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Commentary

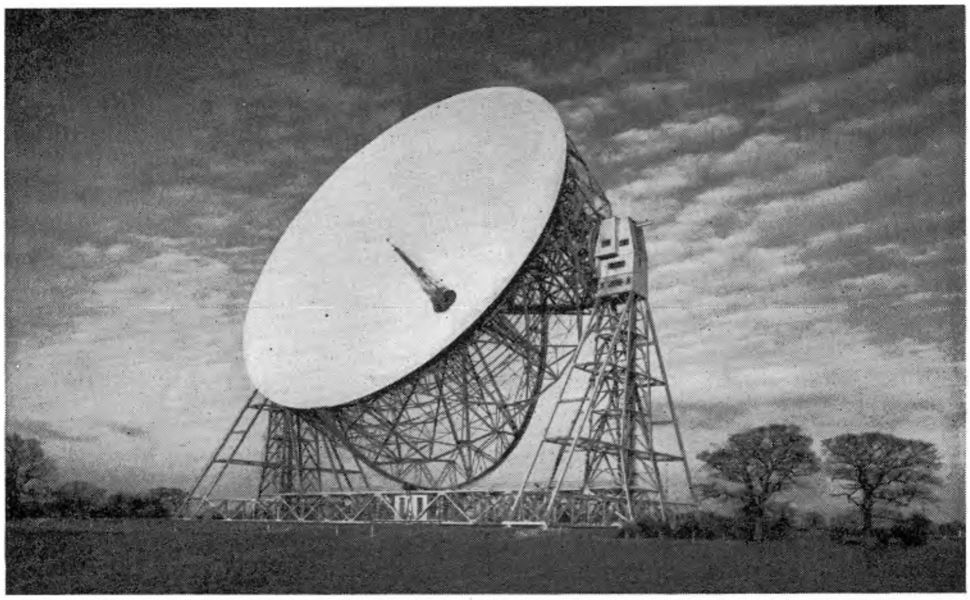
STANDARD Telephone and Cables Limited have recently celebrated their seventy-fifth anniversary, for it was on 2 May 1883 that J. E. Kingsbury, who had made a special study of the telephone and who, the year previously, had written the first handbook on telephone practice ever to be published, opened an office in Moorgate Street, London, to handle the telephone business of the Western Electric Company in Great Britain. From that modest import agency, begun with a staff of three only seven years after the invention of Bell's telephone, there has grown up one of the largest telecommunication engineering firms in the world, employing some 25 000 men and women at eight major factories in England and Wales and making a range of products extending over the whole field of telecommunications.

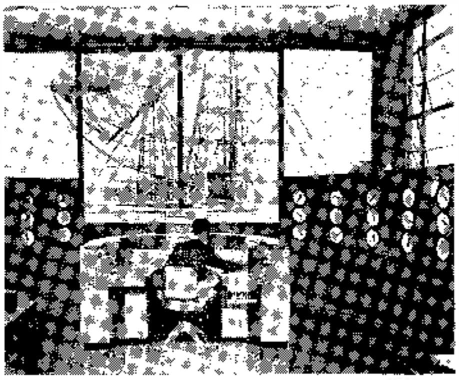
At the time of the foundation of S.T.C. the telephone, described by one historian as the "most beautiful and most useful application of science which the nineteenth century produced", was hardly out of the experimental stage. During the succeeding three-quarters of a century S.T.C., among others, have helped to give it status and prestige and helped to make it the ready and able servant of millions of people throughout the world. Today one can lift the telephone and talk to a relative or business associate a country, an ocean or a continent away, in less time than it takes to walk to the end of the road; but modern society, reared on a daily diet of scientific wonders, no longer marvels at the magic of the telephone. It takes for granted the fact that the telephone industry is equal to any problem of communication, no matter how difficult. Indeed, this blasé acceptance by the public of the wonders of the telephone is, in itself, a measure of the industry's success. The creation of such a climate of opinion has not been accomplished, however, overnight; it has been a long process made possible only by the vision, the technical skill and the craftsmanship of the pioneers and their successors.

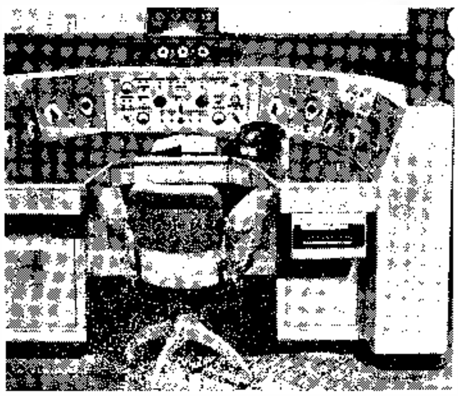
The ever present urge for man to communicate, which gave birth to the telephone, has inevitably driven him beyond the frontiers of telephony into realms about which the early pioneers could hardly have dreamt. Radio, television, radar, have all followed, directly or indirectly, from the telephone art and now the wheel has, so to speak, turned full cycle and the new era of electronic engineering is opening up new and exciting possibilities for the telecommunication engineer.

S.T.C. in its seventy-fifth anniversary year may be justly proud of the part that it has played in this development process for, mirrored in its achievements over three-quarters of a century can be seen the progress of a whole industry. From its earliest days the Company has been equipped for the development and manufacture of telephone apparatus and cable and the record of achievements which has arisen from its small beginnings is, indeed, considerable. The Company has been responsible for numerous major telephone exchanges, thousands of miles of telephone cable, a number of the world's most powerful broadcasting and radio communication transmitters and radio and line transmission links installed in all parts of the world. Their installation of telephone exchanges has ranged from the early manual switchboards such as the first Central Battery Telephone switchboard at Bristol to the latest type of board such as that installed at London Airport, which is probably the busiest telephone exchange in Britain. In addition to being one of the main suppliers of telephone equipment to the British Post Office, S.T.C. have the distinction of being the sole supplier, since 1898, of telephone equipment to the Corporation of Kingston-upon-Hull who run the only privately owned public telephone service in Britain.

The progress during these last seventy-five years has certainly been remarkable for, not only is the telephone taken for granted, but so is radio and television; indeed, a great many of the engineers now working on these projects can hardly visualize life without such amenities and, while it is reasonable that this should be so it is as well to take an occasional look back and to realize what has been achieved through the enthusiasm and foresight of our forebears. It should serve to put our present achievements in true perspective and to give courage and enthusiasm to the would be pioneers of today.







The Computer and Control for the Telescope at Jodrell Bank

In use the radio telescope is required to carry out relatively complex motions. In this article the method of computing and controlling these movements is described.

THE 250ft diameter radio telescope built for the University of Manchester at Jodrell Bank, Cheshire, weighs approximately 2 000 tons of which some 700 tons comprise the bowl and bowl structure.

The bowl or mirror is parabolical in shape and its aperture is 250ft in diameter; the focus lies in the aperture plane. The bowl surface is constructed of sheet steel welded plates and the aerial is carried on a tower 60ft high. The bowl is pivoted on bearings at the top of two towers some 180ft high and is rotated by four 50 h.p. motors (2 to each pivot) through circular racks which were taken from gun mountings on the battleship 'Royal Sovereign'.

The two towers are each mounted on 4 bogies running on double tracks laid on concrete, the bogies being driven by four 50 h.p. motors.

A central pivot is provided about which the structure revolves in azimuth and all power and signal cables are taken up through the pivot. Immediately above the pivot is the motor room housing the Ward-Leonard sets and motor control gear.

Suspended immediately below the centre of the bowl is a swinging laboratory housing pre-amplifying equipment, etc.

The basic movements are in azimuth—about a vertical axis—and, elevation—about a horizontal axis, the movements being independent; the azimuth movement is restricted to a total of 420° from Az—90° to Az + 330°, and the elevation to 400°, 20° either side of vertically downwards.

During operation of the telescope the aerial is, of course, always above the horizon, provision for tilting downwards being made for maintenance purposes and for the changing of aerials and aerial tubes.

Automatic stops are provided to prevent the telescope from moving beyond the pre-set limits on either co-ordinate.

The automatic control equipment is housed in a control room in the main building some 100yd from the telescope.

The telescope is remotely controlled and has the following basic facilities:—

(a) It is capable of being locked in any given azimuth and elevation.

(b) It is capable of continuous motion in azimuth with fixed elevation and vice versa and at different rates of azimuth and elevation.

(c) It is capable of continuous sidereal motion for following a particular star.

(d) It may be driven to follow several scanning motions (described below).

The control equipment comprises a central console at which the 'controller' sits and two groups of panels with indicator dials visible from the desk.

While the telescope is capable of movement in only two basic co-ordinates, azimuth and elevation, in following a star it is necessary to set up the control in its fixed co-ordinates—right ascension and declination—and a complex computer is required to convert these into the

The above photographs show a general view of the telescope and close-up views of the control desk.

motions of which the telescope is capable; similarly, it is required to be able to scan across or along the Milky Way employing another set of co-ordinates—Galactic latitude and longitude. Again, it will sometimes be necessary to scan in azimuth and elevation and to be able to read off the positions in the other two pairs of co-ordinates—right ascension/declination and Galactic latitude/longitude. The computer must, therefore, be capable of solving instantaneously and continuously, the fundamental equations of spherical trigonometry and to be capable of control in any one of the three pairs of co-ordinates RA/Dec—Az/El—Long/Lat.

The computer takes the form of an electrical analogue in which Magslip resolvers (giving output signals proportional to the sine and cosine of the angle through which their rotors are turned) are employed to solve the trigonometrical equations. The resolvers employed are Muirhead type 2, in which excitation is applied to the stator windings through feedback amplifiers, the feedback voltage being derived from auxiliary stator windings. The function of the electronic amplifier is to obviate the non-linearity in output due to electrical losses and unwanted flux. Fig. 1 shows the type of twin high gain two-stage *RC* coupled amplifier employed in the computer.

A typical example of the computation required, may be seen from that employed to compute E (elevation) from the known quantities L (latitude of station), D (declination) and H (hour angle, which is the difference between the right ascension of a given star and sidereal (star) time). The equation may be written:—

$$\sin E = \sin L, \sin D + \cos L, \cos D, \cos \mathrm{H}.$$

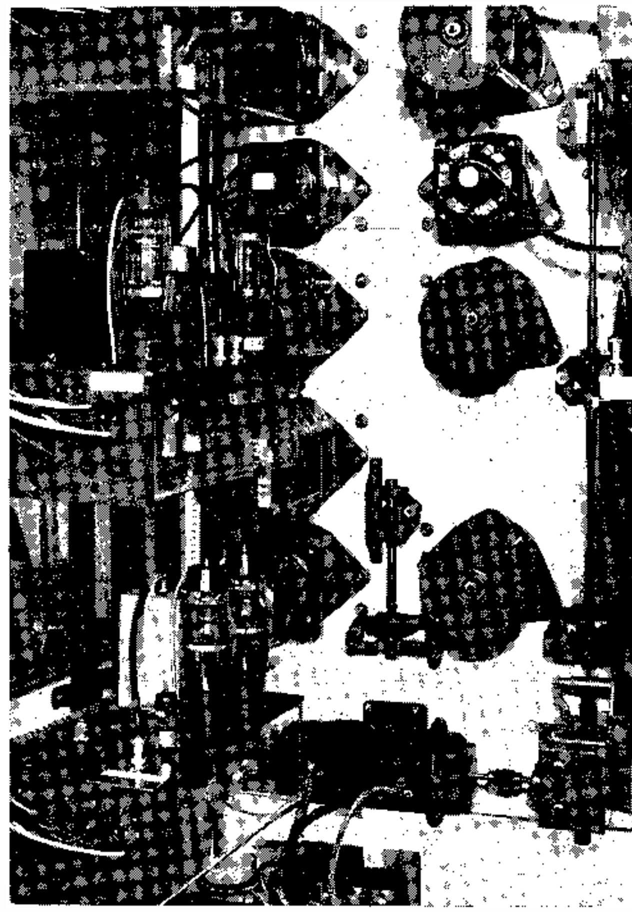
The three known terms are derived from resolvers and combined to provide an 'error' signal given by—

$$F = \sin E - \sin L, \sin D + \cos L, \cos D, \cos H,$$

the error signal being fed to an *RC* coupled amplifier with phase sensitive rectification (Fig. 2) whose output drives a d.c. motor generator (via a d.c. amplifier, incorporating feedback control (Fig. 3)) to which is coupled the E Magslip, in such a sense that a positive value of F drives E towards zero. The basic circuit is shown in diagrammatic form in Fig. 4.

Unfortunately, since no single equation gives sufficient accuracy of control over all parts of the sky, it has been necessary to provide a number of equations which are brought into operation according to the position in the sky of the target. For this purpose cams are provided on the RA, Dec. Az. El. Lat. and Long. shafts which automatically switch in the appropriate equations.

Rear view of control desk showing galactic computer, reference phase selector and servo motor controls



Rear view of the Azimuth co-ordinate panel showing four resolver amplifiers, servo amplifier and servo motor

In all, 14 different equations are employed in the computer:—seven being for calculating azimuth and elevation from hour angle and declination and vice versa, and seven for calculating longitude and latitude from right ascension and declination and vice versa. A block diagram showing the former is shown in Fig. 5, that for solving Long./Lat. from RA/Dec. being very similar.

When a body near the earth is to be observed such as the moon, since the equations are computed on the assumption that the viewer is at the centre of the earth while the telescope is mounted at its periphery, a correction for parallax must be made to the elevation. For this purpose an additional elevation resolver (Elp on Fig. 5) is employed, the amount of parallax correction required, which may be taken from the Nautical Almanac, being set upon a calibrated potentiometer on the control desk.

While the position of the stars in right ascension and declination, remain, for all practical purposes constant, those of the sun, moon and planets are continually moving across the sky. It is, therefore, necessary to apply corrections in right ascension and declination when observing these bodies. The amount of correction—per hour or per day—is shown in the Nautical Almanac in sec/arc for declination and sec/time for right ascension. The required quantities are set up on the control desk which cause motors coupled through differential gears to the

