset this advantage.

This, of course, is only true in a valve operating under conditions such that the cavity properties are important compared with beam loading, as, for example, in a low d.c. level amplifier. In this case the reduced gain resulting from the use of glass dome tuners would be increased by the addition of an extra cavity other conditions, beam focusing etc., permitting.

The small change in cavity resonant frequency with insertion of the glass tube indicates that the losses introduced are largely dielectric and would therefore be reduced by the use of a lower loss glass than the Kodial glass employed. A suitable glass would be Dow Corning 7070 with a loss factor approximately 20 per cent of that of Kodial. In this way the main disadvantage of the method would be largely overcome. No tests have been made with DC7070 glass but it is not anticipated that the use of this glass would involve any particular constructional problem.

No constructional difficulties have been encountered in the sealing or brazing of the glass dome assembly. A breakage of the dome in a valve would, however, be more difficult to repair than a similar mishap with a bellows assembly.

The measurements have been made at 3cm and it is thought that the use of such a mechanism at lower wavelengths would be prohibited by the increasing delicacy of the glass work. At longer wavelengths, however, the construction would be correspondingly more robust.

The device has been considered primarily for use in a low power tube. At high d.c. power levels the presence of stray high energy electrons would probably debar its use.

The use of thin walled glass tubing in the way described would also be effective in a device requiring the coupling of two cavities, only one of the cavities being under vacuum. It would obviously be possible to vary the coupling very easily.

Acknowledgments

The author wishes to thank the Manager of Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this article.

REFERENCES

1. COLLINS, G. Microwave Magnetrons. p. 713 (McGraw-Hill, 1949).
2. SPROULL, R. L., LINDER, E. G. Resonant Cavity Measurements. *Proc. Inst. Radio Engrs. 34*, 305 (1946).
3. BARLOW, H. M., CULLEN, A. L. Microwave Measurements. Ch. 8 (Constable, 1950).

Communications via Meteors

Redifon radio transmitters are being employed in a new communications system in which the trails of minute meteors are used to reflect radio messages round the curvature of the earth.

This new communications technique has been developed by Canada's Defence Research Board under the code-name "Janet".

Billions of tiny meteors no larger than grains of sand enter the earth's atmosphere every hour. Each one, as it flashes through, leaves a long trail of ionized particles in the region of 60 miles up. Project "Janet" uses these trails to reflect v.h.f. radio waves over long distances. Experiments show that signals can be transmitted by this method up to 1 000 miles.

The equipment used in the technique is relatively simple. Each circuit employs two ground stations incorporating a transmitter and receiver and carefully aligned aerials. Tht transmitters and receivers are on continuously, but no messages are passed until a meteor passes the right spot in the atmosphere. The two receivers immediately sense the "opening" of the radio circuit and automatically switch on devices which feed messages to their respective transmitter at very high speed (approximately 2 000 words/min). Because each trail can only be used for periods varying from milliseconds to a few seconds, the messages have to be recorded and stored before transmission, then sent in short, fast "bursts". The transmission speed is too great for standard teletype equipment and incoming messages are therefore also stored by special electronic equipment which prints them at normal speeds during the intervals between transmissions.

The exceptionally high speed of the transmissions, coupled with the fact that hundreds of usable ionized trails are formed every hour between any two fixed points, permits the passing of many messages in a relatively short time.

The "Janet" system has been developed to provide an economical alternative to the ionospheric scatter system. It appears to be largely unaffected by electrical and cosmic disturbances such as northern lights, electric storms etc. Further, the spasmodic and limited nature of the reflecting medium ensures a high degree of privacy.

The fact that relatively low power can be used should show substantial savings in the costs of stations. Further, the aerial systems appear to be much simpler.

As the system will operate in the 30 to 60Mc/s band and uses a bandwidth of only about 3kc/s, this will bring relief to the heavily over-crowded conventional h.f. band.

The Analysis of Three-Terminal Null Networks

By T. H. O'Dell*

A method of analysis is described which is particularly suitable for three-terminal networks that exhibit infinite attenuation at one frequency. The popular twin-T and bridged-T networks are analyzed as examples.

THERE are a number of three-terminal networks such as the twin-T and the bridged-T, that exhibit infinite attenuation at one frequency. These networks are frequently termed null networks and are used as filters, in oscillators and in selective amplifiers.

The complete analysis of these networks can become very lengthy and cumbersome by conventional methods. The method described below enables an expression for the output voltage to be written down after only four simple operations, because the symmetry of the equations used reduces confusion of terms and makes common factors easy to identify.

The Twin-T

Consider the network shown in Fig. 1(a). This is a twin-T network of three inductors and three resistors, the shunt sections are k_1R and $jk_2\omega L$, where k_1 and k_2 are real positive numbers.

Fig. 1(a) can be redrawn as Fig. 1(b). Provided the network is driven by a constant voltage generator the two circuits are identical. An application of Thevenin's Theorem to both sides of Fig. 1(b) gives Fig. 1(c) where:

$$e_1 = \frac{V_{in}\, k_2 j\omega L}{R + k_2 j\omega L} \qquad (1)$$

$$Z_1 = \frac{R(2k_2 j\omega L + R)}{R + k_2 j\omega L} \qquad (2)$$

$$e_2 = \frac{V_{in}\, k_1 R}{k_1 R + j\omega L} \qquad (3)$$

$$Z_2 = \frac{\omega L(2k_1 jR - \omega L)}{k_1 R + j\omega L} \qquad (4)$$

From Fig. 1(c) it is seen that:

$$V_{out} = \frac{e_1 Z_2 + e_2 Z_1}{Z_1 + Z_2} \qquad (5)$$

The substitution of equations (1), (2), (3) and (4) into equation (5) gives:

$$V_{out}/V_{in} = \frac{(k_1R^3 - 2k_1k_2\omega^2L^2R) + j(2k_1k_2\omega LR^2 - k_2\omega^3L^3)}{(2k_2j\omega LR + R^2)(k_1R + j\omega L) + (2k_1j\omega LR - \omega^2L^2)(R + k_2j\omega L)} \qquad (6)$$

For the output voltage to be zero at a frequency ω_0 the numerator of equation (6) must be zero at ω_0.

Thus:

$$\omega_0^2 = (R^2/2k_2L^2) = (2k_1R^2/L^2) \qquad (7)$$

$$\therefore\ k_1k_2 = \tfrac{1}{4} \qquad (8)$$

From equation (8) it is clear that there are an infinite number of possible networks that exhibit a null at a given frequency. All these networks have different characteristics.

Consider the case when $k_1 = k_2 = \frac{1}{2}$.

And let:

$$p = (\omega/\omega_0) \qquad (9)$$

Then:

$$V_{out}/V_{in} = \frac{(1 - p^2)}{(1 - p^2) + 4jp} \qquad (10)$$

* *King's College, University of Durham.*

Equation (10) leads at once to the required expressions for the magnitude and phase of the output voltage.

$$|V_{out}/V_{in}| = \frac{|(1 - p^2)|}{\sqrt{[(p^4 + 14p^2 + 1)]}} \qquad (11)$$

$$\tan\phi = [-4p/(1 - p^2)] \qquad (12)$$

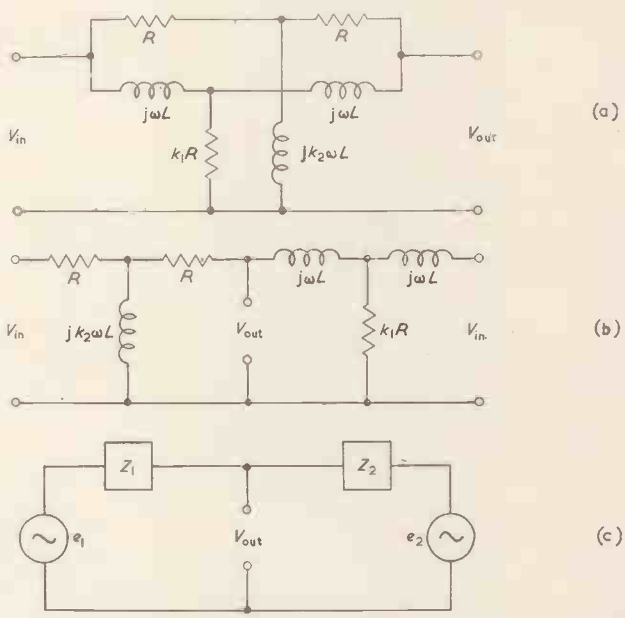


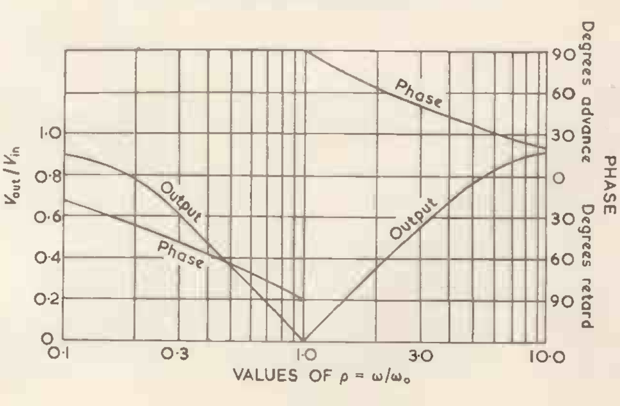
Fig. 1(a) Twin-T network (b) Rearrangement of (a) for analysis (c) Application of Thevenin's Theorem to (b)

Where ϕ is the angle between the input and output voltages. ϕ varies from $+\pi/2$ to $-\pi/2$, a positive angle representing a phase advance and *vice versa*.

The phase and output voltage functions for this particular case are shown in Fig. 2, plotted over two decades.

The other possible values for k_1 and k_2 give just such simple results.

Fig. 2. Characteristics of circuit of Fig. 1(a)



For example when $k_1 = 1$ and $k_2 = \frac{1}{4}$:

$$V_{out}/V_{in} = \frac{(1-p^2)}{(1-p^2) + 3\sqrt{2}jp} \quad \ldots\ldots\ldots (13)$$

It can be seen by inspection of equations (10) and (13) that equation (10) represents a slightly sharper null than equation (13). The difference is, however, small.

The Loaded Twin-T

Fig. 3(a) shows a very popular network, the Wien bridge, that is fed from a constant voltage generator and loaded by the resistor R_L.

The procedure for analysis is exactly the same as in the first case, but to preserve symmetry the load resistor is split into two halves, each of value $2R_L$, in parallel. This is shown in Fig. 3(b).

The application of Thevenin's Theorem to each side produces Fig. 1(c) again, where in this case:

$$e_1 = \frac{V_{in}\, 2k_4 R R_L}{(2k_4 R R_L + X_c^2 + 2R_L X_c + 2k_4 R X_c)} \quad \ldots\ldots (14)$$

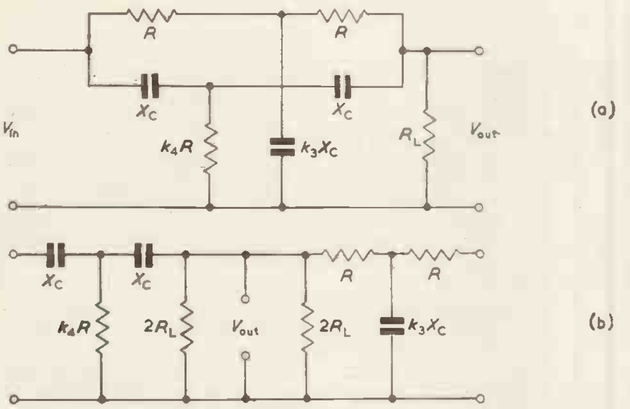


Fig. 3(a). Wien bridge network
(b) Rearrangement of (a) for analysis

$$Z_1 = \frac{2R_L(X_c^2 + 2k_4 X_c R)}{(2k_4 R R_L + X_c^2 + 2R_L X_c + 2k_4 R X_c)} \quad \ldots\ldots (15)$$

$$e_2 = \frac{V_{in}\, 2k_3 R_L X_c}{(2k_3 R_L X_c + R^2 + 2RR_L + 2k_3 R X_c)} \quad \ldots\ldots (16)$$

$$Z_2 = \frac{2R_L(R^2 + 2k_3 X_c R)}{(2k_3 R_L X_c + R^2 + 2RR_L + 2k_3 R X_c)} \quad \ldots\ldots (17)$$

where $X_c = -j/\omega C$.

By substituting equations (14), (15), (16) and (17) in equation (5):

$$V_{out}/V_{in} = \frac{2k_4 R R_L(R^2 + 2k_3 X_c R) + 2k_3 X_c R_L(X_c^2 + 2k_4 X_c R)}{(X_c^2 + 2k_4 X_c R)(R^2 + 2RR_L + 2k_3 R X_c + 2k_3 R_L X_c) + (R^2 + 2k_3 X_c R)(X_c^2 + 2X_c R_L + 2k_4 X_c R + 2k_4 R R_L)} \quad \ldots\ldots (18)$$

Thus for null:

$$\omega_0^2 = (1/2k_4 C^2 R^2) = (2k_3/C^2 R^2) \quad \ldots\ldots\ldots (19)$$

$$\therefore k_4 k_3 = \tfrac{1}{4} \quad \ldots\ldots\ldots (20)$$

Again there are an infinite number of possible networks for all the values given for k_3 and k_4 by equation (20).