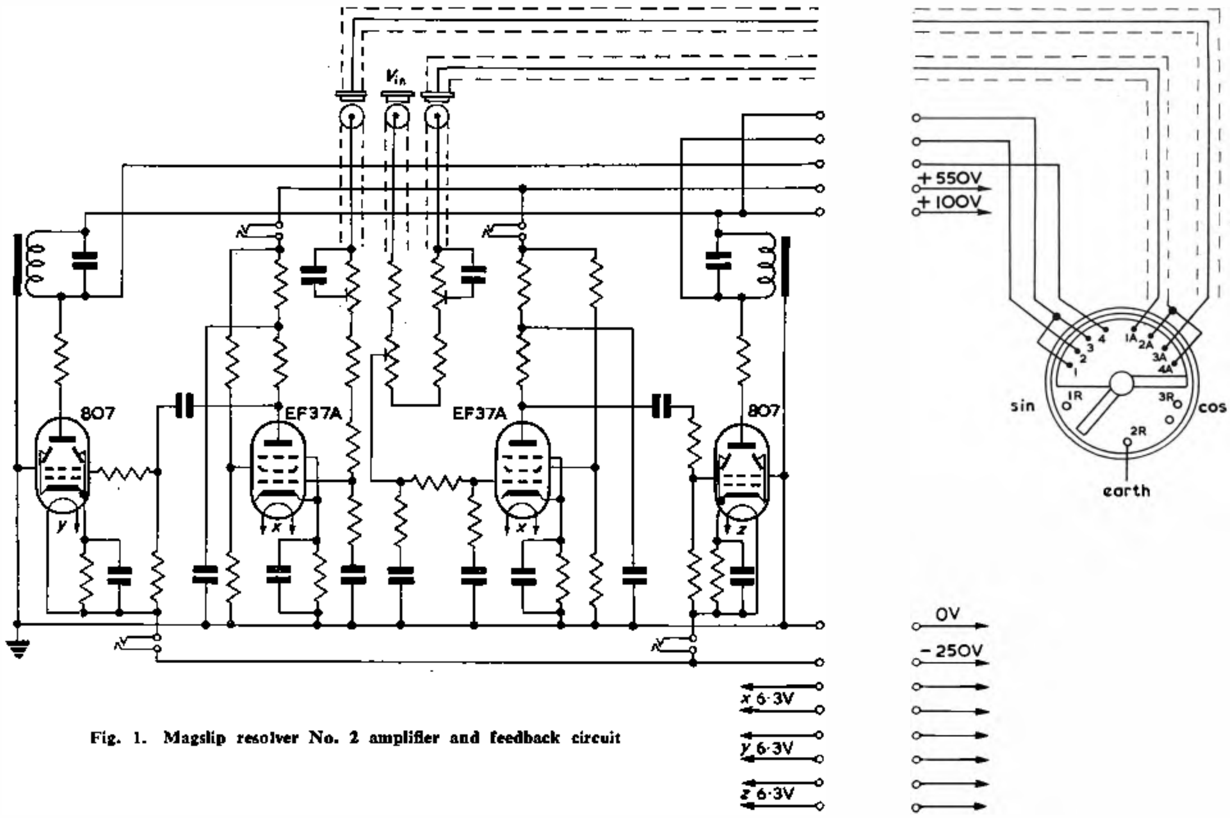
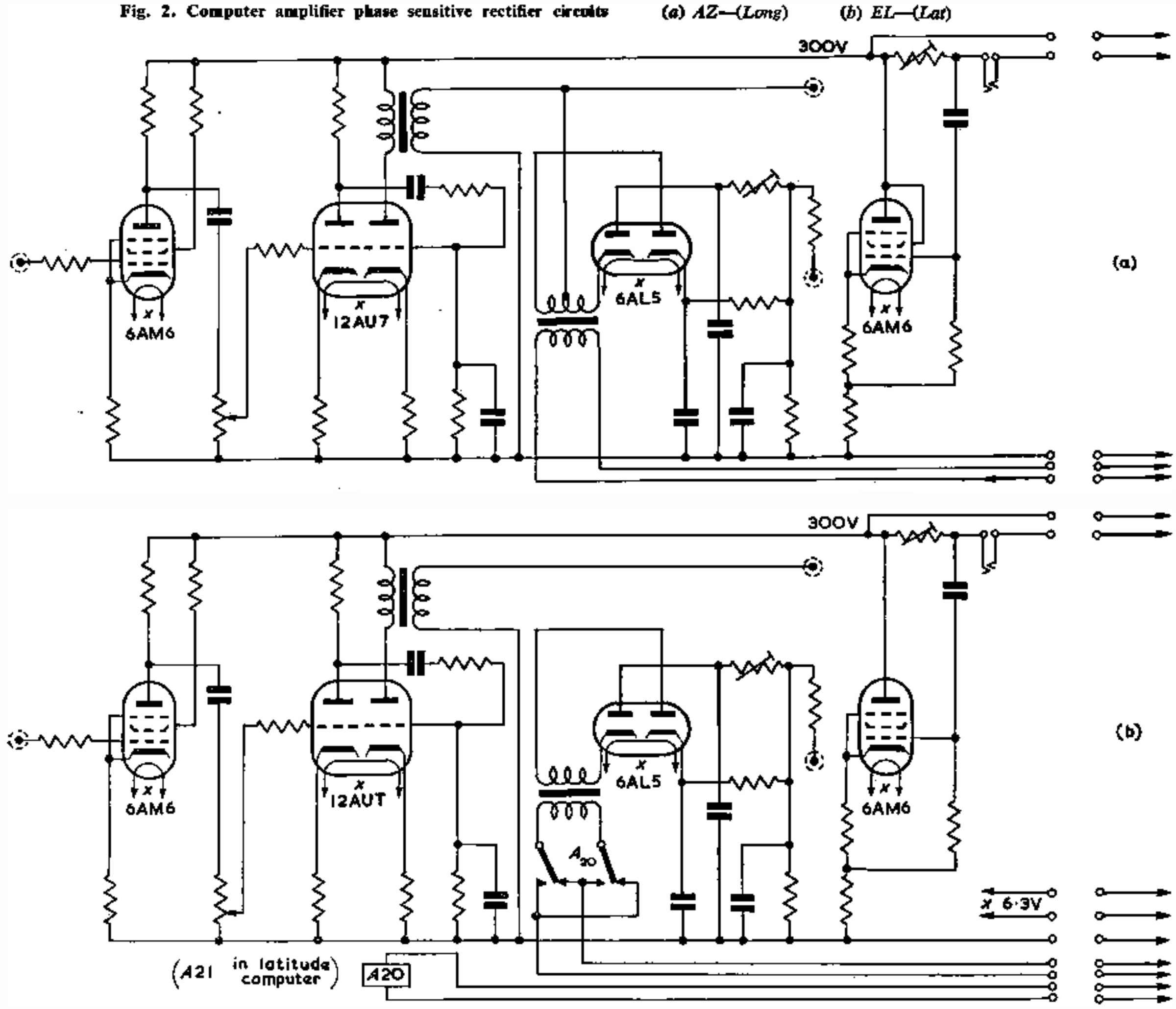


Fig. 1. Magslip resolver No. 2 amplifier and feedback circuit

Fig. 2. Computer amplifier phase sensitive rectifier circuits (*a*) *AZ*—(*Long*) (*b*) *EL*—(*Lat*)



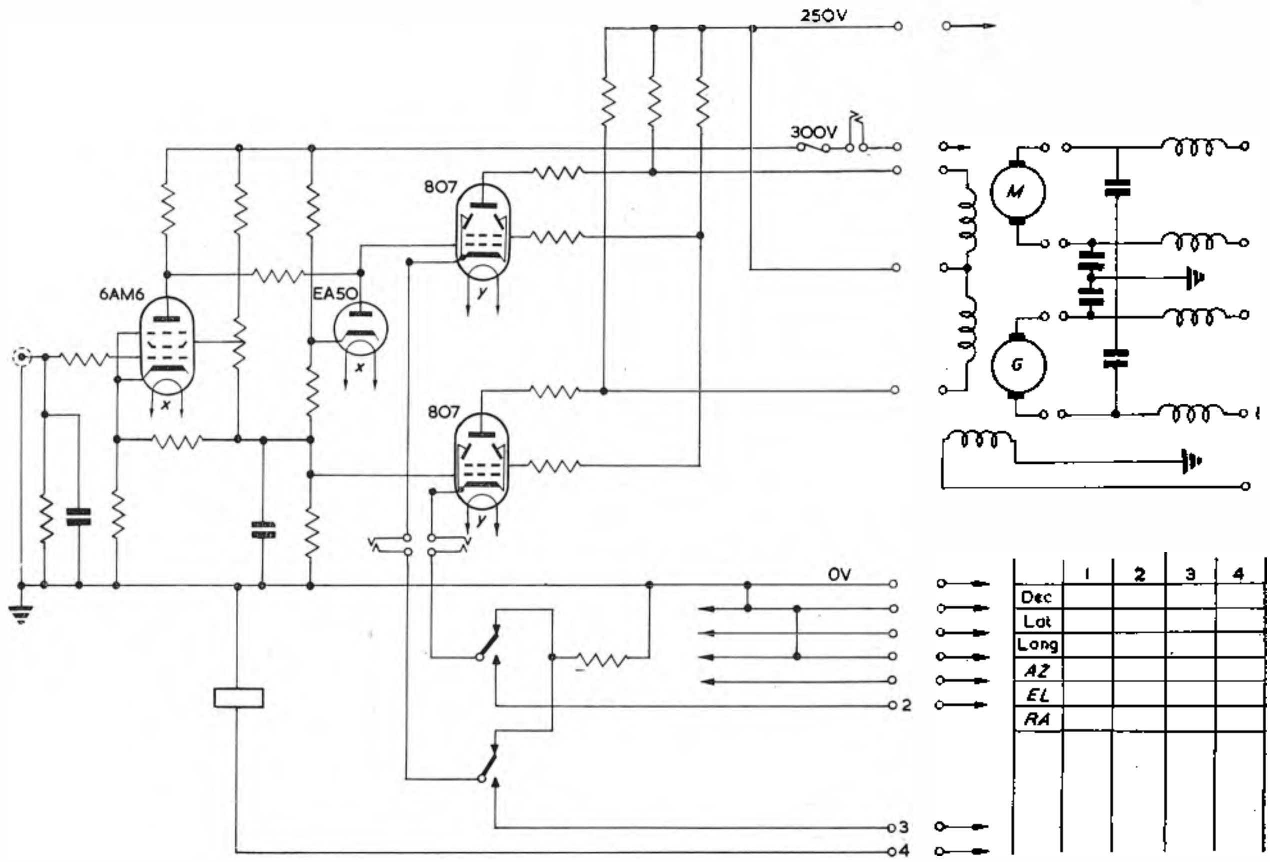


Fig. 3. Velodyne amplifier and Velodyne circuit

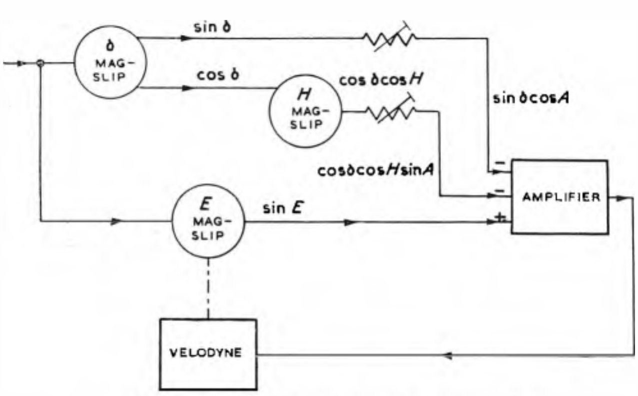


Fig. 4. Schematic layout of co-ordinate control

right ascension and declination computing mechanisms to revolve the required number of seconds of time per hour or per day for right ascension and seconds of arc for declination respectively. The impulses for driving the correction motors are selected by switches from impulses generated by commutators on the sidereal time shaft.

The motors driving the resolvers for the three pairs of co-ordinates are coupled also to indicator dials which indicate in degrees and minutes of arc, and hours, minutes and seconds of time. In addition, sets of dials are provided to indicate sidereal time, universal time and the repeated back position of the telescope itself in azimuth and elevation.

Also coupled to and driven by the motors on the azimuth and elevation shafts are coarse and fine Magslip transmitters (geared 90:1 to each other) which are electrically connected to coincidence transmitters driven by the telescope itself. The output of the coincidence transmitters, which is proportional to the error between the required position given by the control transmitters and the actual position given by the coincidence transmitters is fed to servo amplifiers controlling the Ward-Leonard sets and telescope driving motors in such a sense as to cause the telescope to be driven so as to reduce the error signal to zero. Similarly, transmitters driven by the telescope are coupled to coincidence transmitters on the repeater shafts in the control room to control the repeater drive motors. The fine Magslips, which make one revolution per 10° of the telescope, are provided so that during normal operations where a high degree of accuracy is required, the signal from the fine Magslips gives a tight and accurate control. On the other hand, should the telescope be moved while the control is switched off as must happen on occasions, the coarse Magslips are required to ensure that control is never lost. Valve operated relays are provided to switch automatically from fine to coarse control according to the displacement between the required and actual positions.

The hour angle Magslips are driven by the right ascension motor on the one hand and a synchronous motor running at Sidereal time on the other. The synchronous motor is controlled by an *RC* coupled regenerative oscillator through a power amplifier (Fig. 6). The speed of the motor is compared every 30sec with a pendulum driven master clock and is arranged to drive very slightly faster than the required sidereal time. If, at the time of checking, the motor-driven clock is in advance of the master clock, a check capacitor is closed across the input to V_1

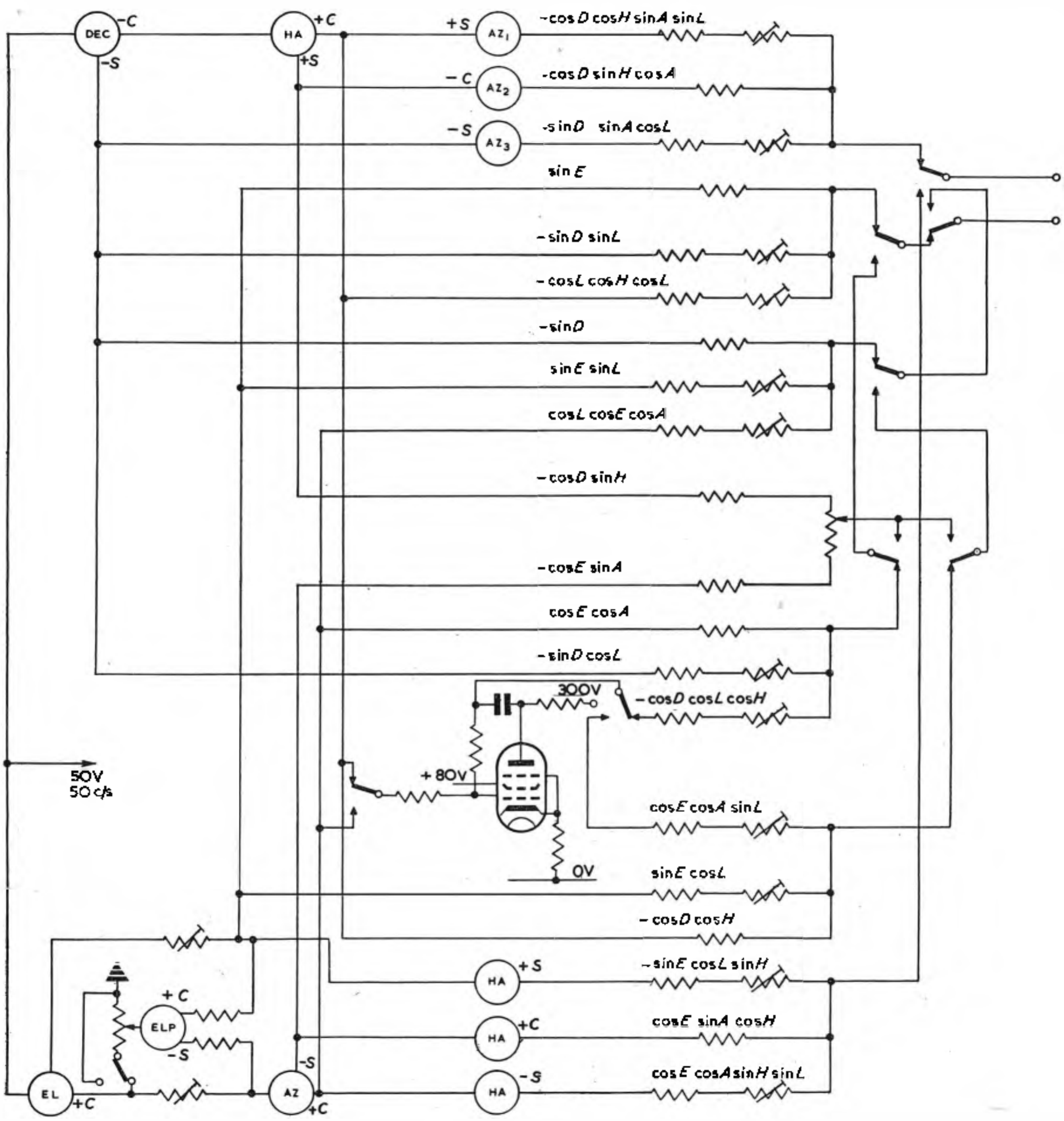
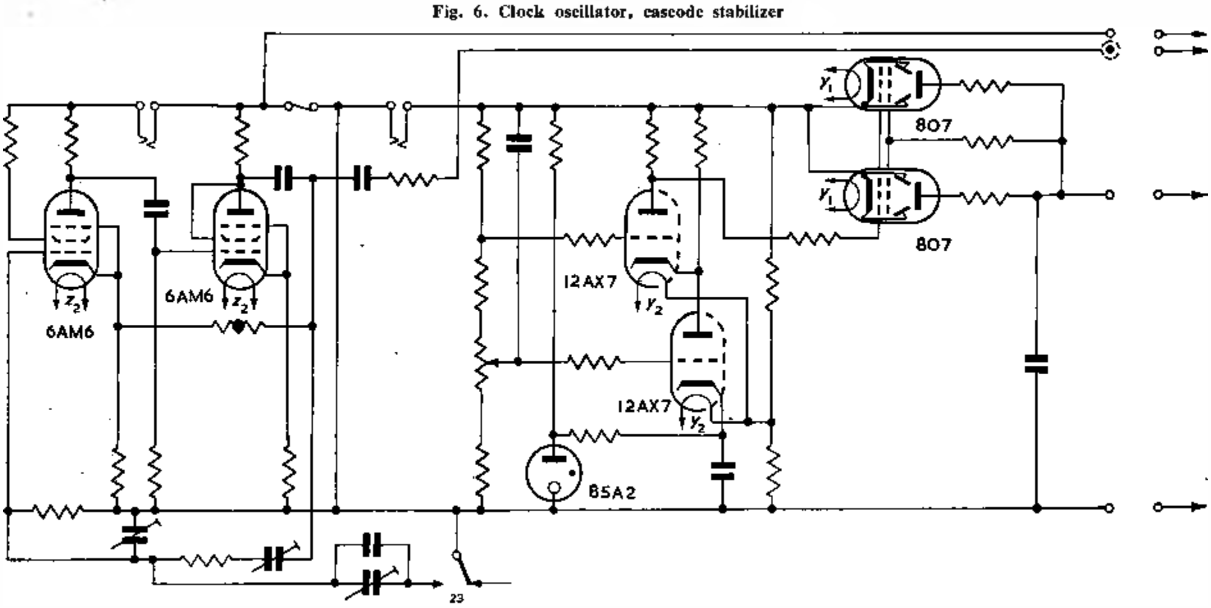


Fig. 5. Azimuth and elevation equation unit

Fig. 6. Clock oscillator, cascode stabilizer



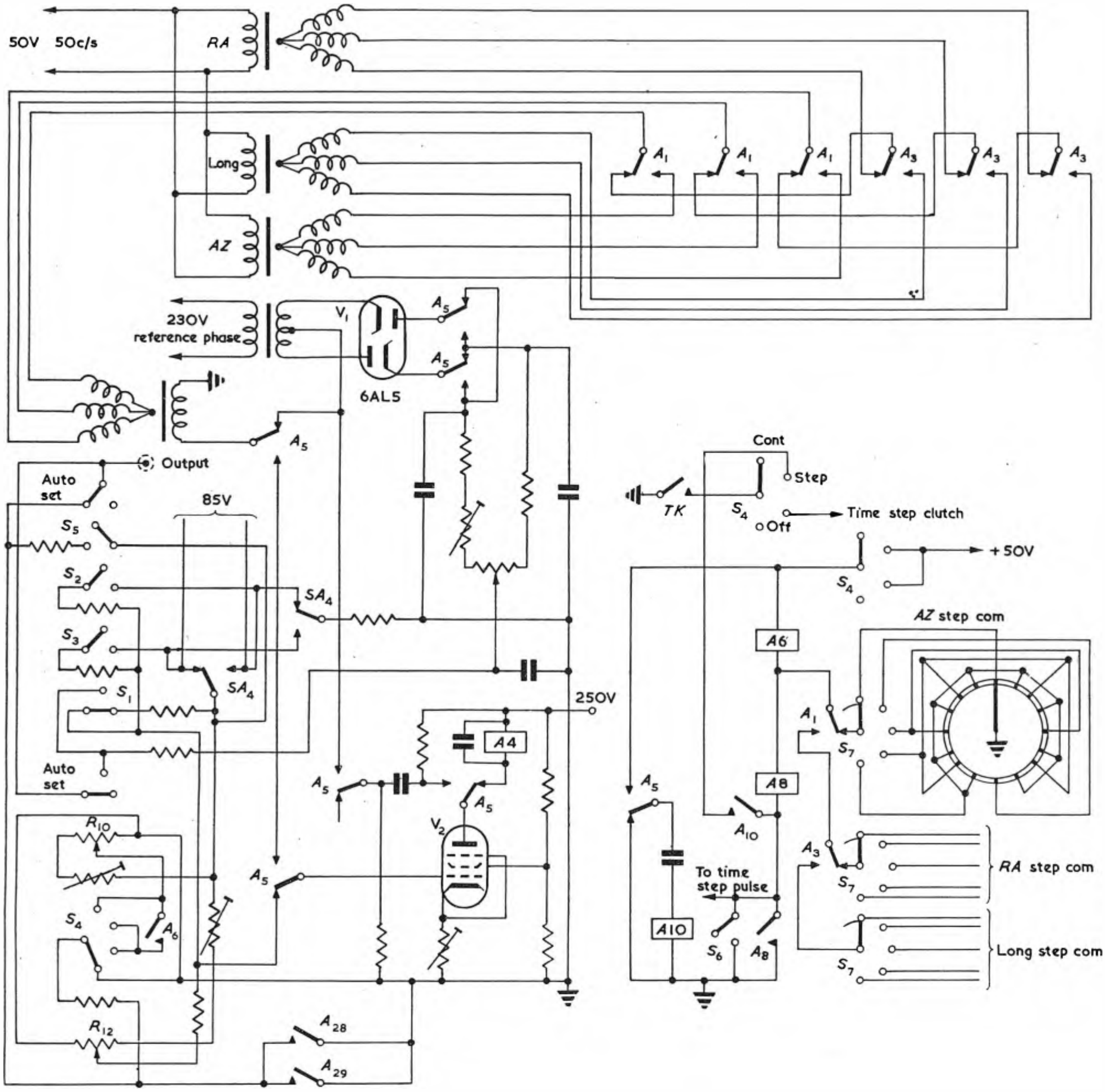


Fig. 7. Step unit, scanning and auto set (AZ, RA, Long)

(see Fig. 6) thereby reducing the frequency of oscillation and the speed of the motor, bringing it back into coincidence with the master.

At any given time, one of the three pairs of co-ordinates (Az/El, RA/Dec, Galactic Long/Lat) is the controlling set, the other two sets being obtained from these by means of the computer. In following a star or planet the controlling pair is RA/Dec, and they may be required to remain constant as in following a star, or, to vary very slowly as in the case of following a planet, the sun or the moon as described above.

On the other hand, it may be required to scan a given area of the sky in a regular pattern in any of the three sets of co-ordinates. Means are provided for scanning in a number of different ways and for setting up the required positions of the driving co-ordinates.

In Fig. 7 with switch S_1 in the normal position, a continuous drive, whose speed is controlled by the potentiometer R_{10}, is applied to the motor generator of the controlling co-ordinate, the sense, positive or negative, of the drive being controlled by operation of S_2 or S_3.

By operation of the 'auto set' switch S_6, the position of the co-ordinate may be set up automatically. Upon operation of this switch valve V_2 becomes a single stage amplifier which drives, via the phase sensitive rectifier V_1, the selected co-ordinate motor until concidence between the position setting Magslip and a coincidence Magslip on the co-ordinate shaft is reached.

Again, the controlling co-ordinate may be arranged to scan across a given arc about a pre-determined centre, the centre being established by the position setting Magslip which is again coupled to the coincidence Magslip on the controlling co-ordinate. The output of the coincidence Magslip is coupled to and compared with the pre-set amplitude bias on the relay valve, the amplitude of the bias being set by R_{12}. The bias determines the limit of the scan by operating or releasing relay A_4 when the Magslip error signal exceeds that of the bias. Operation and release of A_4 changes the sense of the driving voltage, the speed being governed by the setting of the scan speed control R_{10}.

While one of the co-ordinates is scanning about a given centre, in order to provide a second dimension to the scanning raster, the other controlling co-ordinate can be varied by a finite amount. The variation may be initiated each time the first co-ordinate reverses direction or by time impulses which are set upon the control desk. The source of initiation is selected by operation of S_4. If the stepping motion is to be initiated by reversal of direction of scan of the first co-ordinate, operation and release of A_5 equivalent to A_4 in duplicate unit controlling the first co-ordinate) causes a relay A_{10} to pulse. Pulsing of A_{10} operates A_6 and A_8 which connect the driving voltage to the motor generator. The step commutator on the co-ordinate shaft is driven round until an earthed segment is reached which causes A_8 to release which in turn releases A_6 which breaks the driving signal to the motor generator, A_6 will remain released until a further pulse of A_{10}. The size of the step is decided by the commutator segment which is selected by S_7. When it is desired to step on a time-base, relay A_6 is operated by impulses received from the timer instead of by pulsing of A_{10}.

Duplicate scanning units are provided so that each of the co-ordinates in the driving pair may be controlled independently.

During scanning, the telescope and the other sets of co-ordinates follow via the computer.

Means are provided so that if, while following a star, the telescope reaches its limit of rotation in azimuth, a contact on the co-ordinate shaft causes the normal signal to the azimuth motor generator to be disconnected and a fixed input is fed in causing the motor to drive in the reverse direction for 360° when a further contact restores the normal signal and the drive is resumed and the star is again picked up. During the auto-reversing of azimuth, the input to the elevation is disconnected so that the elevation position is locked.

Circuits are provided so that in the event of component breakdown or other fault causing errors either in computation or following, normal tracking of the telescope will cease and alarms will be raised. Interlock facilities are provided to prevent entry into the bowl or to the swinging laboratory while the telescope is in motion and to prevent the telescope from being moved while personnel are in the bowl.

A feature of the control gear is the easy accessibility of all chassis and components for maintenance and servicing. All chassis connexions are made by plug and sockets for easy and quick replacement.

Ratios of 21 600:1 exist between the Magslip resolvers and the motor generators which they control and which in turn drive them, a small amount of backlash in the drive could therefore introduce large errors and ambiguities in the computations and control system. Great care, has therefore, had to be taken in the mechanical design and construction to avoid backlash in gearing and to keep friction to a minimum.

A New Radar Fire Control System

The first details of a new radar and fire control system which was developed for Britain's new supersonic all-weather fighter—the English Electric P.1B have now been released. Developed by Ferranti Ltd, at Edinburgh, the system is called AIRPASS (Airborne Interception Radar and Pilots' Attack Sight System) and is one of the most advanced and versatile airborne radar systems in the world.

At the time of the Battle of Britain ground radar was necessary to direct single seat fighters so that they could make contact with the enemy. Today aircraft speeds and operating altitudes have increased vastly and longer range weapons are available so that unaided visual contact by pilots flying day fighters can no longer ensure a successful attack.

It was to meet this requirement that AIRPASS was developed. The pilot using this system may never see his target and yet still destroy it. The system forms a vital link in the chain of defence, enabling a fighter pilot to search for and find any hostile aircraft by day or night, manœuvre into a suitable position for attack, aim his weapons and destroy the target.

AIRPASS comprises a very advanced radar unit and a sighting system. The former gives information to the pilot in simple usable form, enabling him to intercept the target and lock the radar beam to it. Thereafter the radar accurately and automatically tracks the target. It feeds information to the sighting system where it is presented to the pilot in such a way that he can quickly and easily manœuvre into a suitable position from which to attack with guns, rockets, or guided weapons. The sight ensures that the weapons are correctly aimed even though the target may be invisible to the pilot and travelling at supersonic speed.

In a typical interception, ground radar will set the pilot in the general direction of the approaching target. AIRPASS will then take over. The radar beam will scan a sector of sky covering a wide angle horizontally and vertically, on either side of the flight path, and extending many miles ahead of the fighter. When the target is located the radar will lock-on and a computer in the radar unit will automatically work out the best approach course for the pilot to follow.

The pilot follows the correct approach and completes his attack using the information presented to him in his 'sight'. The whole operation may be completed without the target ever being seen by the pilot, but if it should come into his view he can take advantage of the normal visual sighting methods which are included in the system. An automatic warning device is incorporated to tell the pilot when to break off the attack if he should approach close enough to be in danger of colliding with the target.

In AIRPASS the radar unit is housed in a single container or 'bullet'. This design has many advantages. In the case of the English Electric P.1B, the 'bullet' forms a complete centre-body in the engine air-intake. In this way the P.1B carries a powerful radar equipment, without increasing the frontal area unnecessarily, and at the same time preserves a good aero-dynamic form. A considerable saving in weight is effected by the elimination of the usual number of boxes and interconnecting cables, particularly since the radar 'bullet' actually forms part of the aircraft structure.

The packaged design of AIRPASS is small enough to be installed even in lightweight fighters.

The Advantages and Disadvantages of

DIRECT OR A.C. COUPLING FOR DEFLEXION AMPLIFIERS

By H. L. Mansford*, B.Sc., A.C.G.I., A.M.I.E.E.

To give of their best, oscilloscopes must both display the waveform and provide for amplitude and time measurement. Amplitude measurement should be independent of amplifier gain and linearity.

The advantages and disadvantages of direct and a.c. coupled amplifiers are discussed, and there follows a brief description of a quasi d.c. coupling system using a vertical deflexion signal amplifier input switch, covering the frequency band d.c. to 100Mc/s.

GENERAL purpose c.r.t. oscilloscopes may be classified in accordance with the method of circuit coupling used for the vertical, or Y deflexion, e.g.,

(1). *Direct Coupling*, generally using long-tailed pairs of valves.

The advantages and disadvantages of both balanced and unbalanced d.c. coupled amplifiers have been examined elsewhere[1,2].

It has been shown that direct coupling offers great advantages to the user for many circuit applications, and also enables both d.c. and a.c. signal amplitudes to be metered on a calibrated Y shift, without reliance being placed upon an amplifier gain figure.

(2). *A.C. Coupling*, requiring time-constants adequate for the lowest signal frequency.

The l.f. limitation of a.c. couplings shows as a phase shift giving pulse shape distortion, unless time-constants are made at least one hundred times greater than the maximum signal pulse width.

Large coupling capacitors restrict the h.f. response; and interference from supply mains bumps, via the h.t. line, becomes a major problem.

Signal amplitude measurement is usually restricted to the use of a calibrating signal in conjunction with a graticule, and the d.c. signal component cannot be displayed or measured.

(3). *Carrier Frequency Systems* using a signal modulated r.f. carrier, enable demodulation also to restore the d.c. signal component, after a.c. couplings.

Such systems give the advantages of d.c. together with the simplicity of a.c. inter-stage couplings. A system has been described by Goddard[3], and like (1) it also makes use of the d.c. measuring facilities at the amplifier input.

An alternative method, which gives less restriction at the upper limit of the frequency band, will now be described.

Capacitive Coupling, with Pilot Pulses and D.C. Restoration

The system is inspired by television circuits, and it requires a small part of the main signal to be suppressed in favour of a pilot signal. This gives a reference potential for d.c. restoration by clamps.

Amplifier capacitive couplings, followed by d.c. restoring circuits, use a technique familiar to television circuit engineers; but they were not the first to make use of the basic principle involved.

Imagine a centipede on the march. His body is the waveform being transmitted. Gravity provides the bias, moving the waveform down, while each leg gives a point of support or d.c. (demi-centipede) restoration; raising the body to a constant height above the reference level: in this case moulding it to the contour of the surface being traversed.

A signal may behave in this same way if it be given a 'leg' in the form of a pulse to push the waveform away from the reference potential, together with some bias to edge it steadily towards this potential. This is outlined by the circuit of Fig. 1(a) with the waveform of Fig. 1(b) for the signal. The negative going pulses, which are an

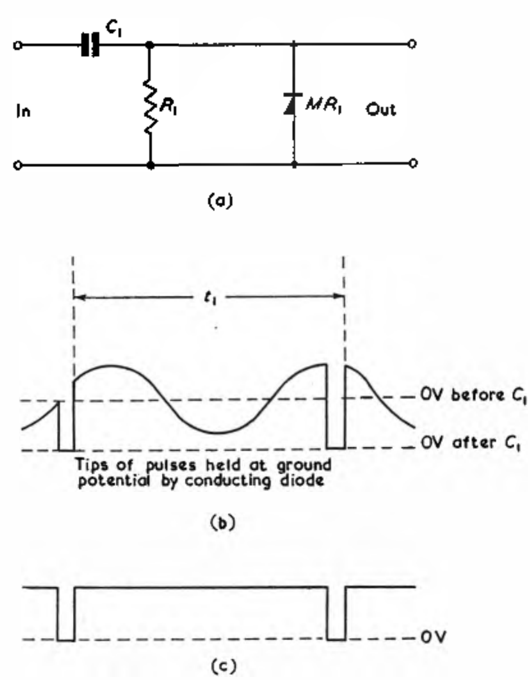


Fig. 1(a). An a.c. coupling followed by a diode for the restoration of a d.c. signal component
(b). Input waveform plus a pilot signal to be used for d.c. restoration by the tips of pulses held at ground potential by a conducting diode
(c). Fig. 1(b) input waveform becomes d.c. only

essential and constant part of the signal waveform, provide the 'legs'. The remainder of the signal may take any form so long as it makes no negative excursion to exceed that of the 'legs'.

This is a simple asymmetrical form of d.c. restoration, and it is seen that it requires a waveform which is a mixture of two signals. The main and normally variable part is considered as existing continuously because should its a.c. component cease, there may still be a d.c. level upon which any a.c. signal may build (Fig. 1(c)).

This continuity is broken periodically by inserting the second or pilot signal, to the total exclusion of the first

* *Bell Punch Co. Ltd, formerly E.M.I. Electronics Ltd.*

or main signal. Obviously the pilot pulse period is time lost to the display, and for this reason it is made a short part only of the cycle.

Reference to Fig. 1 will show that the diode current flowing during the short pilot pulse period has to balance the leak (R_1) current flowing for the remainder of the period. Also the change of level during the pilot pulse period—and therefore also during the display period—must be a fraction of one per cent only, if the error is to escape detection. It follows from this that the time-constant C_1R_1 (of Fig. 1) must be adequately long relative to the time interval (t_1) between pilot pulses.

As one of the objects to be attained by the introduction of d.c. signal restoration is to shorten the coupling time-constants, the pilot pulse interval must be short compared with the lower signal frequencies to be displayed.

An alternative method of meeting the requirement of a long time-constant is offered by the use of a double acting clamp switch[3], such as that of Fig. 2. Here the Fig. 1 bias potential operating through resistor R_1, is

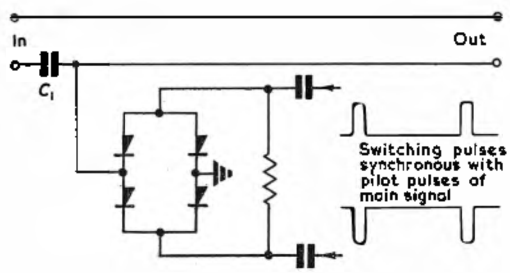


Fig. 2. A four diode clamp switch pulling the pilot signal up or down to chassis potential

omitted, leaving no direct circuit connexion except during the pilot pulse period. This offers advantages:—

(a) It gives a pilot pulse reference potential which may lie within the envelope of the normal signal excursion (Fig. 5(b), because its action is symmetrical.

(b) It gives long coupling time-constants with small capacitor values.

(c) The clamp switch diodes are normally only called upon to shunt away from the coupling capacitor those currents (charges) arising from the capacitor leakage, or from the following grid.