As the expression given by equation (18) is dimensionless, equation (18) can be expressed in terms of dimensionless parameters in the same way as equation (6). Two parameters will be needed, let these be:

$$p = (\omega/\omega_0) \quad \ldots\ldots\ldots (21)$$

$$n = (R/R_L) \quad \ldots\ldots\ldots (22)$$

And let:

$$k_3 = k_4 = \tfrac{1}{2} \quad \ldots\ldots\ldots (23)$$

Now equation (18) becomes:

$$V_{out}/V_{in} = \frac{(1-p^2)}{(1 + 2n - p^2) + j(4p + 2np)} \quad \ldots (24)$$

Equation (24) gives the required function for the magnitude of the output as:

$$|V_{out}/V_{in}| = \frac{|(1-p^2)|}{\sqrt{[(2n+1)^2 + p^2(14 + 12n + 4n^2) + p^4]}} \quad (25)$$

Note that equations (24) and (10) are identical when $n = 0$, i.e. when the network is unloaded.

The function given by equation (25) is plotted over two decades for $n = 0$ and $n = 0{\cdot}5$ in Fig. 4. It can be seen that the load has a marked effect upon the characteristic.

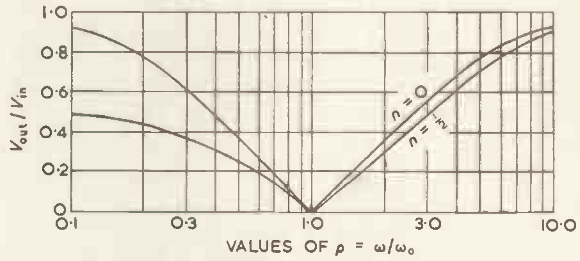


Fig. 4. Characteristics of circuit of Fig. 3(a)

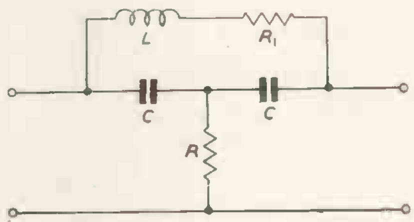


Fig. 5. Bridged-T network

The Bridged-T

The circuit shown in Fig. 5 is one type of bridged-T in common use.

By following the previous lines of analysis:

$$e_1 = \frac{V_{in} R}{R - j/\omega C} \quad \ldots\ldots\ldots (26)$$

$$Z_1 = \frac{-1/\omega^2 C^2 - 2jR/\omega C}{R - j/\omega C} \quad \ldots\ldots\ldots (27)$$

$$e_2 = V_{in} \quad \ldots\ldots\ldots (28)$$

$$Z_2 = R_1 + j\omega L \quad \ldots\ldots\ldots (29)$$

Substitution in equation (5) gives:

$$V_{out}/V_{in} = \frac{(RR_1 - 1/\omega^2 C^2) + j(\omega LR - 2R/\omega C)}{(RR_1 + L/C - 1/\omega^2 C^2) + j(\omega LR - 2R/\omega C - R_1/\omega C)} \quad \ldots (30)$$

Thus for a null at ω_0:

$$\omega_0^2 = (1/C^2 R R_1) = (2/LC) \quad \ldots\ldots\ldots (31)$$

From equation (31) any one of the four components L, C, R, or R_1, may be put in terms of the other three.

Thus:

$$L = 2CRR_1 \quad \ldots\ldots\ldots (32)$$

Let $p = (\omega/\omega_0)$

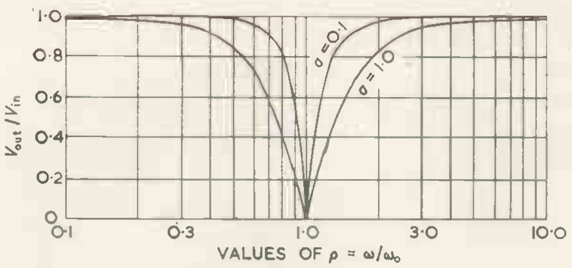


Fig. 6. Characteristics of circuit of Fig. 5

and $a = (R_1/R)$.

Then:

$$V_{out}/V_{in} = \frac{(p^2 - 1)}{(p^2 - 1) - \sqrt{a}jp} \quad \ldots\ldots\ldots (33)$$

Equation (33) shows that a bridged-T network of this type is very versatile because the value of a can be varied over a wide range. By inspection it can be seen that the null becomes sharper as a is reduced. Fig. 6 shows the response of the network over two decades for $a = 1·0$ and $a = 0·1$.

Design Considerations

In the design of a network of the twin-T type advantage can be taken of the type described, in which $k_1 = 1·0$, $k_2 = \frac{1}{4}$, or *vice versa*. In this case three of the six components are identical and this will reduce the cost of a variable network. The additional sharpness obtained from a symmetrical type is, for many purposes, too small to be of importance.

In the bridged-T, of the type described, the parameter a can be varied over a very wide range. It is possible to reduce the value of a below 10^{-4} by using an inductor of high Q factor, and this will result in an extremely sharp null.

The effect of stray capacitance can be minimized by making the resistors in a network as low as possible. In the twin-T, and to a lesser degree in the bridged-T, the components can be varied in proportion to one another with no change in theoretical performance at all.

ERNIE

(Electronic Random Number Indicator Equipment)

In the Government's "Premium Savings Bonds" Scheme, which starts next November, it is necessary to make a draw each month to determine the prizewinners and, due to the large numbers that are likely to be involved it has been decided that the only feasible method of making the draw is by electronic means. For this purpose an Electronic Random Number Indicator Equipment (ERNIE) has been evolved.

The Electronic Random Number Indicator Equipment was designed and will be built at the Post Office Engineering Research Station to meet the following operational requirements:—

(1) Each number will contain 9 digits;

(2) Each 9 digit random number will be associated with a 5 digit sequential draw number;

(3) The second and third digits of the random number will be alphabetical characters, this means that for the second digit, counting must be in a scale of 23 (I, O and U are not used), for the third digit only ten letters will be used so that counting can be arranged as for the "numerical" digits;

(4) The final output of the equipment will be presented on a page-printing teleprinter and also on gummed tape;

(5) Associated with the equipment will be a distributor which will automatically route each random number to the appropriate one of a maximum of 20 teleprinters located at suitable points where the numerical registers of bonds are kept.

The starting point for the generation of each random digit is a source of electrical noise. That used is a gas discharge tube containing neon. When an electrical potential is applied to such a tube a discharge occurs which is manifest in a visible glow. In the tube the gas molecules move about in a random manner. The current which arrives at the anode consists of electrons whose passage through the tube has been subject to collision with the molecules of gas. They therefore arrive in varying numbers from instant to instant, this giving rise to random variations of current which can be detected by a suitable amplifier. By the use of a "slicing" valve it is possible to neglect the preponderant number of changes which are of low amplitude and to amplify and "square up" the remainder to operate a counter. In the noise generator the slicing level is such that about 3 000 such "squared up" pulses are recorded every second.

The number of pulses received by the counter in any one period will vary over wide limits, but there will be some average number and numbers near this average will occur more frequently than much smaller or much larger numbers of pulses. If the average number is large enough, as it is under the conditions of counting in the equipment, the number of pulses counted will terminate in the digits 0, 1, 2 up to 9 with equal frequency, subject to the conditions stated in the following paragraph.

If the counter started from zero every time, the digits produced would not—in theory—be equally likely, even though—in practice—the differences would be negligible. The digits can, however, be made equally likely by having different starting points for the counter and arranging that all starting points occur in the same proportions. There are various theoretically satisfactory ways of doing this, of which the most convenient is simply to start the counter from where it was left at the end of the previous count.

The foregoing description relates to the production of a succession of random numbers, or letters, relating to one digit of the Bond number. In the complete equipment 9 sets of noise generators will drive 9 separate counters, one for each of the 9 digits. Eight of these counters read from 0 to 9 and one, in order to meet the requirements of the "alphabetical" digit, reads from 0 to 22. The counters are stopped simultaneously at fixed intervals. After the counters have been read the noise generators are allowed to drive the counters again. The counters move on from the position in which they were left so that the theoretical requirements are satisfied.

In order to safeguard the randomness of number generation and in order to ensure that there is no correlation between individual digits of the 9 digit number, each digit will be generated separately as already stated. To safeguard the possibility of any failure, each generator together with its associated counter, is provided in duplicate. The two counter outputs thus obtained will be added in combining units having 10 outputs. This safeguards the possibility of bias appearing in the number presented at the output due to a fault condition in one of the electronic tubes used for counting.

The outputs from the 9 combiners are transferred to primary storage equipment from which they, together with the fixed codes (figure shift, letter shift, line feed, carriage return and spaces) needed for the printing of the number on a teleprinter are assembled and scanned and fed to the teleprinters together with the sequential draw number.

A separate master pulse generating unit will provide the basic timing signals which start and stop the counting and control the teleprinters. The signals from the output stores will be released at a rate of about one every three or four seconds into the teleprinter distribution network.

In order that ineffective search time may be reduced to a minimum, arrangements are incorporated in the primary number storage for the inhibition of random numbers generated above a certain value, i.e., corresponding to ranges of bonds not yet offered for sale. This value can be pre-set for each of the 23 denominational codes separately. This "inhibitor" will be adjusted immediately before each draw on the basis of the maximum number of bonds sold in any particular denomination.

The equipment will also examine the first two digits of each number and as a result of that examination select and route the information to the appropriate one of a number of teleprinter stations located at the points where the numerical registers of Bonds are kept.

Characteristics for Half-Wave Rectifier Circuits

By H. A. Enge*, Dr.Philos.

Voltage regulation characteristics, form factor, first harmonic, and peak current values are given for half wave rectifiers with common type loads. All in non-dimensional variables.

CALCULATIONS of currents and voltages in rectifier circuits are too tedious to be carried out for every individual design. For simple circuits the number of parameters is fortunately so low that results of general validity, once obtained, can be presented in the form of curves.

The main purpose of this article is to give such curves for half-wave rectifier circuits with resistive, inductive and battery load. A sufficiently large, externally loaded capacitor can also be regarded as a battery in this connexion. The variables used are non-dimensional. Hence the results can be used for predicting characteristics for high power as well as low power half-wave rectifiers.

Rectifier Elements

The rectifier element is the non-linear circuit element that accomplishes the rectification (e.g. selenium rectifier,

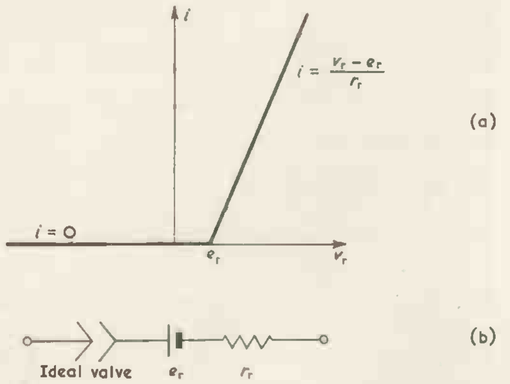


Fig. 1. Rectifier element characteristic and equivalent diagram

vacuum tube etc.). The most important properties of the rectifier element are exhibited through the *element characteristic* giving instantaneous current as a function of voltage across the element.

Most rectifier elements have characteristics which to a good approximation can be represented by the curve in Fig. 1(a). An equivalent diagram of such an element is shown in Fig. 1(b). The double arrow symbol is used to designate an ideal electric valve, i.e. a valve which offers no resistance against a positive current in direction of the arrow and infinite resistance against a current in the opposite direction. The relationship between voltage drop and current through the element in this idealized case is:

$$i = 0 \qquad \text{for } v_r \leq e_r \qquad \qquad (1)$$

$$i = \frac{v_r - e_r}{r_r} \text{ for } v_r > e_r$$

Selenium and copper oxide rectifiers both have element characteristics that come quite close to the one in Fig. 1(a). For most selenium rectifiers e_r is approximately 0·5V per plate and r_r is of the order of 10Ω-cm² per plate. The corresponding numbers for copper oxide rectifiers are of the order of 0·2V and 5Ω-cm².

Gas-filled hot cathode rectifiers have almost constant voltage drop when conducting, corresponding to $r_r \simeq 0$. The constant voltage drop is of the order of $e_r \simeq 10$ to 15V.

The characteristics of vacuum tube rectifiers follow approximately a power law, $i = kv^a$. Even such a simple relationship between current and voltage drop would introduce considerable difficulties in the calculations unless the exponent a were equal to unity. One is thereby lead to try

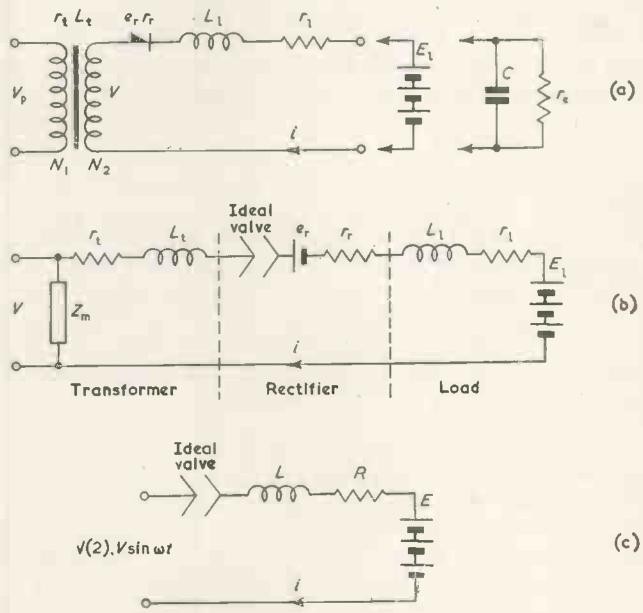


Fig. 2. Half-wave rectifier circuit

(a) Wiring diagram, (b) equivalent diagram, (c) components combined

the relationship (1) also for vacuum tube rectifiers. The characteristic for one section of the valves 5Y3, 5Y4, or 80 for instance, can be approximated by equation (1) with $e_r = 10$V and $r_r = 400\Omega$. (For other characteristics see Reference 2, page 1171).