But should the need arise, as will follow if a hum or other interference signal be introduced to the amplifier circuit after the addition of the pilot pulse, the low impedance clamp switch can quickly change the charge in each low value coupling capacitor, so as to retain the correct mean signal potential. In effect, the clamps provide for a short time-constant when interference signals are involved, but a long one for the wanted signal.

This operation is illustrated by the waveforms of Fig. 3. A d.c. only input signal (negative) has been chosen to make the distortion which follows, more obvious.

The correct Fig. 3(a) waveform may have an a.c. interference signal added to it by an unwanted pick-up in the amplifier chain; and because this is of a lower frequency than that of the pilot pulse, Fig. 3(b) indicates the errors due to a part only of one cycle of this interference. Fig. 3(c) shows that the d.c. clamp switch gives the signal a new and correct start after each pilot pulse has been clamped to the d.c. reference line, and the maximum error can never exceed that accumulated during a t_1 period.

An oscilloscope display amplifier is naturally associated with those pulse circuits to be found in the time-base generator. It should be practicable to utilize clamp switching pulses synchronous with the time-base, if use is made of a double acting switch, such as that of Fig. 2. This gives a display which is self blanking, because the pilot pulse period may be the normal time-base return time.

One insurmountable objection remains however, and that is the chaotic amplifier circuit conditions which can

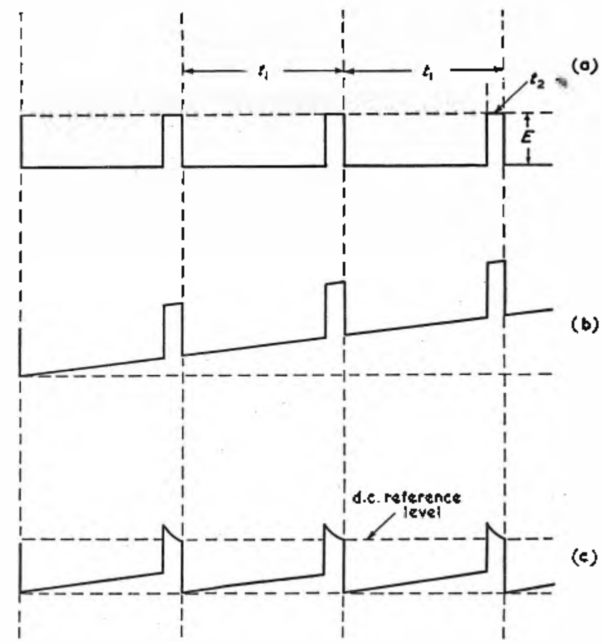


Fig. 3. A negative d.c. potential E plus the pilot signal (for period t_2):
(a) Before the pick-up of an error (hum) signal
(b) The wanted plus unwanted signals
(c) The unwanted signal minimized by the clamp switch action

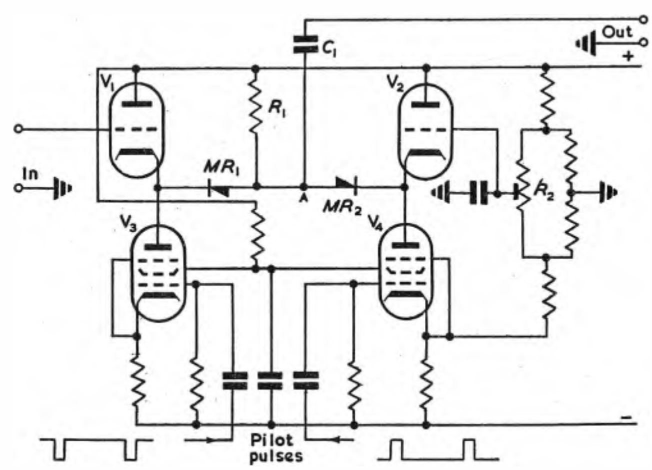


Fig. 4. A cathode-follower and crystal diode signal switch, for the addition of a pilot (d.c. reference) signal

arise while the time-base is inactive, as it may be if it is awaiting a triggering signal.

A simple way of overcoming this obstacle is not to use the time-base pulse source, but an alternative free running source[4], not synchronous with the time-base repetition period. The c.r.t. beam may readily be blanked by the pilot pulse, which leaves any interference which could become visible, in the form of short dark sections of the trace.

Practical tests on an oscilloscope have shown that by

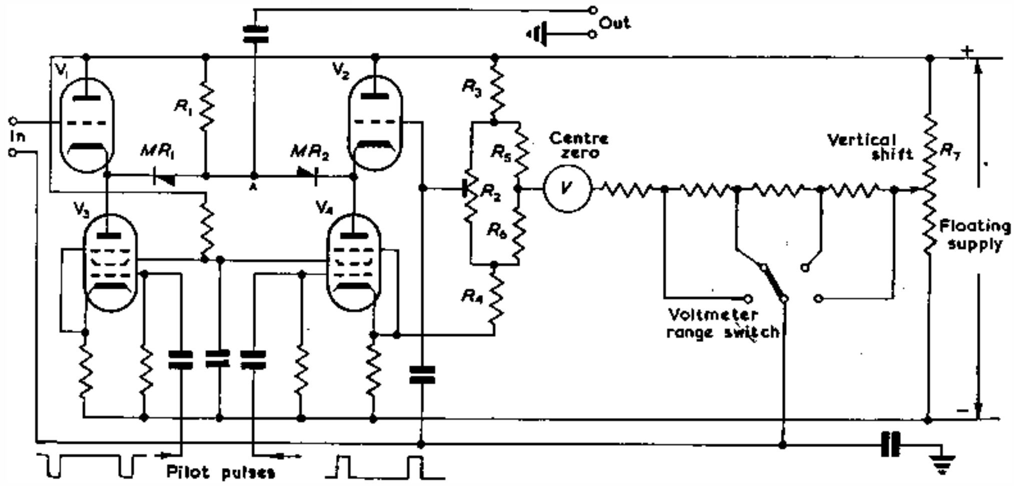


Fig. 6. An extension of the Fig. 4 circuit to cover a large d.c. voltage range for the reference potential

keeping the pilot pulse period t_2 at 4μsec or less, it is impossible to resolve these dark intervals, because on slow time-base rates the width is too little, while on faster sweeps the persistence of vision fills in the gap, assuming asynchronous conditions.

A pulse frequency control oscillator was used at around 10kc/s, with provision for some tuning to avoid beats with the time-base frequencies, but this was found unnecessary.

A separate amplifier is of course essential for the time-base synchronizing or triggering signal, in order to avoid injecting a signal at pilot pulse frequency where it is least wanted.

The design of an oscilloscope amplifier input switch is now briefly described. Additions have been made to provide d.c. vertical shift potentials, which are metered for signal amplitude measurement.

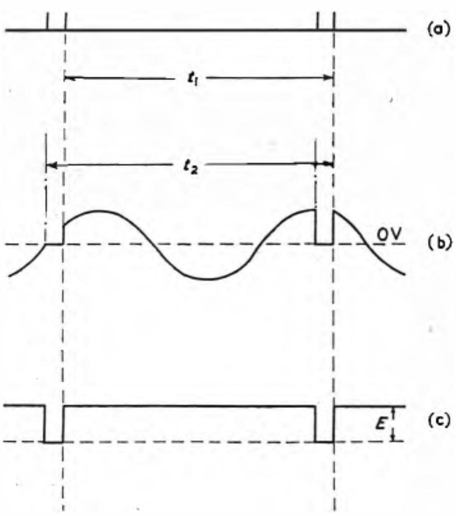


Fig. 5. Waveforms at circuit point A of Fig. 4
(a) Circuit balanced. Zero signal input. Signal due to switching
(b) With an a.c. signal at the input
(c) With a positive direct potential at the input

The pilot pulse mixing circuit of Fig. 4, outlines an electronic switch found suitable over the bandwidth; d.c. to 100Mc/s. V_1 and V_2 are similar valves connected as cathode-followers; and a valve type is chosen to give a high mutual conductance and short grid base.

V_3 and V_4 switch current alternately between V_1 and V_2. The switching is controlled by pulses, going negative at V_3 control grid for the pilot pulse period, and positive at V_4 grid for this same period.

In the absence of grid pulses, V_4 is biased to cut off its anode current, so leaving V_2 without current and germanium crystal diode MR_2 not conducting.

V_3 draws some anode current from R_1 through MR_1, and the remainder through V_1. For a fixed current value in V_3, V_1 grid d.c. potential fixes also the somewhat more positive potential at its cathode; which is also the potential at A, if the drop across MR_1 (conducting) be ignored.

During the pilot pulse period t_2, V_2 takes control, and its grid potential may be set at R_2 to give the same potential at A for this shorter period. Obviously there should then be no output signal, excepting short duration 'whiskers' during the switch-over. Fig. 5(a), illustrates this waveform. Fig. 5(b) shows the output signal waveform when an a.c. signal is applied at the input. Fig. 5(c) gives the output for a change of d.c. level at the input; the height E representing the d.c. input signal which, with clamps for d.c. restoration, can be either positive or negative.

In order to measure input potentials up to ±10V (say), it is necessary first to adjust R_2 setting for zero signal out when the input is zero. This may set the trace to the horizontal cursor in front of the c.r.t. screen. Then a d.c. potential at the input (V_1 grid) may be balanced by an equal change at R_2 (V_2 grid). This restores the trace to the cursor line, and R_2 voltage change may be read on a meter to give also the voltage at the input.

Fig. 6 shows a modification to the d.c. shift facilities, which enables the range of the shift voltage to be extended indefinitely without circuit limitations other than those arising from insulation. R_2 setting now reverts solely to a zero balance control to give zero output signal when there is zero signal input. For this adjustment, the voltmeter must also be set to its centre scale zero by using the vertical shift control.

R_3, R_5, R_6, R_4 constitute a potential divider, passing a current several times greater than that required by the voltmeter V. Now voltage changes at the input can be balanced by the vertical shift potential, which is metered by V, as in circuits using 'old style' d.c. coupling[1].

From this brief survey it is apparent that the full measuring facilities are made available at the amplifier input; so making gain variations and non-linearity of no importance when the amplifier and c.r.t. are used in this way as a null indicator.

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Performance Calculations for D.C. Chopper Amplifiers

By I. C. Hutcheon*, M.A., A.M.I.Mech.E., A.M.I.E.E.

It is shown how input resistance, output resistance and overall d.c. gain can be calculated for a chopper amplifier employing any arrangement of perfect switches, resistors and capacitors in the chopping and rectifying circuits. Formulae are derived for two particular arrangements, one of which can provide fast response with little output ripple if the switching sequence is suitably chosen.

THE use of chopper amplifiers for amplifying small direct voltages with a minimum of drift is now common. The input signal is converted by the chopper into a.c., amplified as such by an a.c. coupled amplifier, and usually reconverted to d.c. by a demodulator.

A common application is the stabilization of a direct coupled computer amplifier[1] by a chopper amplifier connected in a subsidiary loop so as to observe the input drift and inject a correcting signal. The composite amplifier then has a wide frequency response, and a long term drift, referred to the input, which is typically below 1mV.

Industrial self-balancing potentiometers frequently incorporate a chopper amplifier whose output, however, is retained as a rule in a.c. form to drive a reversible balancing motor. In such cases the input voltage drift may be as low as $\pm$ 1 or even $\pm$ 0·1μV and the current drift of the order 10^{-11}A. Drift depends very largely on the properties of the chopper, and a mechanical switch is generally used for results of this order.

Still lower current drift can be obtained with the vibrating capacitor modulator, which gives a sine wave output but is difficult to make with a voltage drift below about 100μV. Transistor choppers, in which the transistors are used as switches, have already been developed[2] to give stabilities of $\pm$ 100μV and $\pm 3 \times 10^{-8}$A, and by virtue of their reliability and low cost will certainly supersede the mechanical chopper in suitable applications[3]. Lower current drifts are probable with the advent of silicon transistors, and useful results have been achieved with silicon diode choppers.

There exists a considerable body of literature dealing with chopper design and drift problems, but the analysis of switched chopping and demodulating circuits in general terms seems to have been somewhat neglected.

This article shows how the following characteristics can be calculated for a chopper amplifier which employs any arrangement of perfect switches, resistors and capacitors in its chopping and rectifying circuits.

(a) Input resistance to d.c.

(b) Overall d.c. gain.

(c) Output resistance.

These are the parameters whose values usually have to meet desired specifications. General formulae are derived for two particular circuits in terms of resistor values and switching times. It is also shown that a suitable choice of switching sequence permits one of the circuits to deliver a smooth output with small smoothing-circuit time-constants; and this in turn permits faster response.

Input Circuits

Consider the simple chopping circuit shown in Fig. 1. V_1 and r_2 represent the d.c. source, r_3 is the load imposed by the following amplifier, and C_2 is a high quality coupling capacitor which has zero leakage. The chopper is a perfect switch connected across the circuit and operates continuously, being closed for a fraction x of the time and open for $1 - x$. This is the simplest possible circuit, requiring only an on-off switch one side of which is connected to circuit earth.

When the switch is open a current flows into the capacitor and tends to charge it, the reverse occurring when the switch is closed. C_2 is thus charged to a potential V_2 over a number of initial cycles and is assumed sufficiently large to maintain this potential steady during each ensuing cycle. It may therefore be treated as a battery of low resistance.

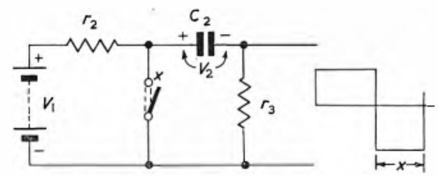


Fig. 1. On-off chopping circuit

Equating the charge flowing into C_2 during $1 - x$ to that flowing out during x gives the value of V_2:

$$V_2 = \frac{(1 - x)\, r_3}{x r_2 + r_3} V_1 \quad \ldots\ldots\ldots\ldots \quad (1)$$

The output waveform is a perfect rectangle whose peak-to-peak amplitude is:

$$\text{Output} = V_2 + \left(\frac{r_3}{r_2 + r_3}\right)(V_1 - V_2) \quad \ldots\ldots\ldots \quad (2)$$

which, with equation (1), gives the circuit gain:

$$\text{Gain} = \frac{\text{Output}}{V_1} = \frac{r_3}{x r_2 + r_3} \quad \ldots\ldots\ldots \quad (3)$$

Equation (3) fully defines the performance of the circuit under steady conditions. x is generally of the order 0·5, and r_3 is chosen rather larger than r_2 to provide a reasonable gain.

Sometimes it is desired to use a filter to remove stray a.c. from the incoming signal. The basic arrangement is shown in Fig. 2, where C_1 is large enough to maintain a steady potential V_1 and can be treated as a battery in the same way as C_2. A pulsating current is drawn from C_1 whose mean value is:

$$I = \frac{V_1\, x\, (r_2 + r_3)}{r_2\, (x\, r_2 + r_3)} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

This current flows steadily through r_1, giving a voltage difference $I r_1$ between V and V_1. The latter can thus be found in terms of the former, and the new circuit gain, from equation (3) is:

$$\text{Gain} = \frac{\text{Output}}{V} = \frac{1}{1 + x\,((r_1/r_2) + (r_2/r_3) + (r_1/r_3))} \quad (5)$$

* *George Kent Ltd.*

If V and r_1 are considered as the source, then the circuit from C_1 onwards acts like a pure resistance V_1/I which is its 'input resistance.' From equation (4)

$$\text{Input resistance} = r_2 + \left(\frac{1-x}{x}\right)\left(\frac{r_2 r_3}{r_2 + r_3}\right) \quad \ldots\ldots\ldots (6)$$

The gain of this circuit tends to zero if r_2 is very large or very small. There is therefore an optimum value for r_2 which gives greatest gain for given values of r_1 (the source) and r_3. It is found, by differentiating equation (6) to be

$$\text{Optimum } r_2 = \sqrt{(r_1 r_3)} \quad \ldots\ldots\ldots\ldots (7)$$

These two circuits are the simplest possible, and are the ones generally used when only on-off switching is available.

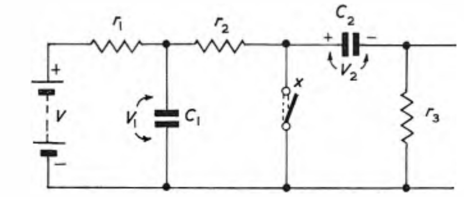


Fig. 2. Basic circuit with filter

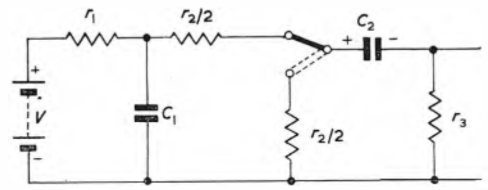


Fig. 3. S.P.D.T. chopping circuit

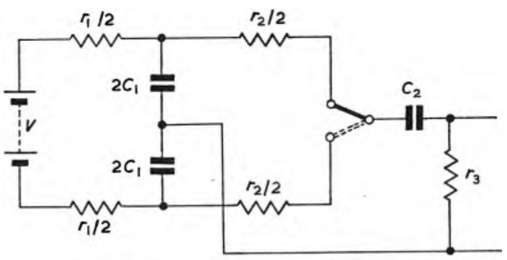


Fig. 4. Balanced circuit

With transistor choppers in particular, it is desirable that the base-drive circuit should be joined to circuit earth, so that stray currents are not injected into the chopping circuit. This makes double-throw operation more difficult. Again, if a s.p.d.t. mechanical chopper is required both to chop and to demodulate, only on-off chopping is possible, and one side of each switch must be earthed.