Half-Wave Rectifier Circuit

Fig. 2(a) shows a half-wave rectifier circuit with a load consisting of an inductor, a resistor, and a battery in series. The battery can be replaced by a large externally loaded capacitor. The circuit generalizes the following common cases:

(1) Battery charging. Usually the load inductance is then zero (or negligible).

(2) Pure resistive load $L_l = 0$ and $E_l = 0$.

(3) Inductive load. $E_l = 0$.

(4) Rectifier with large input filter capacitor. If r_e is the total external load resistance across the capacitor (including resistance of filter chokes etc.), the load "electromotive force" is:

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$$E_1 = Ir_e \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

Here I is the direct current in the circuit. For further calculations, the capacitor and external load is replaced by a battery of e.m.f. given by equation (2). This can be done only when the time-constant of capacitor and external load circuit is considerably larger than one cycle of the line voltage, or:

$$r_e C \gg (1/f) \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

where f is the frequency. Special cases for which this condition is not fulfilled to an adequate degree have been treated in the literature[1-4].

Fig. 2(b) shows the same circuit as Fig. 2(a), but equivalent diagrams have replaced the transformer and rectifier symbols. The magnetizing impedance Z_m constitutes a circuit of its own. It is assumed to be directly connected to the input and hence can be disregarded in the further treatment of the rectifier circuit. The input voltage in the equivalent circuit is the open-circuit secondary (r.m.s.) voltage of the transformer:

$$V = V_p (N_2/N_1) \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

where V_p is the primary (line) voltage and N_1 and N_2 are the numbers of primary and secondary turns.

The equivalent series resistance and "leakage" inductance of the transformer are those measured at the secondary terminals with the primary short-circuited or:

$$r_t = r_p (N_2/N_1)^2 + r_s = r_p' + r_s \quad \ldots\ldots \quad (5)$$

$$L_t = L_p (N_2/N_1)^2 + L_s = L_p' + L_s \quad \ldots\ldots \quad (6)$$

Subscripts p and s stand for primary and secondary respectively, and the primes indicate so-called reduced values. The total leakage inductance L_t of the transformer is much easier to calculate or measure directly than are the quantities L_p and L_s. Hence equation (6) is ordinarily unimportant.

It has been assumed that the primary of the transformer is directly connected with an a.c. power line with negligible internal resistance and inductance. If this is not the case, any series resistance and inductance in the primary should be included in r_p and L_p.

The equivalent diagram given here for the transformer is commonly used for ordinary a.c. circuits. In rectifier circuits it should be used with caution, however. It is well known that in a half-wave rectifier, the d.c. component flows only in the secondary circuit. The primary ampere-turns match the secondary ampere-turns except for this d.c. component. By transforming r_p over to the secondary circuit, one therefore introduces an error which is usually quite small, however. It will be shown in a separate section below how it can be corrected for.

The unbalance in ampere-turns also tends to saturate the magnetic core, unless this effect is considered when designing the transformer.

The various components in Fig. 2(b) are combined in Fig. 2(c), where the circuit elements are:

$$R = r_t + r_r + r_l \quad \ldots\ldots\ldots\ldots\ldots \quad (7)$$

$$L = L_t + L_l \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

$$E = E_l + e_r \quad \ldots\ldots\ldots\ldots\ldots \quad (9)$$

The transformer resistance and inductance are given by equations (5) and (6). The expression for E will be modified below when the r_p-correction is introduced.

Now, assume that the input voltage and frequency is given and also the quantities R, L and E. The instantaneous current as a function of time is sought.

Instantaneous Current, Ignition Angle and Extinction Angle

The differential equation for the circuit in Fig. 2(c) is:

$$\sqrt{2}\, V \sin \omega t = Ri + L\,(di/dt) + E \quad \ldots\ldots \quad (10)$$

when the rectifier is conducting. The angular frequency is $\omega = 2\pi f$ and V is given by equation (4). In the nonconducting part of the cycle, equation (10) should be replaced by $i = 0$, of course.

It is convenient now to introduce the *ignition angle* α. This is the value of ωt that makes the increasing input voltage equal to E, the opposing e.m.f. on the other side of the rectifier.

$$\sin \alpha = (E/\sqrt{2}V) \quad \ldots\ldots\ldots\ldots\ldots \quad (11)$$

When ωt passes the value $\omega t = \alpha$, the circuit starts conducting. The instantaneous current is thereafter given by the solution of equation (10) for a fraction of a cycle until the current again reaches zero. This happens at another angle $\omega t = \beta$, the extinction angle, which is discussed below.

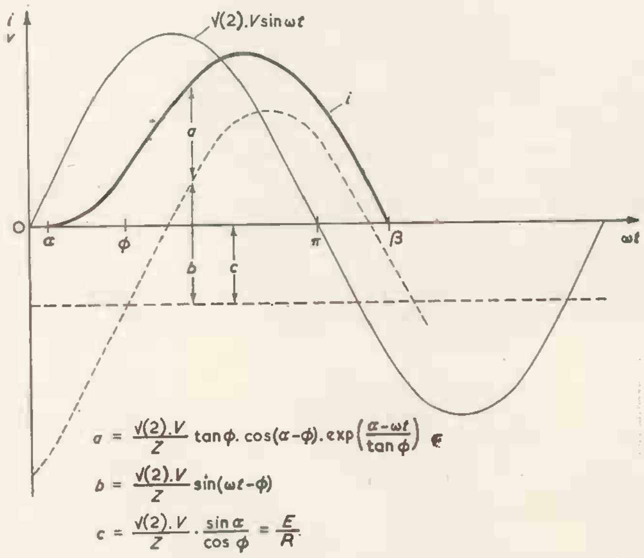


Fig. 3. Current in rectifier circuit with ignition angle $\alpha = 12°$ and phase angle $\phi = 60°$

The general solution of equation (10) is fairly simple but still not very well suited for further treatment when $R = 0$ or $L = 0$. Therefore, these two cases are treated separately after the general case.

General case: $R > 0$ and $L > 0$.

The complete solution for equation (10) is for the conductive region $\alpha < \omega t < \beta$:

$$i = (\sqrt{2}V/Z)\left[\tan\phi \cos(\alpha - \phi) . \exp\left(\frac{\alpha - \omega t}{\tan\phi}\right) + \sin(\omega t - \phi) - \frac{\sin\alpha}{\cos\phi}\right] \ldots \quad (12)$$

where

$$Z = \sqrt{(R^2 + \omega^2 L^2)} \quad \ldots\ldots\ldots\ldots \quad (13)$$

$$\tan\phi = (\omega L/R) \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

In obtaining this solution $\sqrt{2}V \sin\alpha$ is substituted for E, and the integration constant is found by making use of the initial condition: $i = 0$ when $\omega t = \alpha$.

Fig. 3 shows as an example the instantaneous current versus the time angle ωt when the ignition angle is 12° and the phase-angle is $\phi = 60°$. It is shown how the instantaneous current function is built up of the three terms in equation (12). The first term a, is what is usually called the transient term in ordinary a.c. circuits. In the present case this term is not any more transient than the other terms because it reappears with full strength every cycle. The second term, b, is the familiar stationary sinusoidal current lagging the voltage by an angle ϕ. The last term, c equal to E/R, is the direct current due to the opposing e.m.f. in the circuit.

The extinction angle β can be found as the value of ωt between α and $2\pi + \alpha$ that makes the expression (12) equal to zero, i.e. β is given implicity by:

$$\tan\phi \cdot \cos(\alpha - \phi) \cdot \exp\left(\frac{\alpha - \beta}{\tan\phi}\right) + \sin(\beta - \phi) - \frac{\sin\alpha}{\cos\phi} = 0 \quad .. (15)$$

The extinction angle is seen to be a function of the given quantities α and ϕ only. Numerical solution for β from equation (15) have been obtained for $\alpha = 0°, 15°, 30°, 45°, 60°, 75°$ and $90°$ combined with $\phi = 15°, 30°, 45°, 60°, 75°$ and $85°$. In Fig. 4 the extinction angle has been plotted versus the ignition angle α for a few values of the phase angle ϕ.

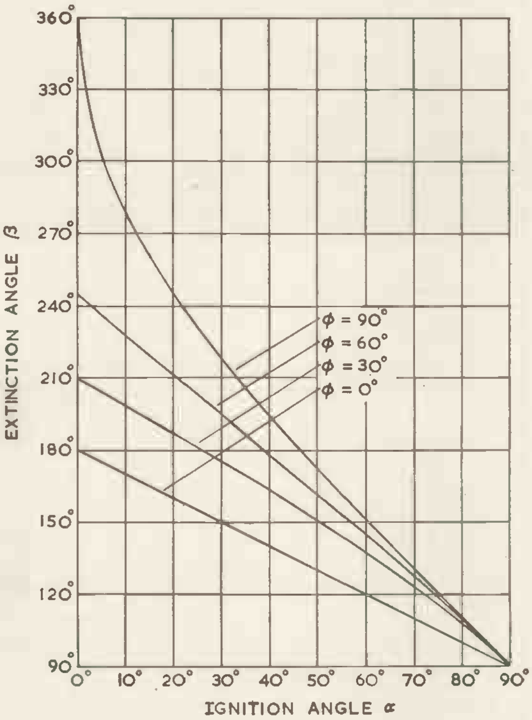


Fig. 4. Variation of extinction angle with ignition angle

HYPOTHETICAL CASE: $R > 0$ and $L = 0$. ($\phi = 0$)

Strictly speaking neither this case nor the next one can be realized in practice. When the leakage inductance of the transformer is very low and there is no load inductance, the case under discussion may be a quite good approximation, however. For $L = 0$ the solution of equation (10) is:

$$i = (\sqrt{2}V/R)(\sin\omega t - \sin\alpha) \quad \ldots\ldots\ldots (16)$$

and the extinction angle:

$$\beta = \pi - \alpha \quad \ldots\ldots\ldots (17)$$

HYPOTHETICAL CASE: $R = 0$ and $L > 0$. ($\phi = 90°$)

The solution of equation (10) for this case is:

$$i = (\sqrt{2}V/\omega L)[\cos\alpha - \cos\omega t - (\omega t - \alpha)\sin\alpha] \quad .. (18)$$

and the extinction angle is given by:

$$\cos\alpha - \cos\beta - (\beta - \alpha)\sin\alpha = 0 \quad \ldots\ldots (19)$$

Average Current (D.C.) Load Ratio

The average current in the circuit can be found by integrating equation (12) from $\omega t = \alpha$ to $\omega t = \beta$ and dividing the result by 2π. However, it is simpler to integrate equation 10 as follows:

$$I = (1/2\pi R)\int_\alpha^\beta (\sqrt{2}V\sin\omega t - L(di/dt) - E)\,d(\omega t)$$

This integration is straightforward. The result is:

$$IR/\sqrt{2}V = (1/2\pi)[\cos\alpha - \cos\beta - (\beta - \alpha)\sin\alpha] \quad .. (20)$$

Instead of the direct current I, it is more convenient to operate with the non-dimensional ratio $IR/\sqrt{2}V$ which may be called the load ratio.

As stated above, the extinction angle β is a function of α and ϕ. Hence the load ratio is also a function of the given quantities α and ϕ:

$$IR/\sqrt{2}V = f(\alpha,\phi) \quad \ldots\ldots\ldots (21)$$

Making use of the numerical solutions for β, a table of load ratios has been obtained for the values of α and ϕ mentioned in the previous section. Graphic interpolations have yielded load ratios also for other values of ϕ. The final result is presented in Fig. 5 which gives $\sin\alpha$ versus the load ratio for several values of the phase angle ϕ. The curves give in relative measure E as a function of I and are thus the sought voltage regulation characteristics. Because of the non-dimensional form of the variables, the graph has general validity for all possible values of the circuit parameters except negative E-values.

The average current in a circuit where $R = 0$ is found by integrating equation (18). After rearrangements and use of equation (19) one finds:

$$(I\omega L/\sqrt{2}V) = \frac{(\sin\beta - \sin\alpha)^2}{4\pi\sin\alpha} \quad \text{for } \phi = 90° \quad \ldots (22)$$

Equation (22) is of interest only for calculating form factor etc. for $\phi = 90°$ as a hypothetical limit case.

Correction for Lack of D.C. in Transformer Primary

As previously mentioned, an error is introduced into the calculation when the resistance of the primary circuit is "reduced" to the secondary or rectifier circuit (Cf. equation (5)). When the resistance of the primary is small, the error is also small and can usually be neglected. In other cases, however, it may be necessary to correct for the error made, and this can easily be done as follows.

It has been assumed that in the circuit of Fig. 2(b) there flows a current composed of the direct current I and a superimposed alternating current. The truth is that the a.c. component flows through the whole circuit and the d.c. component flows through the whole circuit except part of the transformer resistance r_t. In other words, into the calculations has been entered a constant voltage drop Ir_p' that does not exist. The curves of Fig. 5 together with equation (9) hence yield a too low output voltage (load voltage E_1) for a given value of the direct current I. Equation (9) should therefore be replaced by:

$$E = E_1 + e_r - Ir_p' \quad \ldots\ldots\ldots (9a)$$

Form Factor

The effective or root mean square value of the current flowing through the rectifier circuit (not the transformer primary) is given by the following integral:

$$J^2 = (1/2\pi)\int_\alpha^\beta i^2 d(\omega t) \quad \ldots\ldots\ldots (23)$$

The form factor is the ratio between the r.m.s. value and the d.c. value: $F = J/I$. The integral (23) is evaluated most easily by making use of equation (10) multiplied by i. The resulting formula for the square of the form factor is:

$$F^2 = (\sqrt{2}V/IR)[\sqrt{(2V/IR)}(\cos\phi/2\pi)(S\sin\phi + T\cos\phi) - \sin\alpha] \quad (24)$$

where the load ratio is given by equation (20) and:

$$S = \int_\alpha^\beta (\sin\omega t - \sin\alpha)\cos\omega t\, d(\omega t) = \tfrac{1}{2}(\sin\alpha - \sin\beta)^2 \quad .. (25)$$

$$T = \int_\alpha^\beta (\sin\omega t - \sin\alpha)\sin\omega t\, d(\omega t) = \tfrac{1}{2}(\beta - \alpha) + \sin\alpha\cos\beta - \tfrac{1}{4}\sin 2\alpha - \tfrac{1}{4}\sin 2\beta \ldots (26)$$

The above equation for the form factor cannot be used

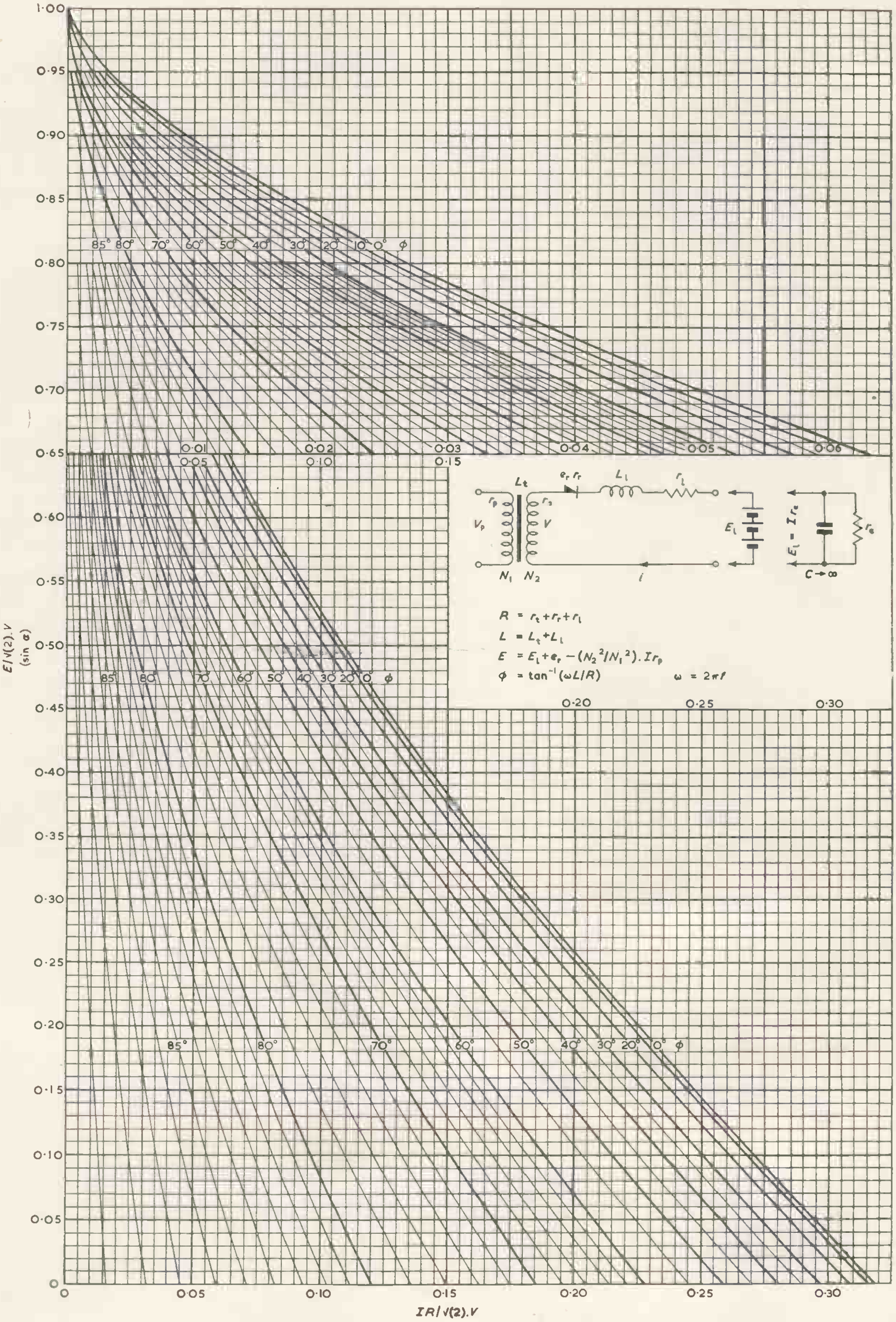


Fig. 5. Variation of the ratio (opposing e.m.f./a.c. peak voltage) with load ratio

for $\alpha = 90°$ ($R = 0$). A similar formula is obtained, however, for this case by substituting equation (18) into equation (23) and by use of equation (22).

For values of $\sin\alpha$ close to unity, equation (24) is not very well suited for numerical calculations. Power expansions yield the following approximate formulas, valid when $(\pi/2) - \alpha \ll 1$:

$$F \simeq \sqrt{\left[\frac{6\pi}{5[(\pi/2) - \alpha]}\right]} \text{ for } \phi = 0° \quad \ldots\ldots (27)$$

$$F \simeq \sqrt{\left[\frac{32\pi}{35[(\pi/2) - \alpha]}\right]} \text{ for } \phi = 90° \quad \ldots\ldots (28)$$

Equation (27) is correct within 1·5 per cent in the whole region of α from 0 to $\pi/2$.

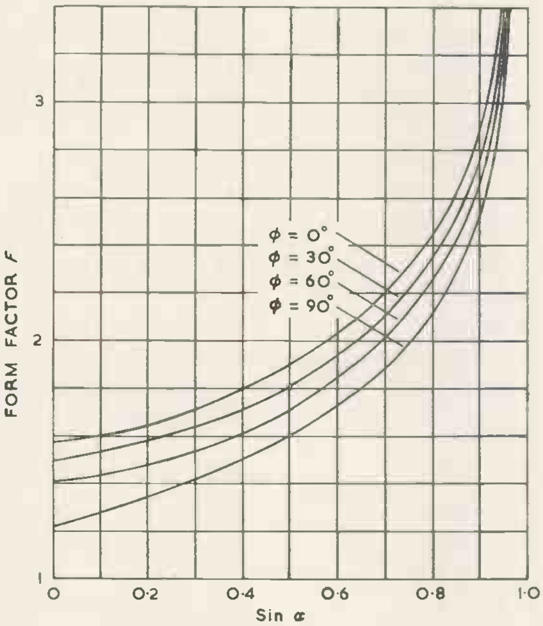


Fig. 6. Variation of form factor with sin α

The form factor has been plotted in Fig. 6 versus $\sin\alpha$ for $\phi = 0°$, 30°, 60°, and 90°.

In the primary circuit of the transformer the r.m.s. value of the current is:

$$J_p = (N_2/N_1)\sqrt{(J^2 - I^2)} = (N_2I/N_1)\sqrt{(F^2 - 1)} \quad \ldots (29)$$

First Harmonic of the Current

When the load in the rectifier circuit is of the type indicated in Fig. 2(a) to the right, there is always a ripple on the voltage across the capacitor. Unless this ripple voltage is small, the calculations carried out above are not valid.

When the inequality (3) is fulfilled to an adequate degree, the approximate magnitude of the first harmonic of the ripple voltage is:

$$V_{c1} = (J_1/\omega C) \quad \ldots\ldots\ldots\ldots (30)$$

Here J_1 is the r.m.s.-value of the first harmonic of the rectified current. It can be expressed as:

$$J_1 = \sqrt{[\tfrac{1}{2}(a_1^2 + b_1^2)]} \quad \ldots\ldots\ldots\ldots (31)$$

where:

$$a_1 = (1/\pi)\int_\alpha^\beta i\cos\omega t\, d(\omega t) = \frac{\sqrt{2}V}{R} \cdot \frac{\cos\phi}{\pi}(S\cos\phi - T\sin\phi) \quad \ldots\ldots\ldots\ldots (32)$$

$$b_1 = (1/\pi)\int_\alpha^\beta i\sin\omega t\, d(\omega t) = \frac{\sqrt{2}V}{R} \cdot \frac{\cos\phi}{\pi}(S\sin\phi + T\cos\phi) \quad \ldots\ldots\ldots\ldots (33)$$

The integrals S and T are given by equations (25) and (26).

It is convenient to give the ratio between the r.m.s. of the first harmonic and the direct current. This is:

$$h_1 = (J_1/I) = (\sqrt{2}V/IR)(\cos\phi/\sqrt{2\pi})\sqrt{(S^2 + T^2)} \quad \ldots\ldots\ldots\ldots (34)$$

This relative first harmonic has been plotted in Fig. 7 versus $\sin\alpha$ for $\phi = 0°$, 30°, 60°, and 90°.

Peak Forward Current

For some types of rectifier elements the peak forward current must stay below a certain upper limit. For a specific design it is therefore important to be able to compute this

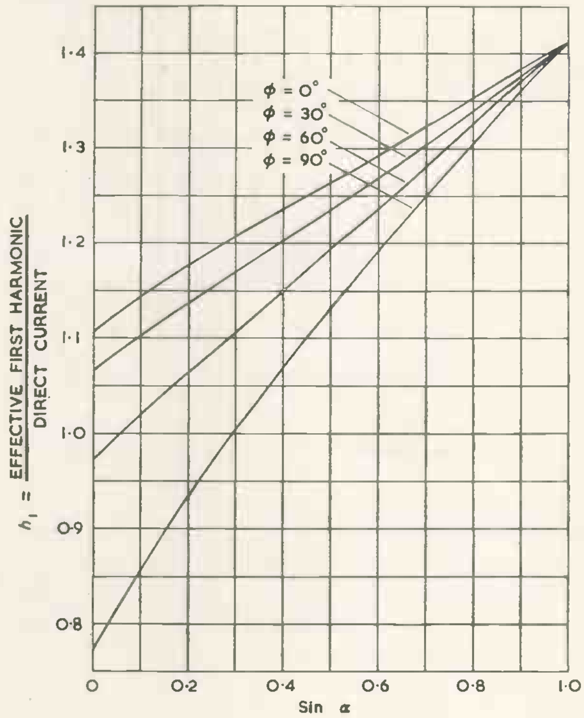


Fig. 7. Variation of relative first harmonic with sin α

peak value. It is found, of course, as the maximum of the time function of the instantaneous current. Only two cases have been considered here, namely $\phi = 0$ and $\phi = 90°$. The instantaneous current functions are given by equations (16) and (18) respectively.

The ratios between the maximum current and the direct current for the two cases are:

$$i_{max}/I = (\sqrt{2}V/IR)(1 - \sin\alpha) \text{ for } \phi = 0 \quad \ldots\ldots\ldots (35)$$

$$i_{max}/I = (\sqrt{2}V/I\omega L)(2\cos\alpha + 2\alpha\sin\alpha - \pi\sin\alpha) \text{ for } \phi = 90° \quad \ldots\ldots\ldots\ldots (36)$$

For values of $\sin\alpha$ close to unity equations (35) and (36) are not very well suited for numerical calculations. Power expansions yield the following approximate formulas valid when $(\pi/2) - \alpha \ll 1$:

$$i_{max}/I \simeq \frac{3\pi}{2[(\pi/2) - \alpha]} \quad \text{for } \phi = 0 \quad \ldots\ldots (37)$$

$$i_{max}/I \simeq \frac{32\pi}{27[(\pi/2) - \alpha]} \quad \text{for } \phi = 90° \quad \ldots\ldots (38)$$

The two peak current ratios are plotted in Fig. 8 versus $\sin\alpha$. For any intermediate value of ϕ, the ratio lies between the limits given by the two curves.