They are not the most efficient circuits, however, since for a large part x of each cycle, the input is shorted out to no good purpose, and the input resistance is lower than it need be for a given load resistance. This does not matter when the source resistance is low, but may be a limitation when it is high. More seriously, the circuits are unsymmetrical and therefore affected by a.c. pick-up at the chopping frequency or its harmonics.

The efficiency can be improved by the arrangement of Fig. 3, wherein the input is disconnected when the coupling capacitor is earthed. If the switch is break-before-make, r_2 may be omitted, but undesirable transients may be set up during the transit time since the following amplifier then sees a relatively high impedance. Make-before-break operation seems generally preferable if the overlap is kept small. r_2 must then be present to limit the drain from the filter capacitor during the overlap, and is found to have an optimum value depending on r_1, r_3 and the overlap duration, being small when the duration is small. Better symmetry is obtained if r_2 is split into two parts as shown.

The circuit remains susceptible to pick-up unless a balanced version is adopted as in Fig. 4. This separates the input from circuit earth and may not be convenient, e.g. where d.c. feedback is to be applied. In such a case transformer coupling is a solution. Alternately the circuit may be operated at a frequency differing from that of the supply mains. Capacitively coupled input circuits can all be analysed in the manner outlined above, and some, including two with transformer coupling, are dealt with elsewhere[4].

Output Circuits

The simplest switch-type demodulator circuit is shown in Fig. 5. R_1 represents the amplifier output impedance,

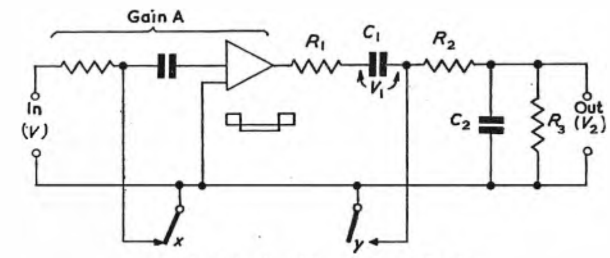


Fig. 5. Circuit with on-off rectification

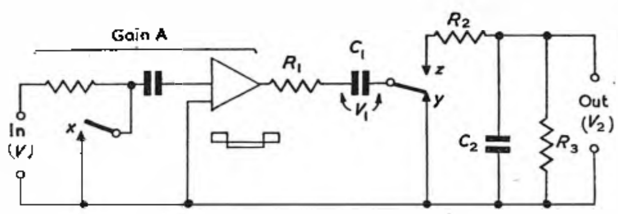


Fig. 6. Circuit with s.p.d.t. rectification

R_2C_2 is a smoothing circuit, and R_3 is the load. An on-off switch operates continuously at the chopping frequency and earths the coupling capacitor C_1 while the amplifier output is (say) positive; C_1 is thus charged positively on the left. When the output reverses, the switch is opened, and the negative signal, assisted by the potential of C_1, is applied to the reservoir capacitor C_2.

This circuit suffers from the disadvantage that C_2 is drained whenever the switch is closed. This is overcome by the circuit of Fig. 6 which, however, needs a double throw switch.

Ideally the output switch, however arranged, should open and close in exact synchronism with the signal reversals. This prevents signal reversals reaching C_2, gives greatest efficiency and minimizes output ripple. Exact synchronism cannot in practice be achieved with mechanical choppers (it may be approached with transistors) and it may therefore be desirable to work away from this condition in the best direction, e.g. that giving least ripple. There are also some advantages in using unsymmetrical mark-to-space ratios, and it is therefore worthwhile analysing the circuits for the general case in terms of the switching times.

Analysis

First, assume that the input circuit and amplifier together have a gain A, giving a rectangular wave output of magnitude AV for a direct input V. For convenience, the waveform is considered as generated by an on-off switch closed for a fraction x of each cycle and open for $1 - x$. C_1 and C_2 are assumed large.

Each circuit exists in a number of different states which occur either once or twice every cycle. The total time spent in each state, expressed as a fraction of the cycle, is denoted by $p, q \ldots w$ which are defined in Table 1. These circuit state durations fully define the two switching circuits and their phase relationship.

The durations x, y, z, for which each switch is closed, are given conversely in terms of the circuit state durations in the same table. These define the two circuits individually, but not their phase relationship. Analysis is made in terms of circuit state durations, and the results are reconverted as far as possible into expressions in x, y, z.

During each circuit state, steady currents flow through C_1 and C_2 which are large and can be regarded as batteries. The currents can be written down in terms of V, V_1, V_2 and the resistances, any convenient arbitrary d.c. level being assumed for the a.c. signal.

Multiplying each current by the appropriate duration gives the charges flowing into or out of the capacitors which must total zero over a complete cycle for each capacitor, whence two equations are obtained which can be solved to give the overall gain V_2/V.

It is more convenient to think in terms of open-circuit gain (R_3 = infinity) and open-circuit output resistance, both of which are easily deduced from V_2/V (R_3 present). These results are listed in Table 2 and fully define the circuit performance.

Output resistances are independent of the waveform and hence reconvert fully into expressions in y and z. Gain depends on x and its phase, and so requires circuit state durations to define it. As far as possible, however, x, y and z are used in the expressions for gain.

TABLE 1.
Definitions of Circuit State Durations

	CONTACT(S) MADE			
	ON-OFF RECTIFYING CIRCUIT	S.P.D.T. RECTIFYING CIRCUIT		
DURATION OF CIRCUIT STATE		y & z OVER-LAPPING	y & z NOT OVER-LAPPING	OVERLAP OR OR UNDERLAP NEGLIGIBLE
p	x	x & z	x & z	x & z
q	y	y	y	y
r	x & y	x & y	x & y	x & y
s	none	z	z	z
t	—	x & y & z	—	—
u	—	y & z	—	—
v*	—	—	x	—
w*	—	—	none	—
$x=$	$p+r$	$p+r+t$	$p+r+v$	$p+r$
$y=$	$q+r$	$q+r+t+u$	$q+r$	$q+r$
$z=$	—	$p+s+t+u$	$p+s$	$p+s$

* *These are not used but are included for completeness.*

Use of Formulæ

If x, y, z and the input-output phase relationship for any of the circuits is known, it is a simple matter to deduce $p, q \ldots$ and hence the gain. Output resistance is already available in terms of y and z (Table 2).

Alternatively one can take any of the circuits and introduce restrictions, e.g. as set by the use of a particular form of chopper, thereby enabling $p, q \ldots$ and hence the gain to be expressed in terms of x, y, z. Consider for example the circuit of Fig. 5 with rectification exactly synchronous with waveform reversals, x and y being assumed to make alternately. This gives $x + y = 1$, $r = s = 0$, $p = x$, $q = y$, and the gain is $-xR_2/(R_1 + R_2)$. Simplified results of this type are listed in Table 3 for a number of practical circuit arrangements.

Smoothing Requirements

With the on-off circuit (Fig. 5), the filter R_2C_2 always receives pulsating d.c. whose amplitude is rather greater than the smoothed output. Smoothing requirements are therefore considerable, depending on the tolerable ripple, and limit the response speed.

With the s.p.d.t. circuit, however, if the load R_3 is high, and y never makes when z is made, no signal reversals are applied to the filter provided that z makes and breaks within a waveform pulse (cases 8, 9, 11, 12, in Table 3). C_2 is, in fact, adjusted once a cycle to a potential which it retains (apart from any drain through R_3) until the next adjustment. Thus if R_3 is, for example, the grid of a following valve, C_2 can be quite small. R_2 may in principle be zero, though a certain amount of resistance is in practice desirable because the a.c. waveform never has a perfectly flat top. If the output impedance R_1 of the amplifier is low, response is correspondingly fast.

Stability with d.c. Feedback

Overall negative feedback is frequently applied to chopper amplifiers to stabilize gain, and this may cause low frequency instability in some circumstances.

If the input to the circuit of Fig. 5 or 6 is suddenly changed, and the switches x and y close alternately, the d.c. applied to the smoothing circuit is delayed while the coupling capacitors charge up. This additional lag, together with that due to R_2C_2, may cause the whole system to oscillate at a low frequency via the d.c. feedback loop.

There is no additional lag if the switches close together, and this is the best arrangement. In the case of alternate

TABLE 2
Properties of Circuits

	ON-OFF CIRCUIT	S.P.D.T. CIRCUIT		
		y & z OVERLAPPING	y & z NOT OVERLAPPING	OVERLAP OR UNDERLAP NEGLIGIBLE
Gain $/A$ ($R_3 = \infty$)	$\left(\frac{rs-pq}{y}\right)\left(\frac{R_2}{R_1+R_2}\right)$	$\frac{s(r+t)-p(q+u)}{(y+z-1)(R_1/R_2)+yz}$	$\frac{rs\ \ pq}{yz}$	$\frac{rs-pq}{yz}$
Output Resistance ($R_3 = \infty$)	$\left(\frac{1}{y}-\frac{R_2}{R_1}\right)\left(\frac{R_1R_2}{R_1+R_2}\right)$	$\frac{R_1+yR_2}{(y+z-1)(R_1/R_2)+yz}$	$\left(\frac{y+z}{yz}\right)R_1+\frac{R_2}{z}$	$\frac{R_1}{z(1-z)}+\frac{R_2}{z}$

closure, the additional lag can be made very small by reducing the amplifier output impedance so that the output coupling capacitor is substantially reset each cycle. This requires the input and output switching to be approximately synchronous, and increases the current through the demodulator switch. It involves a slightly different analysis, but the results given here apply if R_1 is negligible. Alternatively it is sometimes practicable to earth the first stage of the amplifier to the other side of the input, or to inject the feedback in series with the input switch.

Some Conclusions

(1) In every case, greatest efficiency is obtained if rectification is exactly synchronous with waveform reversals.

(2) In circuits with overall d.c. feedback, x and y should preferably close together. Where this is impossible, steps may have to be taken to prevent low frequency oscillation round the d.c. loop.

(3) The on-off circuit has a gain of only $R_2/2(R_1 + R_2)$ (less than $\frac{1}{2}$) if $x = y = \frac{1}{2}$. The gain approaches unity, however, if rectification is synchronous with switching and y is made small, i.e. switching is unsymmetrical. Reducing y increases output resistance, and the useful degree of asymmetry may then depend on amplifier output impedance. On-off circuits need appreciable smoothing.

(4) The s.p.d.t. circuit needs much less smoothing and provides faster response, provided that the closure time z does not include a waveform reversal, y and z never make together, the load is high and the amplifier output impedance low. To ensure this condition in spite of possible drifts in chopper setting requires that z be adjusted to close and open appreciably within a pulse width.

(5) Provided y and z never make together, gain of the s.p.d.t. circuit is entirely independent of all resistances, being unity in the synchronous case.

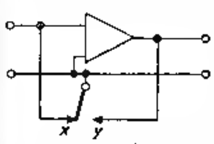
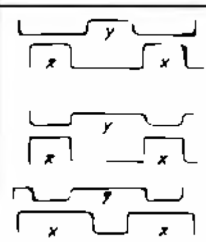
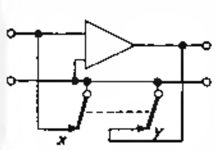
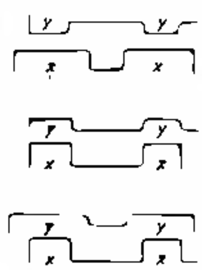
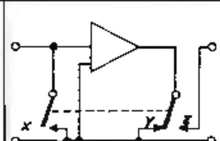
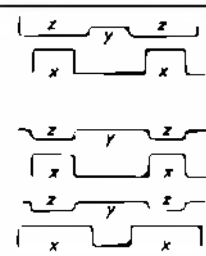
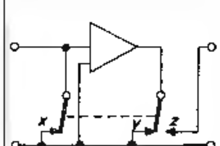
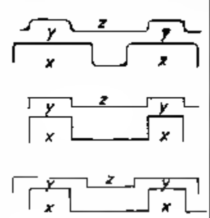
A Suitable Chopper

Switching can be effected by a mechanical chopper or by transistors used as switches[6]. There is always some wear of mechanical contacts when the voltage broken exceeds 0·1 to 0·3V, though it may be extremely slight with currents in the microamp region and voltages below about 10V. Thus the combination of a mechanical chopper for the input switching and a transistor demodulator, may sometimes be useful in proving very low drift together with maximum demodulator life, particularly when appreciable output current is required.

A mechanical chopper[7] with a contact arrangement of the type shown in Fig. 7 permits x and y to close together in the circuit of Fig. 5. Contacts CD are earthed, BC used for x, DE for y Precautions may be necessary to prevent high fre-

TABLE 3

Characteristics of some Practical Circuits

CASE	CIRCUIT AND DESCRIPTION		WAVEFORM	GAIN / A
1		xy underlap $x = y < \frac{1}{2}$		$-x \cdot \frac{R_2}{R_1+R_2}$
2		Synchronous $x + y = 1$		$-x \cdot \frac{R_2}{R_1+R_2}$
3	On-off rectification xy alternate (Fig. 5)	xy overlap $x = y > \frac{1}{2}$		$\frac{-(1-x)^2}{x} \cdot \frac{R_2}{R_1+R_2}$
4		x includes y $x + y = 1$ $(x > \frac{1}{2})$		$(1-x) \cdot \frac{R_2}{R_1+R_2}$
5		Synchronous $x = y$		$(1-x) \cdot \frac{R_2}{R_1+R_2}$
6	On-off rectification xy together (Fig. 5)	y includes x $x + y = 1$ $(x < \frac{1}{2})$		$\frac{x^2}{1-x} \cdot \frac{R_2}{R_1+R_2}$
7		xy underlap $x = y < \frac{1}{2}$		$\frac{x}{x-1}$
8		Synchronous $x + y = 1$		-1
9	S.P.D.T. rectification xy alternate $y = 1 - z$ (Fig. 6)	xy overlap $x = y > \frac{1}{2}$		$\frac{x-1}{x}$
10		x includes y $x + y = 1$ $(x = z > \frac{1}{2})$		$\frac{1-x}{x}$
11		Synchronous $x = y$		$+1$
12	S.P.D.T. rectification xy together $y = 1 - z$ (Fig. 6)	y includes x $x + y = 1$ $(x = z < \frac{1}{2})$		$\frac{x}{1-x}$

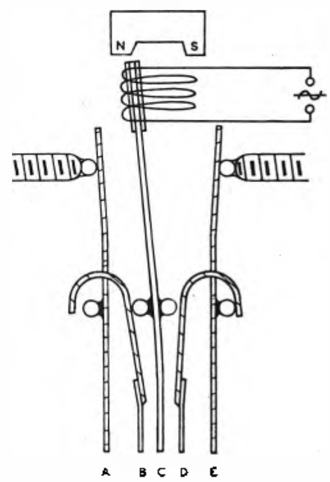


Fig. 7. Contact arrangement of a suitable chopper

quencies feeding back through stray capacities in the chopper.

The arrangement of Fig. 6, with x and y closing alternately, is met by using BC for x, CD for y and DE for z.

The overlap of y and z is inherently small in this arrangement and can be neglected in gain or output resistance calculations provided that R_8 is not made exceptionally low. It does cause a ripple spike, but this is of short duration and readily filtered out. x can be adjusted independently to give synchronous operation ($x = z$), or to encompass z by a definite small amount to ensure minimum ripple.

The xy-together arrangement can also be achieved by using AB for x, CD for y and DE for z, B and C being joined together. This is less desirable, because low-level chopping is best performed with contacts BC or CD which have a wiping action.

The voltages broken by the contacts can if necessary be made considerably smaller than the amplifier output by making y and z (Fig. 6) each fall within a waveform pulse, there being a finite (underlap) time during which neither y nor z is made. This arrangement, however, cannot be accommodated by the contact arrangement of Fig. 7.

Acknowledgments

Thanks are due to the directors of G. Kent Ltd for permission to publish the article.

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A Direct-Coupled Phase-Splitter

By C. Billington*, B.A.

A precision cathode-follower and invertor are described having an output resistance of about 3Ω, and a frequency response from d.c. to beyond 100kc/s.

DURING the course of a particular project, the need arose for a continuously variable phase-shifter to operate in the frequency range 1c/s to 30kc/s, and to permit fine setting at all angles without change of signal amplitude. The well known and convenient properties of a resistance and capacitance in series across a double-ended sine-wave source seemed to be suitable for such a device. It is important that the internal resistance of the double-ended source be negligible compared with the impedance of the phase-shifting network. This balanced source is provided by a low-impedance phase-splitter with matched output. Since the phase-splitter is of more general application, it is preferable to discuss its design and performance primarily, and to describe the phase-shifter briefly as an example of its application.