Peak Inverse Voltage

The maximum voltage across the rectifier in the non-conducting part of the cycle is:

$$V_{max} = \sqrt{2}V + E_1 \quad \ldots\ldots\ldots\ldots (39)$$

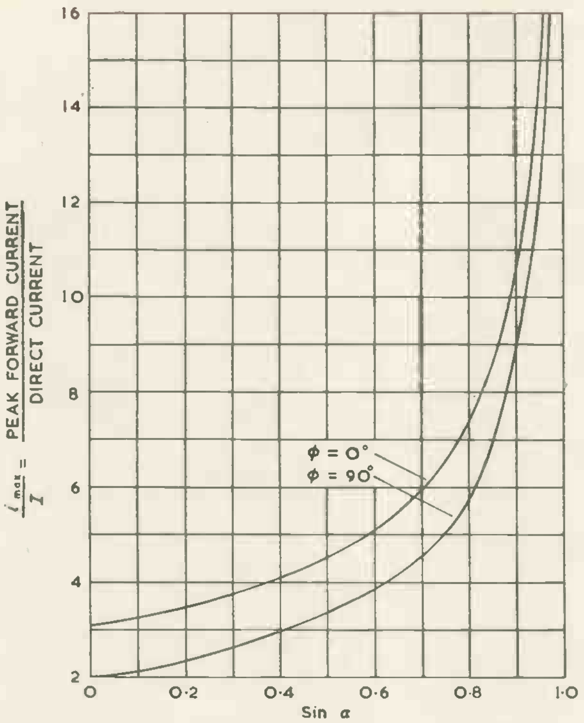


Fig. 8. Variation of peak current ratio with sin α

Example

A selenium rectifier in a half-wave rectifier circuit with capacitor input filter shall deliver 50mA to the load.

Transformer data: $V = 350\text{V}$, $r_s = 120\Omega$, $L_t = 0{\cdot}16\text{H}$
$V_p = 220\text{V}$, $r_p = 30\Omega$, $r_p' = 76\Omega$.

Selenium rectifier data: $e_r = 15\text{V}$, $r_r = 100\Omega$.

Line frequency: $f = 50\text{c/s}$.

Capacitor: $C = 32\mu\text{F}$.

Compute: (a) The d.c. voltage across the capacitor, (b) the effective current values in transformer primary and secondary, (c) the first harmonic of the ripple voltage across the capacitor, and (d) the maximum instantaneous forward current in the circuit.

Total resistance: $R = r_p' + r_s + r_r = 296\Omega$.

Load ratio: $IR/\sqrt{2}V = \dfrac{0{\cdot}05 \times 296}{\sqrt{2} \times 350} = 0{\cdot}0299$

Phase-angle: $\phi = \tan^{-1}\left(\dfrac{2\pi \times 50 \times 0{\cdot}16}{296}\right) = 9{\cdot}62°$

(a) From Fig. 5 one finds $\sin\alpha = 0{\cdot}780$, which gives:

$E = \sqrt{2}V \sin\alpha = \sqrt{2} \times 350 \times 0{\cdot}780 = 386\text{V}$

Equation (9a) yields: $E_1 = E - e_r + Ir'_p = 386 - 15 + 3{\cdot}8 \simeq 375\text{V}$

(b) From Fig. 6 one finds $F = 2{\cdot}37$, which gives:

$J = FI = 2{\cdot}37 \times 50 = 118{\cdot}5\text{mA}$, and

$J_p = (N_2/N_1)\, I\, \sqrt{(F^2 - 1)} = (350/220) \times 50 \times \sqrt{(2{\cdot}37^2 - 1)} = 171\text{mA}$.

(c) From Fig. 7 one finds $J_1/I = 1{\cdot}34$, which gives:

$$V_{c1} = (J_1/\omega C) = \frac{1{\cdot}34 \times 0{\cdot}05}{2\pi \times 50 \times 32 \times 10^{-6}} = 6{\cdot}67\text{V r.m.s.}$$

(d) Finally, from Fig. 8 one finds $i_{max}/I = 7{\cdot}0$ (approx.) which gives $i_{max} = 7{\cdot}0 \times 50 \doteq 350\text{mA}$.

Acknowledgments

The author wishes to thank the Professor at the Technical University of Norway, A. Aanderud, who suggested rectifier circuit theory as a thesis work and encouraged the pursuance of the subject.

REFERENCES

1. Schade, O. H. Analysis of Rectifier Operation. *Proc. Inst. Radio Engrs.* *31*, 341 (1943).
2. Langford-Smith, F. Radio Designer's Handbook. 4th Edn. Ch. 30. (Iliffe & Sons, Ltd., 1954).
3. Terman, F. E. Radio Engineers' Handbook 1st Edn., Sec. 8 (McGraw-Hill, 1943).
4. Gray, T. S. Applied Electronics. 2nd Edn. Ch. 6. (John Wiley & Sons, 1954).

S.H.F. Radio System for Switzerland

The Swiss Posts and Telegraphs Dept has recently placed an order with Standard Telephone et Radio S.A. Zurich, for S.H.F. Radio Equipment, to link Berne with Geneva, providing 600 telephone circuits over a single radio channel, with facilities for future extensions.

The equipment will be designed and manufactured in England by Standard Telephones and Cables Limited (associated with Standard Telephone et Radio S.A. Zurich) and will comprise terminals at Berne and Geneva, with a repeater station at Chasseral.

The transmission path lies over mountainous country, and the section Chasseral-Geneva is some 70 miles (112 km) long. This will be accomplished with the Standard S.H.F. system incorporating the newly-developed Standard 5W travelling-wave amplifier.

The equipment will comprise a system equipped initially with one working and one space r.f. channel for 600 telephone circuits, i.e., ten supergroups. Normal expansion will enable this to be extended to six r.f. channels in the band 3 800-4 200 Mc/s, and if necessary additional channels can be added in the band 3 600-3 800Mc/s. These additional r.f. channels can be used for further telephone circuits or for television links. The frequency division channelling equipment used will provide facilities for "through-group" and "through-supergroup" working with other systems, e.g. a standard coaxial or radio system.

At the terminal stations, the equipment for each radio frequency channel will consist of transmitter and receiver cubicles, the transmitter being equipped with a travelling-wave amplifier which will deliver an output of 5W to the aerial system.

In the case of the repeater station, two one-way 5W repeaters will be required, each consisting of one cubicle mounting the transmitter and receiver.

Radio-Telephone Terminal Equipment

The problem of reducing the amount of radio-telephone terminal equipment required at exchange centres serving a number of out-station links, so that as little as possible remains idle for any length of time, is one that faces many overseas administrations.

Although the radio-telephone is often the only means of communication, the cost of providing separate v.h.f. links from a township to a number of surrounding settlements, is frequently prohibitive.

To meet this need, a new system known as A.S.A.C. (Automatic Selection of Any Channel) has been introduced by Automatic Telephone & Electric Co. Ltd., which more than halves the cost of terminal equipment, while at the same time providing normal exchange termination for each sub-centre, with no loss of secrecy.

Designed for use at exchange centres serving up to six v.h.f. radio links, A.S.A.C. consists of an automatic selection device which searches subscribers' links continuously by means of crystal switching, until a call is originated. The equipment can be used in conjunction with any type of exchange—either manual or automatic.

Using duplex working, separate transmitter-receiver units are normally required for each end of a link, but by employing the A.S.A.C. system only two sets are needed at an exchange to serve six outstations. This enables two calls to be handled simultaneously, or alternatively a call via the exchange between two outstations.

Once a channel is taken into use, the searching arrangement is transferred to the second equipment, the engaged channel being by-passed during searching. A forced release feature is available to prevent the system being held indefinitely by one caller during busy periods.

Short News Items

The National Institute of Agricultural Engineering held Open Days at Silsoe, Bedfordshire, on 25 and 26 July. Most of the exhibits were of a static nature but a number of machines and implements at varying stages of development were seen at work.

The British Thomson-Houston Co. Ltd recently held the third Summer School in Electrical Engineering. Approximately twenty engineers and scientists, each a specialist in his subject, lectured on the latest developments in electrical engineering research, design and manufacture. Visits to works and laboratories and a tour of the BTH works at Blackheath, Birmingham, were included.

Anglo-German Atomic Energy Agreement. An agreement between Her Majesty's Government and the Government of the Federal Republic of Germany for co-operation on the peaceful uses of atomic energy was recently signed in London. Under this agreement, the Federal Republic of Germany may obtain, on terms to be agreed, research reactors and the necessary fuel elements, and unclassified information on topics to be agreed concerning the design, construction and operation of these reactors.

Marconi's Wireless Telegraph Co Ltd have supplied nine of their Type BD.262 series of 20kW h.f. broadcasting transmitters, together with ancillary equipment, for the South African Broadcasting Corporation's broadcasting station at Paradys in the Orange Free State. Marconi engineers are assisting with final adjustments on site but most of the installation is being carried out by S.A.B.C. personnel.

Pye Electric Ltd has been formed as a company to develop and manufacture electrical equipment for the home and export markets. Preparatory work has been proceeding for some years and it is hoped that an announcement will be made within the next few months on the range of appliances to be covered.

The Convention of the British Amateur Television Club will be held at the Bonnington Hotel, Southampton Row, London, W.C.1 from 10 a.m. to 7 p.m. on Saturday 27 October. Tickets will be on sale at the door, or may be obtained in advance from Mr. D. S. Reid, 4 Bishop Road, Chelmsford. (Members 3s 6d, non-members 5s.)

The London Audio Fair 1957 will be held from 12-15 April at the Waldorf Hotel, Aldwych, London, W.C.2. The scope of the 1957 exhibition will be extended to provide facilities for a larger number of exhibitors than this year. The dates have been chosen so that the Fair follows the R.E.C.M.F. Exhibition.

The next meeting of the Electro-Physiological Technologists' Association will be held at the London Hospital on 15 September.

The Radio Industry Council announce new export records for British radio, television and electronic equipment, both for June and for the first six months of 1956. Sales for June, totalling £3·57 million were £170 000 higher than those for May, which had set up a new record. The total for the first six months of the year is now £19·17 million compared with £15·60 million for the first six months of last year.

The Guild of Air Traffic Control Officers are holding a Convention at Southend-on-Sea on 4 and 5 October. The main objective of the Convention is to bring together those who use the services provided by air traffic control, those who manufacture the concerned equipments, and those who actually operate the air traffic control services. Further details and programme can be obtained from the Clerk of the Guild of Air Traffic Control Officers, 118 Mount Street, Berkeley Square, London, W.1.

At the Annual General Meeting of the Scientific Instrument Manufacturers' Association, the retiring President, Mr. C. E. T. Cridland, installed his successor, Mr. G. A. Whipple, in the chair for the year 1956-57. Mr. Whipple is the chairman and managing director of Hilger & Watts Ltd.

Wembley Evening Institute, Copland School, High Road, Wembley, Middlesex, will be holding classes to prepare candidates for the City and Guilds Radio Amateurs' Examination. The classes will be held on Monday and Thursday evenings. Enrolment will be at the school from 10-13 September between 7 and 9 p.m. and the classes will start during the week beginning 17 September.

Battersea Polytechnic have organized two courses of lectures in Microwave Physics for the forthcoming session, on the Theory of Microwave Circuits, and Techniques and Measurements at Microwave Frequencies respectively. Enrolment forms may be obtained from the Secretary (Microwave Techniques Course), Battersea Polytechnic, London, S.W.11.

Norwood Technical College announce that the next full-time course in Radar for the Ministry of Transport Radar Maintenance Certificate will begin on 12 September and prepare for an examination to be held in early December. The next part-time course begins on 24 September and requires attendance on Monday and Tuesday evenings from 6.15 to 9.15 p.m. Two years' attendance is required on this evening course if a Ministry of Transport Maintenance Certificate is to be obtained. Details may be obtained from the Head of the Telecommunications Engineering and Radio Department, Norwood Technical College, Knight's Hill, West Norwood, London, S.E.27.

The Northern Polytechnic have issued a prospectus for the 1956-57 session. This includes full-time and part-time day and evening courses in radio telecommunications, electronics, television and radar. Details of these particular courses may be obtained from the Head of the Department of Telecommunications, The Northern Polytechnic, Holloway, London, N.7.

Erratum. The following amendments should be made in the article entitled "Electronic Methods of Analogue Multiplication" (Part 1) by Z. Czajkowski, which appeared in the July issue. They relate to the paragraph under the heading Crossed Field Multipliers on Page 286. All symbols X and Y are interchanged in the paragraph in question. Small letters x and y were meant to be used to indicate the x and y axis in the plane of the screen whereas X and Y should be used to indicate the variables to be multiplied, in agreement with the rest of the article. The paragraph in question should read as follows.

This ingenious device[15] makes use of the laws which govern the movement of the electrons in a combined electromagnetic 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 proportional to the second variable X. The magnetic field H will, therefore, produce a deflexion in the x-direction proportional to: HV_y i.e. to XY. The electron beam produces a spot on the fluorescent screen, half of which is covered with a mask with a straight edge along the y-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 x-plates. When sufficient voltage is applied to reduce the x-deflexion back to zero, this voltage will be proportional to the deflexion produced by the magnetic field, i.e. to the product $X.Y$.

The arrangement of the deflexion plates in Fig. 7 is in agreement with this description but the edge of the mask appears along the x-axis and not y-axis which is an error.

LETTERS TO THE EDITOR

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

A.C. Voltage Stabilizers

DEAR SIR,—In connexion with Dr. F. A. Benson's article on a.c. voltage stabilizers, (June 1956, issue) our experience of tungsten-filament diodes, which we use in a number of our automatic voltage regulators, may be of interest to your readers.

Some years ago the A2087 tube was selected by our electronic engineering department for use in our a.c. voltage stabilizers, after very extensive tests at different emission currents. At an emission current of approximately 100μA the valve has a calculated life of well in excess of 40 000 hours. A batch of six valves has now been on continuous life test for nearly two years without a single failure. After careful ageing the emission is a very stable function of heater voltage, and remarkably free from random variations.

In order to ensure this very high degree of stability, it is however essential to eliminate all possibilities of variable contact resistance from the filament circuit, e.g. it has been found necessary to spot-weld flying leads to the pins of the valve, and solder these directly into the circuit.

All valves are carefully tested and any showing instability due to light tapping, or excessive drift during the ageing process, are rejected. At one time, a high percentage of these valves had to be rejected.

A special tungsten filament diode has now been developed for us. Quantity deliveries show complete freedom of drift due to mechanical shock, and tests carried out during the last six months indicate a still higher long-term stability. When sufficient data is available the results of our tests will be published.

Yours faithfully,

CLAUDE L. LYONS,

Claude Lyons Ltd.,
Hoddesdon,
Herts.

The authors reply :

DEAR SIR,—We wish to thank Messrs. Claude Lyons for their interesting letter.

The variable contact resistance between valve pin and valveholder is a source of trouble with these valves which we have experienced, especially with paxolin valveholders. In this respect, the A2087 valve is at a disadvantage compared with others, since each end of the filament is brought out to only one pin, compared with three on the 29C1 and AV33 and two on the GRD6. Using the 29C1 in a moulded valveholder, the circuit we have described has given many months' service free from trouble which can be attributed to variable contact resistance at the valve pins.

Our experience of A2087 valves is rather limited but we are not surprised to hear that many samples have been considered unsuitable for this type of work. The use of a low emission current of about 100μA would certainly give the valve a long life but the response time would be increased [1]. The necessity for a careful ageing period is shown by the results of our life tests on other types of valve in which the valves have been found to exhibit a good stability only after a few months' running[2].

As we do not know the circuit of the regulators mentioned in the letter we cannot say whether the advantage of two filaments in parallel, from the safety point of view is desirable in this case. Although in our unit an excessive output voltage would be developed in the event of a complete filament failure, it should be possible to design a circuit in which the reverse would be the case, or to use a safety device which would render the unit inoperative under such circumstances.

Yours faithfully,

F. A. BENSON,

M. S. SEAMAN,

University of Sheffield.

REFERENCES

1. BENSON, F. A., SEAMAN, M. S. Some Characteristics of Saturated Diodes with A. C. Heating. To be published in *Electronic Engng.*
2. BENSON, F. A., SEAMAN, M. S. Saturated Diodes. (Letter to the Editor), *Electronic Engng.*, *28*, 84, 1956.

DEAR SIR,—We refer to the article by Benson and Seaman, which appeared in the June issue.

In paragraph two it is stated "many types of a.c. voltage stabilizers have been marketed. . . . . but they suffer from several defects". The article then goes on to describe the development of an "electronically controlled transducer circuit"—"fairly well known in principle" and "similar to that described by Helterline".

In fact, the circuit arrangements discussed are identical to the a.c. voltage stabilizer developed nearly ten years ago by the Sorensen Company of America (of which Mr. Helterline is General Manager) and which this Company, under licence, has manufactured and marketed for the past three years in this country under the trade name "Sorensen A.C. Voltage Regulator".

The design features of this stabilizer are covered by patents in America, Europe and Great Britain, and the exclusive use of the British patents are vested in this Company. These patents also cover the design features of a special diode control valve, used in all Sorensen voltage stabilizers in which failure of the tungsten filament causes a switch, integral with the electrode assembly to short-circuit the filament-to-anode path, thus completely avoiding any possibility of dangerously high voltages on the auto-transformer or over voltage at the output terminals.

Our sales and Advertising Departments are crestfallen to find that the authors were unaware that valves of this type are now made in this country and are currently available, and that stabilizers of the type described are also commercially available in ratings from 250VA to 10kVA with a regulation better than $\pm$ 0·1 per cent for a $\pm$ 10 per cent change of voltage or frequency, or from zero to full load.

However, both they and our Technical Department are extremely interested to learn that, working quite independently, the authors have come to the same conclusions as did this Company some years ago, namely that this particular form of a.c. voltage stabilizer is an extremely efficient device.

Yours faithfully,

J. LANGHAM THOMPSON,

J. Langham Thompson Ltd.,
Bushey Heath,
Herts.

The authors' reply :

DEAR SIR,—We wish to thank Mr. J. Langham Thompson for his letter. While it might be inferred from the second paragraph of our article that the unit described is new and not available on the market, the third paragraph shows clearly that the basic circuit is due to Helterline (reference 6) and that it is now covered by a British patent (reference 7), the latter entailing that Sorensen's (the holders of the patent) shall take steps to market the circuit in this country.

Unfortunately there was a long delay between the writing of the article and its appearance in *Electronic Engineering*. When the article was written we were unaware that J. Langham Thompson Ltd. were marketing a unit using this circuit, although we knew that this was so by the time the article was published.

Development work on the unit described commenced early in 1951 with the intention of producing apparatus for general laboratory use, and apart from the saturable core reactor (which in our unit uses commercially-available transformer laminations) it can be built up from normally-available components. Although only the basic circuit had been published previously we felt it desirable at the time to publish the data we had accumulated relating to the many variable circuit parameters involved, such as choice of valves, bridge ratio, auto-transformer ratio, reactor flux density and other component values, and also the performances we had obtained. We do not, however, claim that the performance is the best that the circuit can give since certain developments would enable an improvement to be obtained.

The special valves mentioned were not known to be available when the work described was carried out. Attree and ourselves, some three years ago, carried out a comprehensive survey of valves then available which were suitable for this application; the characteristics of these valves have been reported elsewhere (references 8, 9 and 11). During the course of these investigations we found that the A2087 valve had been developed to replace the A1468 which is now obsolete and was therefore not included in the above-mentioned survey.

Yours faithfully,

F. A. BENSON,
M. S. SEAMAN,
University of Sheffield.

A Relay-Operated Serial Adder

DEAR SIR.—It was with great interest that I read J. K. Wood's article in the June issue. It might interest your readers to compare this with a relay-operated parallel adder developed in our laboratories.

The adder was designed as a special purpose instrument in conjunction with data read-out systems. The main requirements to be met were:—

(a) Input to the adder was from an analogue-to-digital convertor. The input was in parallel form and it was available for only 50msec.

(b) The adder had to operate in a binary-decimal code. This notation retains the decimal decades but every decimal digit is coded in binary form. For instance, the decimal number 937 is coded as 1001, 0011, 0111.

The time available for the adding process being short, a parallel adder seemed most suitable. The input from the digitizer was stored in the input store (relay group A) and the already accumulated sum was stored in the main store (relay group B), Fig. 1. Binary 1 and 0 were represented by energized and de-energized relays, respectively.

The contact pattern of the two groups of relays, A and B, comprises the actual adding circuit. Fig. 2 shows a small section of the network, representing the adding circuit for one binary digit or digitizer and accumulator input. Whenever there is a carry from the next less significant digit, line 2 is earthed, otherwise the earth is on line 1.

The operation of the adding circuit is best explained by an example. Consider the case when no carry digit is present and both A and B are energized. This corresponds to the addition of binary 1 + 1. There being no carry, line 1 is earthed, but since contacts A_1 and B_1 are both operated, point C is disconnected from earth and relay C will not energize, representing: sum = 0. Simultaneously, contacts A_3 and B_3 transfer the earth from line 1 to the carry line of the digit of higher significance. (Blocking rectifiers MR_1 and MR_2 ensure that no current can flow from A to B or vice-versa). This circuit is repeated four times for every decade.

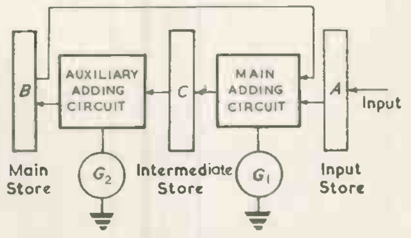


Fig. 1. Block diagram of parallel adder for operation in binary-decimal code

Unfortunately, the output from the adding circuit does not always give the sum in a correct binary-decimal notation. For instance, if the sum is 12 it will appear as 8 + 4 (1100) in the units decade, instead of 2 plus a carry digit to the tens decade. In order to achieve the correct notation, it is customary to use a code generally known as the "excess 3 code". However, in the present case it was decided to retain the ordinary code and to perform a second addition. Whenever the sum exceeded 9 a carry digit was carried over to the decade of higher significance and simultaneously a 6 was added to its own decade. (It can easily be verified that this system gives the correct answer. Thus in the above example 12 + 6 = 18 $= 2^4 + 2$. Neglecting the fifth binary digit, 2^4, the answer is 2, as required).

The programming of the adder is as follows: (Fig. 1)—The new input is stored in the self-locking relays of group A. The already accumulated sum is stored in B. When gate G_1 is opened, an earth is applied to the adding circuit and the self-locking relays of group C are set up according to the contact pattern of relay groups A and B, giving the sum A plus B. Gate G_1 is then closed, admitting no further input into C and the main store

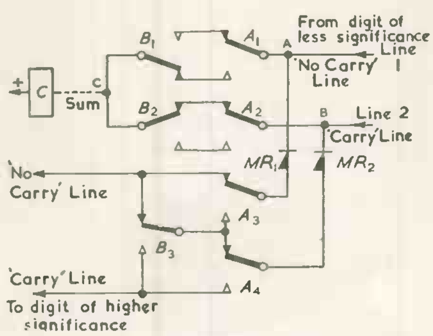


Fig. 2. Section of adding circuit. C = A + B

B is cleared. Gate G_2 is opened, earthing the auxiliary adding circuit and transferring the sum into the main store. (It is this auxiliary circuit which adds 6 to the sum when necessary). To end the cycle of operations, gate G_2 is closed again and stores A and C are cleared. The circuit is ready for the next operation.

J. K. Wood mentions correctly in his article that a relay adder based on parallel operation requires a large number of relays. However, when adding numbers in coded notation it seems to be the most straightforward method. Also, the operation is fast and the programming extremely simple. Moreover, for the described application, most of the relays would have been required in any case for storage purposes.

Yours faithfully

L. J. BENTAL
Elliott Brothers (London) Ltd.,
London, S.E.13.

The author replies:

DEAR SIR.—I was very interested to hear about L. J. Bental's design, especially his method of restoring the correct coding after addition. It is true that a parallel adder is more straightforward when dealing with numbers in coded notation, and as he requires an input store for his application the extra relays required to make an intermediate store for a

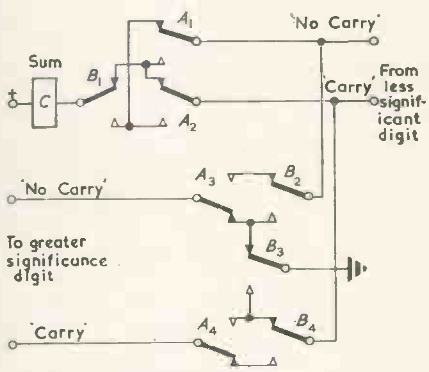


Fig. A. Relay circuit without rectifiers

parallel adder may be even better than the inclusion of a uniselector for serialization, since the operating cycle is also much shorter.

However, with the type of circuit he describes there is a possible disadvantage arising from the use of blocking rectifiers. The current for the intermediate store relays may have to flow through one or more rectifiers in series. The maximum number of rectifiers in series when using binary coded decimal notation is three, which is satisfactory, but for a straight binary parallel adder (say 25 digits), the current to the most significant relay may have to flow through 24 rectifiers in series—one less than the total number of digits.

A circuit developed by the author eliminates the need for rectifiers at the expense of one extra changeover, resulting in more economical design. The circuit is given in Fig. A. Operation is straightforward, there being one changeover less in the sum circuit and two more in the carry circuit.

Yours faithfully,

J. K. WOOD,
United Kingdom Atomic Energy
Authority, Harwell.

BOOK REVIEWS

High Fidelity: The Why and How for Amateurs

By G. A. Briggs. 188 pp., 50 figs. Demy 8vo. Wharfedale Wireless Works Ltd. 1956. Price 12s. 6d.

THE irrepressible Mr. Briggs is in the fortunate position of being able to publish what he cares to pay for. Whether you buy this book or not depends on whether you have Mr. Briggs' other books, for although there is additional material here there is a remarkable similarity between some of the subject matter and his prior publications. It must therefore be concluded that there is little more to be said about high fidelity reproduction in the Briggsian idiom.

The professional engineer has nothing to learn from this book, which is described as non-technical; there are of course many points on which further information would be desirable, and it is unfortunate that the author has not given more references to the many publications dealing so much more explicitly with these matters. In his conclusion Mr. Briggs says that the various chapters have not all taken on the shape first visualized, and that some of them seem abruptly phrased, but he does not explain why. There is however what appears to be a good deal of "padding" in relation to the useful content of the various chapters. As an example, it is difficult to see what contribution to high fidelity listening could be made by the old loudspeakers illustrated, or why such things as a photograph of a 32ft pitch organ pipe should be included. Since Mr. Briggs has calculated the cost of the book to be 9d per word and professes anxiety to give the reader the best value possible, opinion will be divided on the proportions of useful as against interesting information given.