General

The requirement is for two sources of signal power of equal amplitude but opposite phase. The input resistance must be moderately high so as to impose negligible loading upon any previous stage. The average level of the input and the two outputs relative to ground should all be zero. The output resistance must be less than 10Ω. An undistorted signal level of a few volts is necessary with a gain of nearly unity and flat from 1c/s to 30kc/s.

From the outset it has been considered that any solution involving inductors would be expensive and unsatisfactory because of the wide frequency range. Consequently no serious thought has been given to the use of transformers apart from the high tension supplies. An alternative method of providing outputs at ground level is by the use of cathode-followers. The frequency response is flat within the range considered. There are two disadvantages: (a) the need for a negative high tension supply, (b) the useful output power is only about 2 per cent of the d.c. power dissipated in the cathode-follower valve and its cathode resistor. However, in the present instance these disadvantages were tolerable.

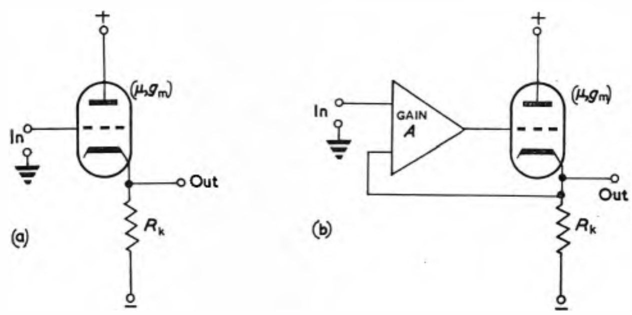

Fig. 1 (a) Direct-coupled cathode-follower
(b) Augmented cathode-follower (schematic)

Since the phase-inversion cannot be achieved by a transformer, the device falls naturally into two parts: first the low output resistance amplifier for the in phase signal, called an 'augmented cathode-follower', and second, the phase invertor. These will be described in the next two sections.

The Augmented Cathode-Follower

Fig. 1(a) shows a simple direct-coupled cathode-follower, and Fig. 1(b) a schematic diagram of the augmented cathode-follower using the same output valve under the

* *Commonwealth Scientific and Industrial Research Organization, Melbourne, Australia.*

same operating conditions. Some of the properties of these two arrangements are compared in Table 1. The mode of operation of the feedback loop does not need further elucidation, but there are certain points deserving comment. For a number of purposes in which feedback is used (particularly operational amplifiers and voltage stabilizers) the success of the circuit depends upon the loop gain being high. Here, since the simple cathode-follower already has a comparatively low output resistance (not greater than 250Ω, say) the effect of using a single stage of amplification with a gain of 50 is to reduce the output resistance to less than 5Ω. Such a low value opens up new possibilities in the accurate handling of signals of comparatively low frequency and wide band. Secondly, the two inputs of the differential amplifier may be set at the same d.c. level, so that in this application both input and output are at the same d.c. level. This is an advantage in direct-coupled systems. Thirdly, the gain of the simple cathode-follower is defined as $1 - \Delta$, and the gain of the augmented cathode-follower is calculated in terms of Δ from the usual feedback equation. The gain of the augmented cathode-follower is not necessarily as close to unity as a simple cathode-follower using valves with high amplification factor. (The 6AM6, Z77, seems to be a valve particularly suited to this purpose, having $\mu = 75$, and $g_m = 8{\cdot}5$mA/V, for a modest current drain of 8mA). In the augmented version the decrement of gain from unity is much more closely related to the loop gain than to the amplification factor of the output valve. Accordingly, a much wider range of output valves is available for the augmented cathode-follower when a medium amount of power is required with negligible reduction of signal voltage. If a gain of exactly unity is desired without the need for any lower output resistance, this can still be achieved by including a resistive attenuator between the output terminal and the differential amplifier in place of the direct connexion. Its loss can then be adjusted with the external load in circuit to make the gain exactly unity. The output resistance and the output level are negligibly affected.

A practical circuit with more than sufficient power for the phase shifting network is shown in Fig. 2. Zero-setting adjustment is necessary if the device is to be used down to zero frequency. Zero-setting by means of the common cathode resistor has some advantages over the more usual potentiometer inserted between the cathodes and the common cathode resistor: (a) the capacitive leakage to ground inherent in all ordinary potentiometers is much less serious; (b) the degeneration introduced by the potentiometer is avoided, and (c) a coarse zero-setting control can be used if desired without causing still further degeneration. The amplifier provides an a.c. gain close to 60 for various units constructed, a dozen different 12AX7's, and slightly different values of supply voltage. The dynamic range of the amplifier is about ±20V, but the limited current capacity of the output valve limits the load into

TABLE 1.

Comparison of simple and augmented cathode-follower

	SIMPLE CATHODE-FOLLOWER	AUGMENTED CATHODE-FOLLOWER
Output resistance	$1/g_m$	$1/Ag_m$
Output level	about +2V	adjustable to zero.
Gain	$1/\left\{1+1/\mu+1/g_mR_k\right\} = 1-\Delta$	$1-\dfrac{1}{1+A(1-\Delta)}$
Input impedance	high	higher

which so large a voltage can be delivered without perceptible distortion. The capacitor value was chosen so that reduction of loop gain was negligible at 1c/s. It is possible upon starting for most of the supply voltage to appear across this capacitor. In the early stages of the development 200V working capacitors were used, and frequent starting produced a partial breakdown which was noticeable only as a slight d.c. instability of the system. Capacitors of 300V rating have since given no trouble.

The effect of heater voltage variation is about one order less than with the ordinary cathode-follower. The effect of change of contact potential in the output valve is reduced by a factor equal to the d.c. loop gain (about 30), and so the unbalance between the two sides of the 12AX7 is likely to have the major effect. The measured output resistance is about 3Ω for a.c., and about twice as much for d.c. The drift of one unit without valve selection or

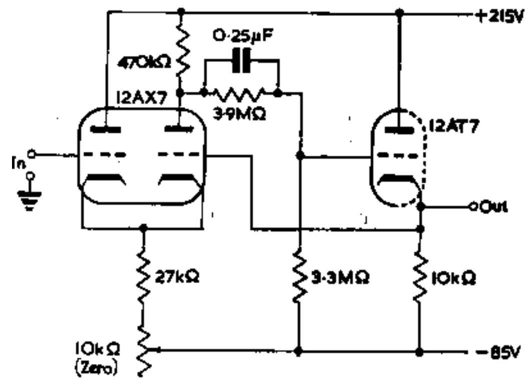

Fig. 2. Augmented cathode-follower (practical circuit)

heater stabilization was found of the order 3mV/h after thirty minutes warm up period. The overall gain for d.c. into no load is about 0·96, and for a.c. from 1c/s to 100kc/s about 0·98. The variation of gain with frequency is much less than ½ per cent, and not readily measurable with the equipment available.

Input Impedance

With the component values of Fig. 2 the grid current is about $1{\cdot}5 \times 10^{-9}$A. The potential of the common cathode is somewhat greater than one volt above ground. Variations of cathode temperature affect this bias value directly, but the grid current is altered only slightly. The differential amplifier can be made to give greater gain at the expense of a sharp rise in grid current by reducing this bias value. For present purposes, however, the values of Fig. 2 were found to be near optimum.

The input capacitance has not been measured, but since the 12AX7 operates similarly to a simple cathode-follower when viewed from the input, there will be an improvement in the case of the augmented cathode if the output valve has larger inter-electrode capacitances.

Phase Invertor

In the arrangement of Fig. 3(a) the gain

$$e_2/e_1 = -AR_2/(R_1 + R_2 + AR_1),$$

assuming the gain, A, of the amplifier to include the loss of the cathode-follower stage. If A is large, the gain approaches $-R_2/R_1$. When the resistors are equal the gain is nearly -1 which is the requirement for an invertor. In practice A is not infinite, but the gain can be adjusted to -1 by making R_2 slightly greater than R_1. The output resistance is approximately $2/Ag_m$. The change in gain caused by a given fractional change in A is proportional to $1/A^2$. For example, if A, nominally 100, declines to 75, the gain of the invertor will change from -1 to $-0{\cdot}995$.

Because of the necessary phase inversion this amplifier can be simpler than in the case of the augmented cathode-follower, and a single pentode stage is sufficient, Fig. 3(b). There are disadvantages in this simplification: grid bias has to be provided, and the correct bias for zero output is more strongly dependent upon cathode temperature. The static bias is economically achieved by the 4·7MΩ resistor which should be made partly variable if exact zero setting is required. However, since the gain of the system

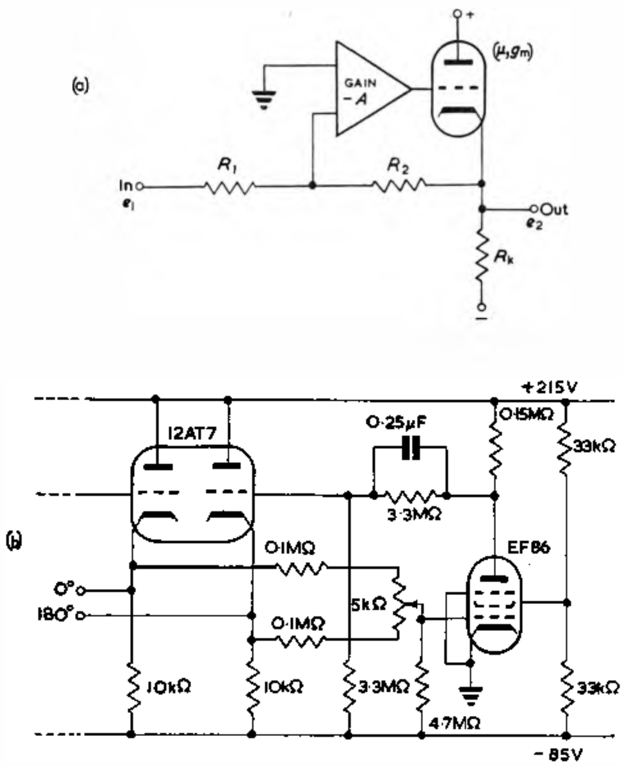

Fig. 3 (a) Inverting circuit (schematic)
(b) Invertor (practical)

referred to the pentode grid as input is about two, changes of contact potential in the pentode (due to heater voltage variation) will appear twice as great at the output.

The gain of the pentode stage is about twice that of the corresponding double triode stage, so the output resistance of the invertor is almost the same as that of the augmented cathode-follower. The frequency response is also similar. A generalized form of this circuit is treated by Thomason[1].

Power Supplies

Although both positive and negative high tension supplies are inevitable in any d.c. system operating at ground level, in this case the major part of the current flows between h.t.+ and h.t.−, and there is negligible flow to ground. Accordingly, a single h.t. supply was used, stabilized at 300V, and with an internal resistance of less than 1Ω. It was tapped at 85V with a resistive potential divider, and the tapping voltage converted to a low resistance source with another augmented cathode-follower similar to the one described. Now upon grounding this point, positive and negative high tensions were made available at +215V and −85V respectively.

Once the overall voltage is fixed the voltage of the tapping point is chosen to give the augmented cathode-follower maximum dynamic range. It will be seen that the dynamic range of the differential amplifier is increased by a greater positive high tension, but that beyond a certain value the efficiency of the cathode-follower valve is impaired by the decreasing negative high tension.

Phase-Shifter Application

As mentioned in the introduction, continuous adjustment of phase was required at all angles over a frequency range 1c/s to 30kc/s. A single resistance and capacitance phase shifting network across a phase-splitter produces at the junction a shift of $2\tan^{-1}\omega RC$ relative to the phase at the input to the resistor. The phase shift is, of course, not independent of frequency. Maximum sensitivity occurs in the region of 90° shift, but the sensitivity is inconveniently small in the region of 0° and 180° shift. A phase reversing switch which is easily incorporated between the splitter and the network, gives access to the region of 270° shift, but cannot improve the sensitivity at small angles. To overcome this difficulty two identical

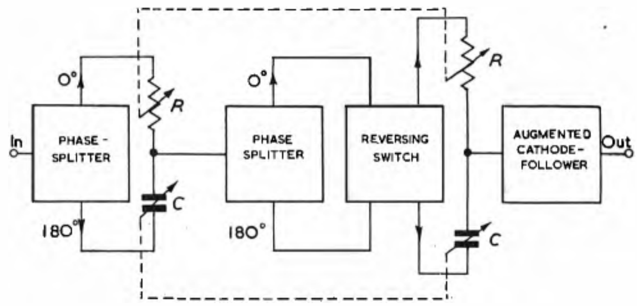

Fig. 4. Block diagram of phase-shifter

stages were used with ganged controls. The output of the second stage has to be taken to some impedance changing device, and a further augmented cathode-follower can be used. The arrangement is shown in Fig. 4.

In the unit constructed the resistor in each stage was made up of a 3·9kΩ fixed resistor and one part of a precision two-gang 10kΩ potentiometer to give continuous shift. The stepped control was provided by a rotary switch selecting any of nine capacitors from 400pF to 4μF at equal logarithmic intervals. The phase-shifting performance of the unit amply met the specification, covering in fact a rather wider frequency range. However, the invariance of amplitude with phase-shift was not as good as had been hoped. This appears to be due to the rather large capacitance associated with the precision two-gang potentiometer, which decreases the output amplitude particularly when the shift is about 90° per stage. With a precision potentiometer and matched capacitors the signal after the first stage has a phase which is close to the mean of the phases at input and output. This signal and its inverse are available at low impedance, and can be useful for calibration or display purposes. If this feature is not wanted a better performance can be obtained with a small carbon dual potentiometer in place of the precision potentiometer.

Conclusion

The two impedance changing circuits described in the foregoing combine well as a phase-splitter, and can be used separately for a number of other purposes where precision and very low output resistance is required.

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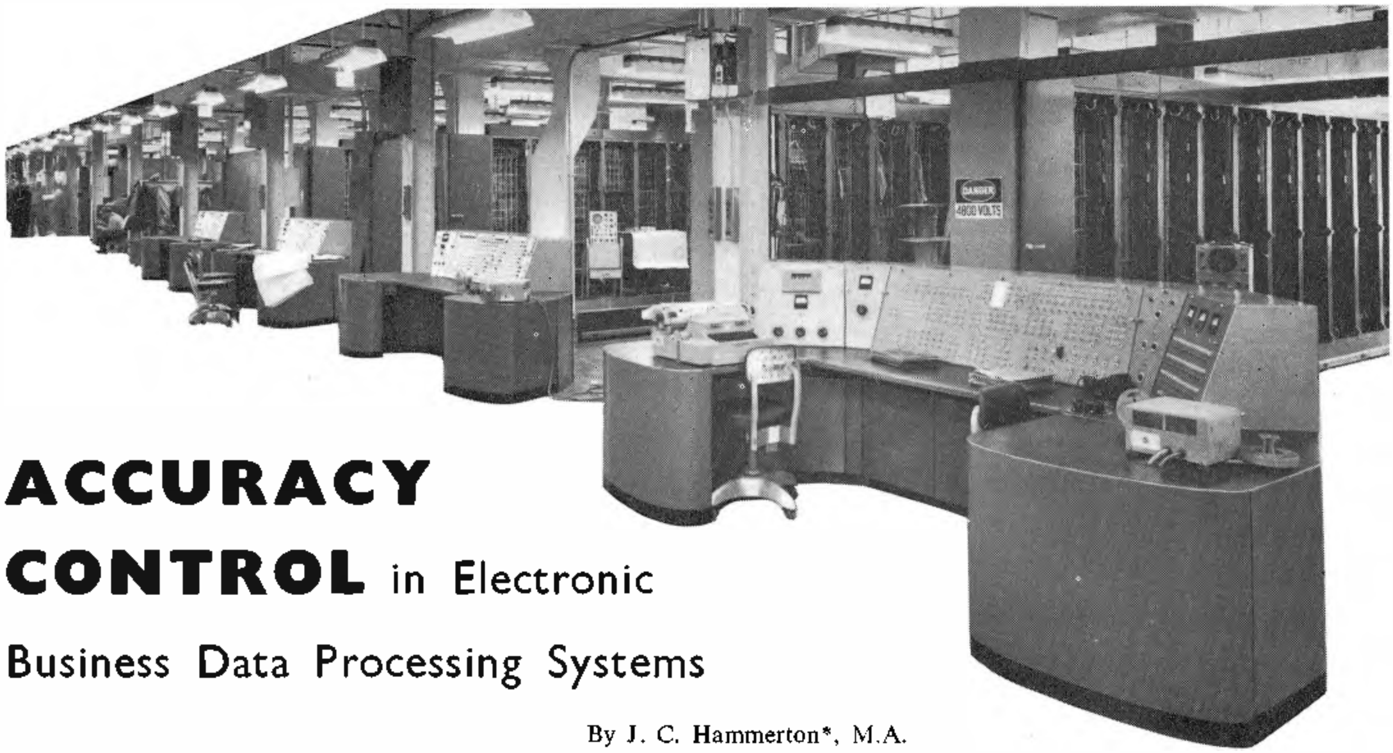


The above illustration shows a general view of the large Bizmac computer installed at OTAC, Detroit.

ACCURACY CONTROL in Electronic Business Data Processing Systems

By J. C. Hammerton*, M.A.

In business data processing systems there are two aspects of reliability. There is the initial estimate to be made of the required complement of equipment to give useful work for a certain fraction of each 24 hours and secondly there is the need to ensure that the equipment will not make undetected errors. In this article an approach to accuracy control in a large electronic data processing system is outlined.

WHEN considering an investment in electronic data-processing machines, a business executive has two major concerns in connexion with reliability. In the first place the equipment manufacturer estimates the required equipment complement on the basis of obtaining useful work for a certain fraction of every 24 hours. The business executive needs assurance that the fraction used is realistic. Secondly, he wishes to be satisfied that the equipment will not make errors without detecting them. The first concern is shared by all users of electronic equipment. The manufacturers of the larger systems seek to make good their guarantees by using the most reliable components available, by using them conservatively and by scheduling a fraction of each day for preventive maintenance. In these respects the wide experience now available from large radar and fire control systems can be applied to business data-processing equipment.

The second requirement, however, is peculiar to users of logical machines. Logical systems customarily deal with random data in which the result of an operation on one block of data bears no relation to the result of the same operation on another block of data. By contrast, in a radar system which is, for example, keeping a plot of a moving target, the accuracy of one computation can be checked by the relationship of the result to the results of the adjacent computations. If it is significantly different it can be ignored. The amount of information yielded by the system is reduced, but no error is caused. Radar systems, in fact, possess more redundancy than logical systems because their input data are not as random.

This randomness of input and output data is responsible for the feeling that a logical system may perpetrate 'unintelligent' errors which can remain undetected for a long time. Mistakes are, of course, made in business offices organized on conventional lines. In general, however, continual perusal of the information by people results in correction of the more obvious errors.

There is reason to believe that the demand for continuing reliability has reached its most exacting in the electronic business machine produced under conditions of commercial competition.

This article examines an approach to accuracy control in electronic data-processing systems.

In a future article the approach will be illustrated by consideration of a particular data-processing machine.

Engineering Approach

Three phases can be identified in the engineering of the accuracy control of a new system. These are a design phase in which the sequence of checks is designed and incorporated, a field phase in which data are collected during a period of operation, and a revision phase in which the available data are studied in order to remedy defects, if a design weakness is revealed, or to redistribute the accuracy checks, if closer monitoring of some parts of the operation is required.

The following discussion is restricted to the design phase and excludes any consideration of the virtues and value of different schemes of marginal checking. Considerable field experience of marginal checking is now available, but a decisive evaluation of its worth does not appear to have been made.

The probable causes of malfunction in a digital data-handling system can be classified as follows:

(1) Transients causing the addition of a bit or bits to the input information or the initiation of an unwanted machine action: the disturbances caused by operating relays in the vicinity of sensitive read amplifiers or by relatively rapid fluctuations in the power lines are examples.

* *Radio Corporation of America.*

(2) Loss of a bit of information or of a machine action due to an unreliable signal at some point in the system: for example, the accumulation of dirt and magnetic oxide on the reading head of a tape station produces low amplitude and, hence, unreliable signals.

(3) Loss of a bit of information or of a machine action due to marginal failure of a component: for example, a low-emission valve or a semi-conductor diode with low back resistance.

(4) Catastrophic component failure causing loss of a bit of information or of a machine action.

Of these the third cause can be reduced by a marginal checking system and a regular schedule of preventive maintenance. The fourth cause tends to become insignificant after the initial shakedown period.

With these errors in mind some general directives can be formulated about the overall system and about the individual machines in the system.

In addition, some specific recommendations can be made arising from a broad knowledge of the system and an understanding of the existing state of the art. For example, with a system which stores its information on magnetic tape, a knowledge of the speed of the tape and of the close positional tolerances imposed indicates that tape wear is a likely source of unreliability. Hence, it is reasonable at the outset to study the possibility of duplicating the information channels on the tape.

Accuracy Control in System

General directives about accuracy control in the overall system are aimed at securing reliable communication between the individual parts of the system.

Three types of checks can be distinguished:

(1) Feedback check.
(2) Parity check,
(3) Verification check.

In the feedback or 'echo' check the receiving machine produces a version of the signal received from the transmitting machine and returns it to the transmitter. The original output signal is stored in a register at the transmitter. The echo signal is used to reset the register, thereby establishing that the signal reached the receiver. The logic at the transmitter includes a method of inspecting the register to see if it has in fact been reset.

Resetting the register detects the loss of a unit bit in the communication channel. By triggering the register with the echo signal, it is possible to detect the addition of a unit bit in the communication channel as well as the loss of a unit bit.

The parity check involves the increase of redundancy in the system's information. For example, in a system using n parallel bits per character an extra bit is added in an $(n + 1)^{th}$ channel in order to make the total number of unit bits in each character an even number (even parity) or an odd number (odd parity).

The adoption of even or odd parity is a matter for discussion. An important consideration is that owing to a large voltage transient in the vicinity of sensitive 'read' amplifiers, all of the parallel information channels may accumulate unit bits. In order to detect this spurious character even parity is adopted if $(n + 1)$ is an odd number and odd parity if $(n + 1)$ is an even number.

The feedback and parity checks are complementary methods of monitoring a communication channel. Their respective merits are summarized below:

(1) In general, a parity check only detects the addition or omission of one unit bit from a character.

(2) An echo check can be designed to monitor all the information channels individually.

(3) The terminal equipment required in both cases is comparable.

(4) A parity check necessitates an extra information channel if the characters are presented in parallel. The echo check requires either time sharing of the information channels or duplication of them.

(5) The logic of an echo checker is located at the transmitter while that of a parity checker is at the receiver.

The latter consideration is important where one machine requires two-way communication with any one of a large number of other machines. For example, a computer may require connexion to a large number of magnetic tape machines either to 'read from' them or 'write to' them. It is evidently economical to avoid duplication of equipment by locating all the checking logic in the computer. Therefore, when the computer transmits, an echo check is used, but when the computer receives signals from the tape machines, a parity check is used.

A verification check is one in which an operation is performed twice and the two results verified to see that they are in agreement. This type of check is sometimes used to monitor arithmetic operations in a computer. It is most valuable if the original and the verifying operations are performed by different equipment.

Verification checks are required when the form of the information, i.e., the coding, changes between two parts of the system making feedback checks impracticable.

In general, the preparation of new input data by manual methods necessitates verification. For example, in preparing punched paper tape from typescript by means of a keyboard machine, an operator makes an original tape. A different operator (or possibly the same one) then prepares the final tape on another machine. The second operator's selection of keys is compared character by character with the punched characters on the original tape. Any discrepancy causes his keyboard to lock until he ascertains the reason for the discrepancy.

If the keyboard operators make one error every n characters on average, the chance of them making an error at the same character is $1:n^2$. The chance of them making the same error at the same character is less and depends on the type of error. If, for example, the error involved is that of hitting an adjacent key, eight different errors are possible. If the probability of any error of the eight being made is equal to that of any other, the chances of two errors being the same is 1:8.

Hence, the chance of two operators making the same error in the same character is $1:8n^2$. If $n = 1\,000$ which is equivalent to the operator making two errors in a double-spaced page of typescript, the overall input operation results in one error every 8 million characters. A reel of magnetic tape 2 400ft long contains approximately 3 million characters. Hence, the manual input operation is responsible for one error every three reels.

Accuracy Control in Machines

In a large system the design of the individual machines is carried through in detail by many engineers. The formulation of a philosophy of accuracy control within machines is, therefore, necessary to secure intelligent uniformity in the use of accuracy checks.

As previously stated, machine errors are expected to be either of a transient, and hence correctable, nature or catastrophic, and hence, irremedial without maintenance action.

The objectives of accuracy control are, therefore:

(1) To determine that an error exists.
(2) To correct the error if possible.
(3) To shut-down if the error is uncorrectable and indicate the type of error.

Since shutting down a machine is expensive in operating time, error correction routines are very important. Hence, also the primary purpose of error detection is to determine that an error of some sort has occurred. Identification and indication of the type of error only become important after attempts to correct the error have failed and the machine is shut down.

Further, when a shut-down is necessary, it is desirable to effect this at a convenient point in the operation so that the operation can be continued after maintenance action. For example, if the operation consists of a sequence of logical actions performed repetitively on blocks of data, a shut-down immediately before the block of data containing the error would be considered convenient.

Error Detection

In general the operations performed by logical machines are based on a sequence of actions in which any one action is dependent on the result of the preceding one. Some of these actions are paralleled by others. However, it is usually possible to locate a number of points throughout the machine at which checks will cover all the actions regardless of the input data.

The minimum number of checks necessary to monitor the operation can be established from a critical appraisal of the machines logic. If more equipment is available, increasing the number of checks beyond the minimum yields returns in swifter isolation of the cause of the error.

Customarily it is not possible to ensure a 'foolproof' system of accuracy control. It is always possible to envisage some combination of events which will produce an unwanted and undetected result. This is difficult for some non-scientific people to understand. Nevertheless it is advisable as a matter of practical engineering to recognize it.

With this in mind the positioning of error checks in a machine depends on the following considerations:

(1) The probability of occurrence of an error of a given type.

(2) The consequences of failure to detect an error.

It appears desirable in principle to position checks where errors occur most frequently either because of the type of data employed or because of the quantity of equipment covered by the check. However, it is equally important to consider the consequences of failure to detect an error. If, for example, a certain machine failure could lead to loss of information from the system, monitoring the equipment associated with the failure is important no matter how few components are involved.

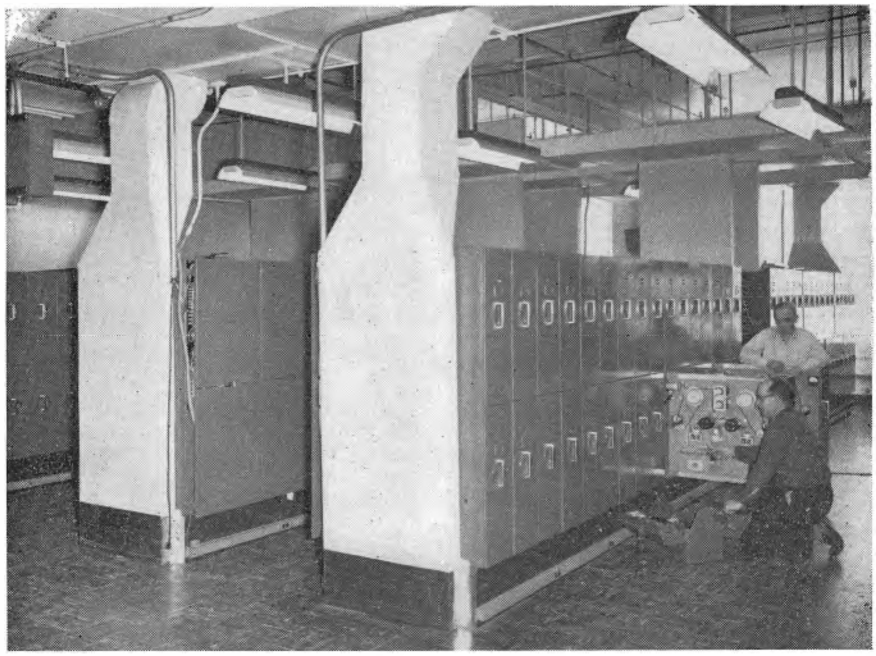
To illustrate this line of thinking with respect to the type of data, an arrangement of data is considered in which the characters are grouped into "blocks'. The beginning and end of a block are defined by special control symbols, a 'start symbol' and an 'end symbol' respectively. The characters in each block have six information bits. A seventh bit is added in order to give each character even parity. Recognition of the characters depends on all seven bits being correct.

A probable type of malfunction in a machine is a parity error arising either from picking up an extra bit or from losing a bit Consequently all characters are checked for correct parity.

Since the control symbols are used to derive orders to the medium storing the input and output data, e.g. magnetic tape, special problems arise when a parity error occurs in one of them.

If a block of data contains on average 200 characters, the probability of a parity error in a control symbol is 1 per cent of the probability of a parity error in an information character.

As an illustrative figure suppose that a parity error occurs once per reel of tape which is once every 3 million characters or once every 6 minutes. A parity error in a control symbol, therefore, occurs once every 10 hours. This is too often in terms of the time required to unravel the after effects of the error if the machine's logic is not designed with this type of error in mind.



The banks of tape stations installed at OTAC

If parity errors occur in successive symbols tape control is further complicated. The probability of this occurring can be estimated as follows:

Let random parity errors occur at an average rate of one error every n characters.

On average, therefore, there are two errors in $2n$ characters. The two errors may be related positionally in any one of p different ways where.

$$p = \frac{2n\ (2n - 1)}{1 \times 2} = n(2n - 1)$$

Since the errors occur randomly by definition, the probability of two consecutive errors occurring in a particular manner is $n(2n - 1)$ to 1.

Consequently, since the probability of an error occurring is $n:1$, two consecutive errors occur in a particular manner once every N' characters where,

$$N' = n^2(2n - 1)$$

If each block of data averages y characters, there are $2n/y = X$ blocks in $2n$ characters and, hence, X pairs of control symbols.

Therefore, there are $2X$ ways in which two parity errors could prevent adjacent control symbols from being recognized.

The probability of this occurring is once in every N characters where,

$$N = n^2(2n - 1)/2X$$
$$= n^2(2n - 1).y/4n$$
$$= n(2n - 1).y/4$$

if $2n$ is much greater than unity,

$$N \rightarrow n^2.y/2.$$

For example, if the rate of occurrence of parity errors is once per 100 000 characters or 30 parity errors per reel of tape, the probability of getting parity errors in adjacent control symbols is $N:1$ where

$$N = 10^{10} \times 200/2 = 10^{12},$$

i.e., once every 300 000 reels of tape.

At 6 minutes per reel and 15 hours system operation per 24 hour-day, 300 000 reels can be run serially in 2 000 days or, approximately 10 years!

Considering the normal rates of random parity errors encountered it appears wise to include means for retaining control in the absence of one control symbol in a block of data but satisfactory to ignore the consequence of errors in consecutive symbols.

In considering machine failures the probability of a failure in any section of the machine is directly dependent on the number and type of the components involved. This general statement assumes that the equipment in the machine is designed consistently to the same operational ratings.

Nevertheless, it may be thought advisable in the design phase to establish a closely integrated series of checks in a certain crucial area of a machine. This conclusion can be reached, for example, by assessing the difficulties of servicing the machine in this area. However, if during the field phase the failure rate proves to be low, the checking equipment is not being used efficiently. Hence, the revision phase includes a re-evaluation of the use of the checking equipment with a view to its redistribution.

Despite these considerations the consequences of not detecting an error must always be taken into account. This is of particular importance where inferential logic is used.

For example, consider a machine which handles three sources of input information, F, A and B. The operation performed by the machine is such that a large proportion of the total input data is derived from source F. The logic of the machine includes rules governing the selection of sources A and B.

The selection of source F, however, is inferred from the following conditions:

(1) The machine is operating, i.e. it has been started and it has not been shut-down for any reason.

(2) Since it is operating, it must require input data.

(3) Neither source A nor source B has been selected.

These conditions are sufficient to justify the selection of source F.

Evidently the equipment required to select F in this way is very simple compared with the equipment required to select it positively in every case.

However, it is noted that if the operation requires the selection of source A but, because of a machine failure, source A is not selected, source F is selected erroneously instead. The inclusion of equipment to monitor the selection of source F depends on the probability of failure in the equipment for selecting sources A and B and the operational consequences of selecting source F erroneously.

The final word on the most efficient method of using a given quantity of accuracy control equipment can only be said after some field experience with the machine. In an industry using rapidly developing techniques, it is not usually possible to apply field experience from one machine to its successor unconditionally.

It must be admitted, therefore, that in a new machine employing new techniques much of the initial work on accuracy control tends to be based on theory alone. This statement admits the inadequacy of theory but emphasizes its importance.

Error Correction

The importance of error correction routines lies in the transient nature of the principal sources of error. The accumulation or loss of unit bits of information form a major part of all the errors in the system. In these cases repetition of the relevant part of the operation may be accomplished without further error. In one machine errors of this type are believed to account for approximately 97 per cent of all the errors which occur.

The usual method of error correction is to repeat a selected part of the operation. For this purpose the operation has to be analysed into a basic cycle, e.g. an instruction, and a close check kept on the movement of tapes within each cycle. The objective of this is to know which tapes to back up and how far to back them when an error occurs. This objective is not realized easily in the case of a machine which controls many tapes or which uses interleaved instructions.

The value of error correction routines is expected to be two-fold. Firstly they enable truly random transient errors to be handled without a shut-down. Secondly, and of equal importance, efficient error correction permits some relaxation of tolerances. For example, the guiding of magnetic tape over the reading head of a tape station is a critical operation. Relaxation of tolerances may lead to failure to read correctly due to 'bounce' or 'skew'. If the error correction routines can be relied on to make good any such errors they do not become a serious problem until their rate of occurrence makes the time spent in correcting them significant.