The author is of course no stickler for the purists and is inclined to be dogmatic on what are well known to be controversial matters; but it is recognized that in a relatively non-technical book dealing largely with the interpretation of individual tastes, the definite views of the writer are perhaps better than a negative attitude.

To comment on every chapter would not be possible, but it is a pity that the work of Dr. Tombs on the corona discharge loudspeaker was not given as much prominence as the Klein system mentioned. The author is well known as an advocate of sand-filled cabinets, evidently inspired by a visit to the BBC in 1938; but in fact sand-filling was standard practice in film recording studios from 1928-1930 to deaden camera noise, the mobile booths having hollow walls filled with sand. Mr. Briggs is "astounded" that the Siemens-Halske ribbon loudspeaker was used to reproduce speech at considerable power in the open air, because such ribbons are now limited to frequencies of 1kc/s and above; these large multiple ribbons were standard equipment in many German cinemas and this reviewer can assure Mr. Briggs, from personal experience of the system in 1931, that it was entirely satisfactory down to 200c/s and below. The description of Sir Oliver Lodge's moving coil loudspeaker suggests that it was developed about the same time as the Rice and Kellog patents, but in fact it was first made in 1898. And so on; the information in this book is tantalizingly incomplete to the engineer, and as this is an engineering journal attention must be drawn to this defect.

However, for the man in the street who wishes rapidly to skim the surface of the technique of high fidelity reproduction methods, the book is excellent value for the money as present book prices go. The printing, paper and illustrations are good and well laid out, although in this particular copy the binding was poor and split in half after about a dozen openings. The subject matter is quite up to date and the book abounds in useful controversial hints and tips. A chapter is devoted to answering questions, and the writing is enlivened throughout by the author's characteristic touches of humour and some apt cartoons.

Since there is no such thing as a definition of high fidelity reproduction—unless it be the depth of one's purse—Mr. Briggs is assured of satisfactory sales in advance, if only because no aspect of electronics has provoked so much curiosity as to what the "other man" does as has "hi-fi".

ALAN DOUGLAS.

Der Transistor Ein neues Verstärkerelement (The Transistor. A New Amplifier Element)

By Joachim Dosse. 109 pp., 44 figs. Demy 8vo. R. Oldenbourg, München. 1955. Price DM 11·80.

THE author's aim in writing this slender volume was to give as easily understandable as possible an introduction to transistor technique. It is perhaps a little doubtful whether he has achieved his aim completely. But this is not astonishing considering the fact that even a specialist in this field has recently stated that "the crystal valve is as yet not properly understood" and "it is not possible to cover semiconductor phenomena adequately without excursions into more advanced physics" (T. R. Scott. *Transistors and other Crystal Valves* p. 203). The present author explains the properties of semiconductors by the use of a simplified model and merely corpuscular concepts and refers to Shockley's book and to E. Spenke's *Electronic Semiconductors* (see the review by Dr. Hoselitz, September 1955 issue of ELECTRONIC ENGINEERING, p. 415) for a treatment using wave mechanics and the recent results of the theory of solids.

The five chapters of the book deal with the historical development, the mode of action, the forms of development, the technical properties and the various circuits. The treatment is non-mathematical except for the mention of a "hybrid"-matrix connecting input and output currents and voltages. A sound knowledge of conventional electronic valves, their properties and circuits, is useful for the understanding of the book.

Welcome features are the multi-coloured figures and the colour plates illustrating the interior structure of point and junction transistors as well as the drawing of a germanium crystal, the various figures and a table comparing the electrical and operational properties of transistors and valves and the delimitation of their respective fields of application. A bibliography not claiming to be complete but sufficiently comprehensive, covering the literature till July 1955 and arranged according to the five chapters, and a subject matter index, complete this well produced booklet which can be recommended to all those wishing to get an idea of the new technique and having only limited time at their disposal.

R. NEUMANN.

Solid State Physics—Advances in Research and Applications. Vol. 1.

Edited by F. Seitz and D. Turnbull. 469 pp., 50 figs. Academic Press Inc. New York. 1955. Price $10.00.

THIS volume is the first of a series which is planned to provide compact reviews of important aspects of Solid State Physics to meet the needs of research workers and students in this and closely related fields. With this objective in mind, three types of reviews are solicited for the series, viz. (1) broad elementary surveys, (2) broad surveys of fields of advanced research and (3) more specialized articles describing new experimental and theoretical techniques. There can be little doubt about the need for such reviews and, to judge by this first volume, the intended series is going to be extremely valuable.

The emphasis in this first volume is on problems in theoretical physics, the one exception being a general survey by H. Y. Fan of the valence semiconductors germanium and silicon. Besides being a good introduction to the subject, this should serve as a useful reference article to those working in this field. Similar comments apply to the article by J. R. Reitz on the methods of the one-electron theory of solids. This describes the basic principles of the method and also, the various procedures which have been proposed for the calculation of the allowed energy levels of electrons in a solid. A more specialized article by F. S. Ham describes more fully the quantum-defect method—a recent development of the one-electron theory which, in certain cases, eliminates the necessity for a knowledge of the ion-core potential, this being essential for the other methods of calculation. In the quantum-defect method, equivalent information is provided by the spectroscopic term values of the free atoms.

Three further specialized articles complete this volume. A qualitative description of the cohesion of metals is given by E. P. Wigner and F. Seitz, on the basis of the one-electron theory. The subject of electron interaction in metals, which, for a long time, has been one of the more serious defects of the one-electron theory, is treated by D. Pines. This author discusses the collective electron description of metals which has been developed by him in collaboration with D. Bohm, and which takes these interactions into account from the beginning of the theory. Finally, there is a description of the theory of order-disorder transitions in alloys by T. Muto and Y. Takazi.

The quality of production is high but one would have thought that it would have been more useful (and more profitable) to restrict the articles within a particular volume to be of the same type. However, provided a prospective purchaser is reconciled to the fact that he may not read all the articles, this book can be highly recommended.

D. R. DRABBLE.

Second Thoughts on Radio Theory

By "Cathode Ray." 69 pp., 40 figs. Demy 8vo. Iliffe & Sons Ltd. 1955. Price 25s.

THIS is not a text book—it is really a collection of essays on the fundamentals of radio practice—but every student should unquestionably have a copy on his desk and very nearly every practicing radio engineer should have a copy by his bedside.

There are many different ways of saying the same thing and, far too frequently, the writers of text-books employ a coldly impersonal style which is difficult to absorb except in small doses. "Cathode-Ray" never forgets that his readers are human beings and in his light-hearted, almost conversational, manner, he makes even abstruse ideas easy to understand and the mathematics both intelligible and interesting. What is more, it is fun to read.

The forty-four essays which the book contains have nearly all appeared in the pages of the *Wireless World* during the past ten years or so although many have now been re-written. They deal with many of the basic principles of radio practice from Ohm's Law to Energy ("most people think energy is the thing they haven't got on Monday mornings"), from Waves to Phase, Modulation and Beats. Subsequent essays take the reader light-heartedly but thoroughly through the complexities of Tuned Circuits, "*Q*", Negative Feedback and Filter Design while a further series, rather more mathematical in nature, serve to introduce the student to the ins and outs of decibels, the calculus, "j", Thevenin's Theorem and Kirchhoff's Laws. If the reader be already acquainted with these conceptions, the book will serve to renew old friendships in a new and refreshing light.

"Cathode-Ray" is one of the most able writers on the technicalities of radio practice. His essays are each a masterpiece of clear exposition and they are written with a blend of humour and understanding which greatly assists in the digestion of the technical "meat".

G. R. M. GARRATT.

The Theory of Sound

By Lord Rayleigh. 480 pp. and 491 pp. (2 volumes). Demy 8vo. 2nd edition revised. Dover Publications, Inc. New York. 1956. Price $1.95 per volume.

THIS is the revised edition of the first American print which appeared in 1945. Originally published in 1877, this treatise still retains a position of importance in the literature of its field, when most scientific treatises of the period are for the most part of historical interest only.

The introduction to Volume 1 gives a biographical sketch of Lord Rayleigh, a note on the historical development of acoustics to the time of Rayleigh and his contributions to acoustics and their significance for modern developments.

Color Television Receiver Practices

Edited by C. E. Dean. 200 pp., 70 figs. Demy 8vo. John F. Rider Inc. New York. Chapman & Hall Ltd., London. 1956. Price 36s.

THE book begins with a broad statement of the fundamental requirements of a colour television system, followed by a description of the standard transmitted signal. After this the various parts of the receiver are discussed in detail. The final chapter describes laboratory equipment for colour television work.

The work is based on a course of lectures given by the Hazeltine Corporation (U.S.A.) for visiting engineers from various manufacturing companies.

Wireless Servicing Manual

By W. T. Cocking. 268 pp., 128 figs. Demy 8vo. 9th Edition. Iliffe & Sons Ltd. 1956. Price 17s. 6d.

SINCE 1936 this book has been known to radio servicemen everywhere as a thorough and comprehensive guide to solving most of the problems that arise in the repair, maintenance and adjustment of the modern receiver. It has been revised to take account of all recent developments in receiving equipment.

This edition appears in a larger and handier format. All illustrations have been modernized and brought into line with current symbolism.

ELECTRONIC EQUIPMENT

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

TRANSISTOR AMPLIFIER

(Illustrated below)

Microwave Instruments Ltd, West Chirton Industrial Estate, North Shields, Northumberland

Type T.A.2 is a small portable combined amplifier and indicating instrument completely free of mains supply, for use in measuring modulated weak signals such as may be encountered in radiation field measurements and microwave test bench assemblies.

The instrument is robust and portable, has self-contained battery supplies with extremely long life and therefore is free from errors which may normally be attributable to mains voltage changes. The controls, grouped on the front panel are:

(a) Continuously variable gain control.
(b) Switched gain control.
(c) Function switch.
(d) Selective/wide-band.

Sensitivity:
Wideband 15μV r.m.s., f.s.d.
Selective 20μV r.m.s., f.s.d.

Frequency response:
(to half deflexion).

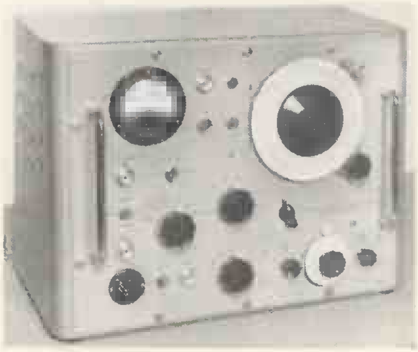
Wideband 100c/s to 10kc/s.
Selective 3·2kc/s ±100c/s.
(Q=30 approx.)

Input impedance: 900Ω ±20 per cent.
Input range: Less than 1μV r.m.s. to 0·3V r.m.s. 100μA d.c. max.
Meter scale: 0 to 100 linear.
Battery drain: 3mA (approx.).

F.M. SIGNAL GENERATOR

(Illustrated above right)

Hatfield Instruments Ltd, Crawley Road, Horsham, Sussex



The type LE.250 signal generator is an a.c. mains operated instrument intended for measurement on f.m. receivers operating in the frequency range 81 to 105Mc/s. It has an additional frequency range of 3 to 20Mc/s for alignment of i.f. circuits. Provision is made for the display of selectivity and discriminator curves in conjunction with a suitable oscilloscope which need not have a time-base. It incorporates the recently developed r.f. attenuators type A 90. The source impedance of the attenuator system is 75Ω and the r.f. open-circuit output voltage can be varied from 2μV to 200mV in 1dB steps. Where the signal generator is connected to a 75Ω load through a length of 75Ω cable the r.f. output will be halved, i.e. 1μV to 100mV in 1dB steps. The r.f. input to the attenuators is monitored by a crystal voltmeter.

The signal generator incorporates two modulation systems, one a sine wave at 400c/s and the other a sawtooth at 25c/s. By means of a six position switch the r.f. carrier can be modulated in various ways and on two switch positions external amplitude modulation can be applied thus giving combined f.m. and a.m.

Both sine and sawtooth oscillators are amplitude stabilized and are independent of mains fluctuations.

THREE-PHASE TRANSFORMERS

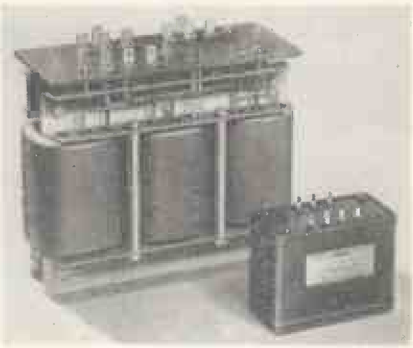
(Illustrated below)

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

These transformers have been designed to meet the conditions set out in R.C.S.C. Specification No. R.C.S. 214 and comply with the recommended dimensional requirements.

The Parmeko three-phase series embraces the full range of "E" type cores extending from the miniature types to those capable of handlings up to 2·5kVA at 50c/s or correspondingly greater ratings at higher frequencies.

The smaller models are totally enclosed and hermetically sealed whereas the larger models, for economic and weight considerations utilize an open-type construction normally impregnated to R.C.S.C. standards.



The design incorporates the "E" type core manufactured from cold rolled grain oriented silicon steel, a development from the now familiar "C" core and has the advantage that it can be operated at a very much higher flux density than the conventional laminated core. Greater efficiency is thereby achieved with reduction in volume and weight.

FLAMEPROOF LOUDSPEAKER

(Illustrated below)

The General Electric Co. Ltd, Magnet House, Kingsway, W.C.2

This unit has received the Ministry of Fuel and Power's certificate approving its use in atmospheres covered by Groups 2 and 3 of B.S.229/1946 dealing with flameproof equipment. It can therefore be used to provide intelligible speech in those sections of the oil, paint, and

chemical industries, for example, where direct sound broadcasting has hitherto been impossible because of the risk of fire or explosion.

The driving unit is a pressure operated device suitable for use with a projector horn or re-entrant horn having a minimum air column length of 20in. The speech coil and driver section, magnet assembly and line matching transformer are totally enclosed in a cast aluminium alloy case.

Enamelled copper wire is used on the speech coil winding, with beryllium copper lead-out strips. The multi-ratio transformer provides various line impedances between 600Ω and 30kΩ and the power fed from the loudspeaker line to any unit may thus be individually adjusted to suit varying site conditions. The maximum power handling capacity is 10W and the frequency response 200c/s to 7kc/s ± 3dB.

H.F. TRANSFORMERS

Balun Ltd, 175 Uxbridge Road, Hanwell, W.7

These transformers are small and compact, hermetically sealed, completely shielded and fully tropical, with a frequency range the order of 1 000:1 at frequencies up to 100Mc/s and in the higher frequency models operating at say 400Mc/s the frequency range is of the order of 100:1.

They are particularly suitable for matching transmission lines to antenna systems and one particular model is designed for matching Rhombic antennae to coaxial lines over the frequency band 200kc/s to 100Mc/s.

The transformers can be used in modulators, mixing units, wideband amplifiers, hybrid units, repeaters etc. and particularly for matching balanced to unbalanced systems. The accuracy of balance on centre tapped windings varies with frequency and falls as the frequency rises.

The attenuation of push-push currents is never less than 20dB. The centre point of balanced windings is left floating or rigidly grounded according to requirements.

INDICATING TEMPERATURE CONTROLLER

(Illustrated below)

Ether Limited, 60 Tyburn Road, Erdington, Birmingham 24

"TRANSITROL" Type 990 is a new self-contained direct-deflexion instrument for indicating and controlling tem-

perature to close accuracy over a wide range. It can also be applied to any process where the signal can be converted into a direct current or voltage.

This instrument incorporates a conventional galvanometer as the measuring system, and an indicating pointer. The latter operates a simple photo-electric system which, in turn controls the heating medium.

The control-arm, adjusted externally to the required temperature set-point, as indicated by a red index pointer, is fitted with photo-sensitive transistor, optical-lens and low-level light-source. When the transistor receives full illumination a relay is energized, giving external switching control. The temperature-indicating pointer is fitted with a flag, capable of passing freely between light-source and transistor and when the desired temperature is attained, the light-source is cut off from the transistor, thus de-energizing the control-relay and shutting off the heating medium. The design of the optical system ensures that only a minute movement of the pointer is required to make and break the control relay thus giving a fine degree of control.

The Transitrol is fully compensated for cold junction variation and internal resistance changes due to the temperature-resistance coefficient of the copper moving coil. To permit final calibration on installation, according to the specific length of the compensating load used, an adjustable rheostat is provided. Calibrations are available for use with all types of thermo-couple or other temperature-sensing media. The entire instrument is designed as a plug-in unit and is easily withdrawn from its casing. It is suitable for wall or panel mounting.

UNIVERSAL MEASURING INSTRUMENT

(Illustrated below)

Philips Electrical Ltd, Century House, Shaftesbury Avenue, London, W.C.2

THE GM.6008 is a versatile and compact instrument which can be used for measuring a.c. or d.c. voltage and current, resistance and capacitance. Measurements, except those of direct voltage and current, are made electronically, achieving, it is claimed, greater accuracy than is possible with conventional moving-coil instruments.

The instrument has the following ranges:

Direct voltage: 20mV to 1kV. (An h.t. probe can be supplied for 1 to 30kV).
Alternating voltage: 100mV to 300V between 20c/s and 100Mc/s. (A diode probe can be supplied for frequencies above 100Mc/s).
Direct current: 10μA to 1A
Alternating current: 0·1μA to 10mA between 20 and 1 000c/s; 10μA to 1A between 20c/s and 100kc/s
Resistance: 1Ω to 1k$M\Omega$
Capacitance: 30pF to 3μF

A switch is provided to reverse polarity when measuring positive and negative voltages in succession and voltages of 1V and 30V are built in for calibration purposes.

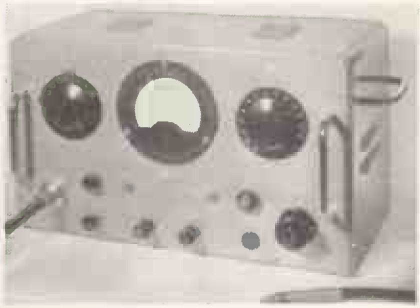
PROXIMITY METER

(Illustrated above right)

Fielden Electronics Ltd, Manchester 22

TYPE P.M.2 is an instrument for micro-measurement of a wide range of physical variables and has been applied to such problems as concentricity measurements of ball races, measurement of magneto striction, measurements of carbon deposit and the evaluation of the eccentricity of revolving shafts.

The application of the P.M.2 depends upon the measurement of the capacitance change which in most applications is



brought about by small mechanical displacements. By suitable arrangement of the electrode it is possible to apply this method to almost any problem of static and dynamic measurement. It has been successfully used as a strain gauge, surface gauge, vibration meter, torque meter, paint thickness gauge, micro-thermometer, pressure gauge and dielectric comparator.

This use of capacitance change allows a sensitivity in measurement far greater than the limits practically imposed by the mechanical stability of the measuring arrangements to which it is applied. For example, in the field of mechanical displacement a movement of less than 10μin may be detected.

As there is no physical contact with the specimen under observation in this measuring method inaccuracies due to contact pressure and wear are eliminated.

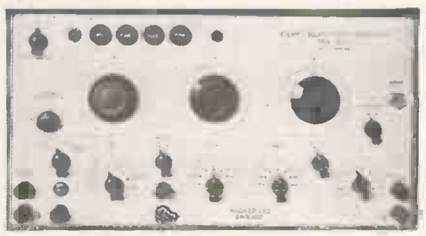
WIDE-RANGE DELAYED SQUARE PULSE GENERATOR

(Illustrated below)

Nagard Ltd, Avenue Road, Belmont, Surrey

THIS instrument was designed to meet the need for a generator of fast-rising pulses and square waves, while at the same time providing a wide range of amplitude, repetition rate, width and delay with direct-reading calibrations, together with full flexiblity as to polarities of main pulse, pre-pulse and trigger pulse and the ability to trigger from any form of external signal.

The fast rise-time, (50mV to 2V negative in less than 10mμsec) in conjunction



with the above features, enables searching investigations to be made. This rise-time necessitates the use of a special high frequency connecting cable, since the output contains important frequency components of several hundred megacycles per second. This cable is supplied with the instrument, together with suitable terminating arrangements.

Notes from ______

NORTH AMERICA

Broadcast Transmission Systems Symposium

The sixth annual symposium of the Professional Group on Broadcast Transmission Systems will be held in Pittsburgh from 14-15 September. The four technical sessions will be held in the Mellon Institute Auditorium and will feature speakers on Television Measurments, Television Studio Development and Broadcast Facilities and Operation.

Department of Defense Announcements

The Working Group on Tube Techniques of the Advisory Group on Electron Tubes will sponsor the third National Conference on Tube Techniques, to be held in New York from 12-14 September. The programme will cover all phases in the manufacture of a large number of electron devices.

The Working Group on Semiconductor Devices will sponsor a Transistor Reliability Symposium, to be held in New York from 17-18 September.

The Speakerphone

Thermistors and varistors have been employed by Bell Telephone Laboratories in the circuits of the Speakerphone, a loudspeaker-microphone combination which allows the user to have both hands free during a telephone conversation. A thermistor serves as an output limiter, and a varistor is employed in a balancing network to maintain line impedance reasonably constant for various loop lengths.

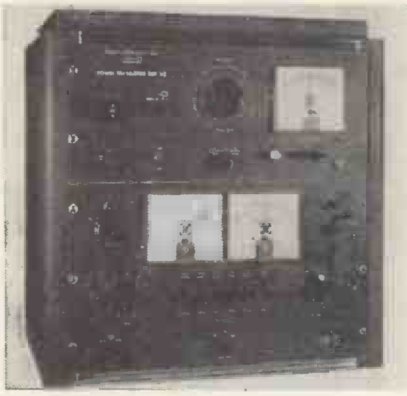
To place a call, it is only necessary to momentarily press the "on" button which lights up when the set is connected to the line. Then, when dialling tone is heard from the loudspeaker, the number is dialed in the normal manner. When answering a call, the "on" button is momentarily pressed, again causing it to light up, and the conversation can start. The transmitter picks up the voice of the user, and the other person is heard on the loudspeaker. The volume of the incoming speech is adjusted by turning the small knob of the volume control. At the end of the call, the "off" button is pressed, disconnecting the set from the line.

If privacy is desired during a call with the Speakerphone, or if some transmission difficulty is encountered, lifting the handset automatically transfers the call to the handset, and turns off the Speakerphone. If the handset is being used, and it is desired to transfer to the Speakerphone, the "on" button is depressed while the handset is being returned to the cradle. The Speakerphone is designed for a talking distance of 10 to 30in. The loudspeaker is placed about 3ft from the transmitter and so located that the normal seated position of the user is midway between the two instruments.

Power Transistor Test Set

(Illustrated below)

A new power transistor test set has been announced by Baird Associates, Atomic Instrument Company, Cambridge, Mass. Model KP1 is designed to measure all the hybrid parameters, including I_{co} in either grounded base or grounded emitter configurations, over the frequency range 100c/s to 200kc/s. I_{co} is indicated on the front panel meter over the range 1μA to 5mA, an external jack being provided to measure I_{co} below 1μA.



The instrument can reverse either emitter current or collector voltage, or both, for observing reverse d.c. characteristics of the transistor under test, the d.c. collector ranges being 0 to 100V and 0 to 300 mA and the d.c. emitter range 0 to 300 mA.

Canadian I.R.E. Convention

The Institute of Radio Engineers in Canada will hold its convention-exposition in Toronto from 1-3 October.

Over 90 leading manufacturers have booked space to exhibit their latest in radar, radio, television, control mechanisms, computers and many related products, including transistors. Peacetime atomic applications will also be shown.

Electronic devices will predominate and since radio, radar and electronics and many aspects of nuclear science overlap, the exposition will appeal to technicians, engineers and scientists, to specialists in medicine, physics and agriculture, and to business and efficiency experts. Government booths at the exposition will include one for the National Research Council of Canada and another for Atomic Energy of Canada Ltd, a Crown company whose display will feature the nuclear project shown at Geneva last year.

The large number of technical papers to be presented at the convention, to run concurrently with the exhibition, will cover such subjects as medical electronics (for diagnoses and treatments), scatter propagation, the application of electronics to atomic energy projects, the use of computers in automation and engineering problems and transistors.

U.H.F. Forward Scatter Transmitter

The Canadian Marconi Company has developed and produced an ultra high frequency forward scatter transmitter, which is broadband and capable of carrying 72 channels.

The power amplifiers consist of power supplies, control and protective circuits and cooling system, and can be equipped with any one of a family of klystrons to provide an output of 10kW c.w. or 1kW at either 700-1 000 or 1 700-2 400Mc/s. Coverage down to 400Mc/s can also be provided by using a family of 3 cavity klystrons.

The amplifier assembly is housed in a walk-in style cabinet, the design of which combines front access to all components; roof-mounted air, water and power connexions to enable back-to-back or adjacent wall location. The klystron is dolly-mounted to facilitate handling.

A Canadian Marconi experimental tropospheric scatter link was installed in July this year between Montreal and Ottawa, which will operate initially in the 2 000Mc/s band. 1kW u.h.f. terminals and large antennae will be used to scatter signals for satisfactory reception over a distance of approximately 86 miles. Later, experiments will be carried out on the 900Mc/s band during which teletype and facsimile traffic will be passed and a rigid check kept on the equipment's performance.

In April an optical microwave link was manufactured and installed by the Canadian Marconi Company for the Canadian Overseas Telecommunication Corporation. Operating between the Corporation's stations at Drummondville and Yamachiche, this system uses the 900Mc/s frequency range, the radio equipment is duplicated and automatic changeover and fault alarm facilities are provided.

Transistorized Magnetic Core Memory

Transistors are under intensive investigation at Bell Telephone Laboratories to determine their suitability in such applications as airborne electronic computers. A current phase of this investigation is the study of a large coincident-current magnetic core memory which is operated entirely by transistors and experience to date would indicate that a transistor-driven memory of this kind is entirely feasible and quite attractive.

The memory system includes the memory proper, or storage array, magnetic core switches for selecting the desired memory locations and transistor amplifiers.

The complete memory can store 1 024 eighteen bit numbers. To accomplish this, 18 432 memory cores and 48 switch cores are employed. Transistor complement includes 98 low-level and 62 high-level units. Total power consumption is less than 50 watts.