Error Indication

When a machine tries to correct an error but fails, it shuts down. At this time the type of error is indicated as an aid to maintenance personnel in assessing the cause of the shut-down.

The importance of error indication in getting a machine back into operation needs no emphasis. If the shut-down is attributable to a machine failure, the machine can be serviced quickly. If the cause of the shut-down can be assessed as non-repetitive, e.g. an error in the information on the input magnetic tape, the error can be over-ridden and the operation continued. For this purpose the provision of facilities for processing a block of data regardless of errors requires careful consideration. This implies that the machine can proceed one instruction at a time under the control of an operator.

Conclusion

This article has attempted to outline an approach to accuracy control in electronic business data-handling systems.

The approach consists of securing reliable communications between the machines which compose the system, and of obtaining a uniformly intelligent use of checking equipment within the individual machines.

In a future article the application of this approach to a particular data-processing machine will be described.

Acknowledgments

The author is indebted to the Radio Corporation of America for permission to publish this article and to his associates for their constructive criticisms.

Testing the Linearity of Modulators and Demodulators in Multi-channel F.M. Transmitters and Receivers

By G. C. Davey*, B.Sc.

This article describes a newly engineered technique for rapid adjustment of the linearity of f.m. modulators and demodulators, where the linearity must be of a high order. A cathode-ray tube display clearly discriminating changes in slope as little as 1 per cent is used.

The description is followed by the principles of design of an equipment performing this function as well as facilitating the alignment of i.f. amplifiers.

ONE of the requirements of modern f.m. carrier telephony systems, in order to prevent crosstalk between channels, is the highest possible degree of linearity of the modulation and demodulation circuits. For the convenience of both design and maintenance engineers it is desirable to provide a means of easy and accurate visual alignment on the lines of the well-known swept oscillator techniques used to adjust r.f. band-pass amplifiers.

Such conventional sweep generators, which plot the voltage output of the circuit under test to a frequency base, do not provide sufficiently sensitive indication of demodulator linearity. A much more sensitive advancement

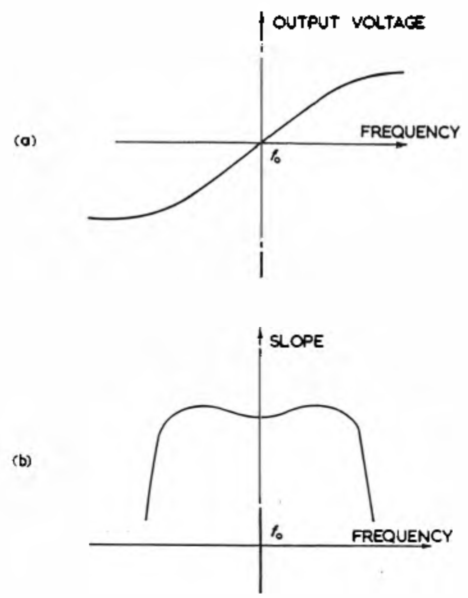


Fig. 1. (a) F.M. discriminator characteristic
(b) First derivative of (a)

of this idea has been suggested by Post Office engineers. This is a method of displaying the slope, i.e. the first derivative, of a demodulator characteristic to a frequency base. A comparison of the two displays is given in Fig. 1.

This article describes an equipment which directly tests demodulators in this way—and indirectly tests modulators. It may also be used as a conventional sweep generator, when it must have an unusually good performance, because the intermediate frequency amplifiers of multi-channel receivers must be adjusted within close tolerances, both as regards flatness of frequency response within the pass band, and rate of attenuation outside.

The Derivative Test

A display of the first derivative of a discriminator characteristic is obtained by applying to it an r.f. signal swept at a low display frequency (F_1) through its normal operating hand, and frequency modulated incrementally at a much higher frequency (F_2) over a much narrower band. Thus the high frequency (F_2) sweep is superimposed on the display sweep (F_1). Considering any instant during the display cycle, reference to Fig. 2 shows that, at that instant the amplitude of F_2 in the output from the discriminator is proportional to the slope of the characteristic.

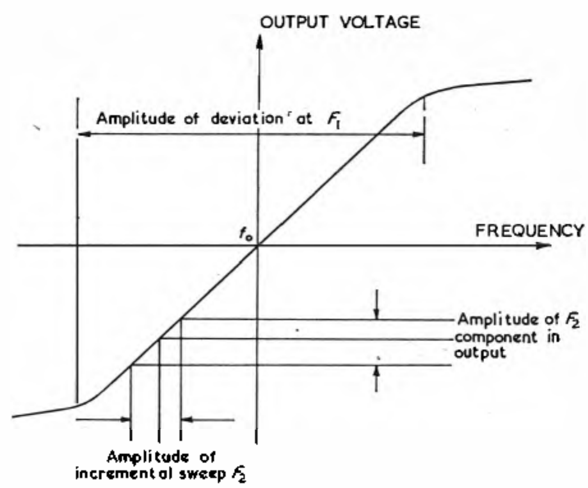


Fig. 2. F.M. discriminator charactistic showing principle of derivative test

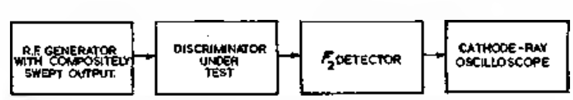


Fig. 3. Elementary block diagram of derivative test

Thus the output from the demodulator will contain a component of frequency F_2, whose amplitude will vary, during the cycle of F_1, proportionally to the slope of the characteristic. If this amplitude variation is detected, for example by a diode rectifier, and amplified, then it can be displayed on a cathode-ray tube.

A block diagram of the equipment thus far conceived is given in Fig. 3.

If $\pm\ \delta f_1$ is the deviation at frequency F_1, and $\pm\ \delta f_2$ is the deviation at F_2, then in order to approach the ideal it is desirable that F_2 be much greater than F_1, and δf_2 much less than δf_1. Then the extent of the characteristic taken in by one cycle of F_2 is reduced, and an analogy can be brought with the differential calculus, in which the increment approaches an infinitesimal limit, and the differential more closely equals the rate of change of the dependent quantity.

* *Marconi Instruments Ltd.*

Design Considerations

For this test to be successful, the deviation at frequency F_2 of the generator output must be constant over the display period. Any variation in it will of course be reproduced in the display.

In particular, if f_o is the centre carrier frequency of the demodulator band, then it is not possible to produce the compositely modulated signal by frequency modulating a signal generator of frequency f_o simultaneously with the two frequencies F_1 and F_2. Any known modulator would pre-modulate the amplitude of F_2 according to its own non-linear characteristic, and its own derivative characteristic would be displayed. The reason for this is shown by Fig. 4, in which equal F_2 modulation amplitudes are considered at two points within the main display sweep.

The solution to this problem is to modulate two oscillators of frequencies f_1 and f_2 respectively with the frequencies F_1 and F_2, with the desired amounts of deviation. If f_1 and f_2 are such that $f_1 \sim f_2 = f_o$, then the two can be mixed to produce the frequency f_o with the composite

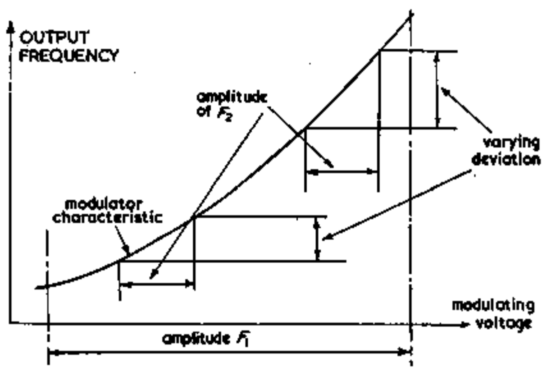


Fig. 4. Ferrite modulator characteristic, illustrating effect of non-linearity

modulation. Since, as regards frequency, mixing is an additive process, the output will have the same deviation as each of the primary oscillators. A block diagram illustrating the proposed system is given in Fig. 5.

Considering the problem of producing the wide, low frequency, deviation ($= \pm \delta f_1$), it is desirable that f_1 be greater than f_o, so that though the initial deviation is a small proportion of its parent frequency, f_1, it is a large proportion of the output frequency, f_o.

The mixing process inherently encourages harmonics, and it is possible that the harmonics mf_1 and nf_2 will beat to produce output within the limits of the operating band, $f_o + \delta f_1$, and $f_o - \delta f_1$. To make m and n as large as possible, and hence mf_1 and nf_2 as weak as possible, it is desirable that f_1 and f_2 be much larger than f_o. As pointed out above, this would carry with it the advantage that δf_1 need only be a small proportion of f_1.

Modulators

The test generator described above provides, in effect, a very linear modulator. In order to test modulators, it would be logical to design a very linear demodulator, or simulation thereof. This is quite possible, but would involve the provision of an extra set of test equipment, with added expense and portability problems. A satisfactory solution is found by testing first a demodulator, taken from the receiver rack, and using it as a standard of linearity with which to compare the transmitter.

At a link repeater then, inevitably possessing both modulator and demodulator, it is only necessary to modulate the transmitter with a composition of the two frequencies F_1 and F_2 defined above, with the appropriate amplitudes, and then feed a suitable part of the output to the receiver. As in the demodulator test, the component of F_2 in the output of the tested demodulator is detected, and its amplitude displayed, when, if the modulator is perfect, a replica of the demodulator characteristic will appear. An attempt can be made to correct any deviation from the original by adjusting the modulator, the accuracy limit being set only by the linearity of the derivative test generator, which is very good.

The only additional equipment necessary to facilitate this test is a circuit mixing the two frequencies F_1 and F_2 with suitable amplitudes.

Design of the Sweep Generator

The more difficult aspects of the present design were set by the specification of the conventional sweep generator.

This, with a band centre of 70Mc/s, produces an output of 500mV into 75Ω, flat within 0·1dB from 50Mc/s to 90Mc/s. The sweep width required for derivative tests is only $\pm$5Mc/s.

The display frequency, F_1, is 50c/s, and the incremental frequency F_2, is 20kc/s.

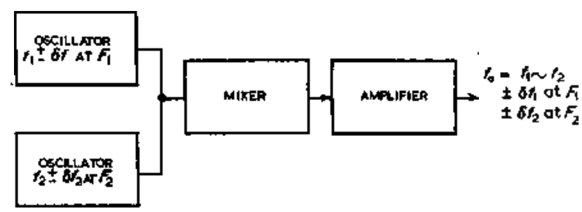


Fig. 5. Elementary block diagram of sweep generator

For the reasons given above, f_1 and f_2 should be of the order of thousands of megacycles per second. This suggests the use of klystrons, which lend themselves to frequency modulation, but which are inconveniently bulky for use in portable equipment, and have poor frequency stability.

The alternative is to bring f_1 and f_2 down to a frequency at which lumped circuit techniques are possible, and oscillators in the v.h.f. band are employed. The centre frequency of that with wide deviation is 200Mc/s, swept $\pm$10Mc/s. The other oscillator has a centre frequency of 470Mc/s. It will be seen that the second harmonic of the former, encouraged by the crystal, will beat with the latter to give a difference frequency of 70Mc/s, $\pm$20Mc/s. This gambit of employing the second harmonic of the wide sweep oscillator reduces the deviation requirements, and gives rise to only insignificant spurious outputs in the working band.

Both oscillators are frequency modulated magnetically. Part of the tuning inductance of each is wound on a ferrite core, which is subjected to the magnetic field of another coil. The flux in the ferrite is controlled by the current in this modulating coil, which is connected in the anode circuit of a modulator valve. The inductance of the oscillator coil, and hence the output frequency, are controlled by the flux in the ferrite.

Both modulation waveforms are sinusoidal. In the case of the wide sweep, this is much simpler than attempting to establish a sawtooth current in an inductor. This requires that a sinusoidal voltage of the correct frequency and phase must be applied to the X plates of the display tube.

The wide sweep oscillator in particular suffers a certain amount of amplitude modulation. In order to minimize the effect of this, the amplitude fed from the other oscillator to the mixer is made considerably the weaker. The output from a mixer can be shown to be primarily dependent upon the weaker of the two inputs to it. Thus a tolerably level output is fed from the mixer to the wide-band amplifier. The mixer is a germanium crystal.

The wide-band amplifier, which uses double-tuned-stagger-tuned interstages, has a voltage gain flat within 1dB over the range from 50 to 90Mc/s when correctly loaded in 75Ω.

A buffer amplifier stage is placed between the weak oscillator and the mixer. Since the output of the equipment is closely controlled by the input from this oscillator to the mixer, the buffer stage is ideal as a means of output level control, and automatic amplitude control. A germanium diode detector connected to the main amplifier output develops an a.g.c. voltage which is amplified by a d.c. coupled amplifier and fed back to this buffer amplifier. By this means, the output of the equipment is held flat within 0·1dB. The same germanium crystal is used to indicate output level on a panel-mounted meter.

The low output impedance of the a.g.c. controlled equipment is corrected by means of an attenuating pad, introducing a 6dB loss.

Provision is also made by which the a.g.c. voltage may be derived from the point of application of the signal. A small artificial section of cable is provided, with a germanium crystal voltmeter connected to the centre conductor, which may be connected in series with the output at the end of a long cable. The voltmeter output may be employed as a source of a.g.c. voltage, rendering the output flat at the input to the equipment under test.

Fig. 6 is a block diagram of the sweep generator thus far discussed.

Output level control is achieved by adding a d.c. component to the a.g.c. voltage.

Calibration

It is desirable that some means of amplitude calibration of the display be provided, so that equipment under test may be adjusted within its specification.

For conventional sweep tests, provision is made for square wave modulation of the output amplitude, so that it falls by 1dB during alternate half cycles of the display. Thus the return trace is at a lower level than the forward trace. This 1dB calibration will hold good so long as the equipment under test is linear at the test level.

In order to avoid tilt problems, the modulating waveform is obtained from a vibrating relay, driven at 50c/s.

As the a.g.c. system is not linear, it is necessary to adjust the 1dB step for each output level. This can be done by switching the relay over with a d.c. voltage, and setting up the step by reference to the output level meter.

A very similar type of calibration is provided for the derivative display. Since the output from the demodulator is proportional to the amplitude of the incremental deviation, it is only necessary to square wave modulate the 20kc/s deviation in a similar manner to the above in order to obtain just the same type of display. In this case the 20kc/s level applied to the 470Mc/s oscillator is lowered in every other half cycle of the display by 1dB. This is done by switching in and out, with another vibrating relay, a pre-adjusted 1dB attenuating pad.

In this case a change of 1dB signifies a 10 per cent change in slope.

For both displays the sensitivity of the display unit is sufficient to provide a one inch deflexion between the two traces 1dB apart. Thus differences of 0·1dB or 1 per cent of slope are discriminable.

For frequency calibration, brightened spots on the display are provided at 1Mc/s intervals. These are crystal controlled, and to facilitate counting, every fifth one is omitted. In order to locate the band centre, unique markers are provided at 70Mc/s.

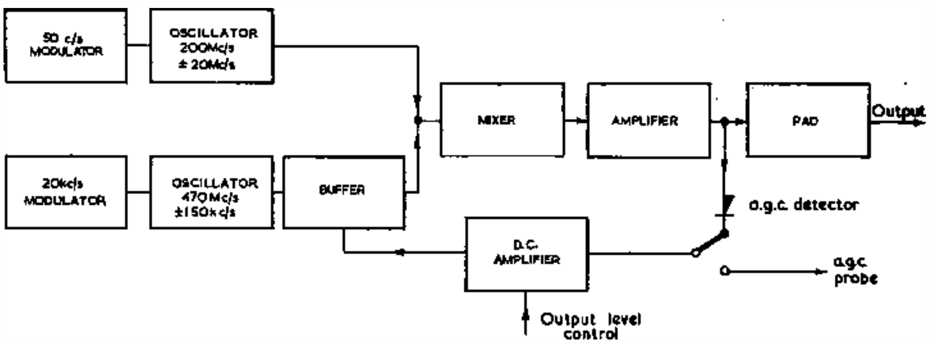


Fig. 6. Complete block diagram of sweep generator

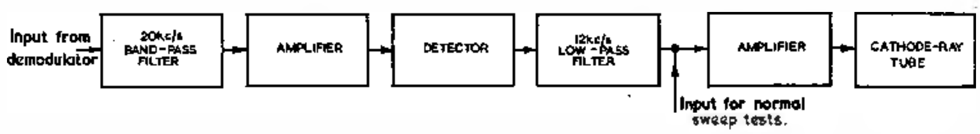


Fig. 7. Block diagram of display unit

The Display Unit

In the equipment designed, a display unit separate from the sweep generator is provided. This unit includes the 20kc/s detector. A block diagram is given in Fig. 7.

The first stage is a 20kc/s band-pass filter, which selects the 20kc/s component in the demodulator output. This also narrows the effective bandwidth of the unit and reduces noise. This is amplified conventionally and detected by a voltage doubling diode. A low-pass filter removes any residual 20kc/s, and the envelope, which consists of square waves when the 1dB calibration is in use, is amplified by a low frequency amplifier, designed for minimum tilt, and displayed.

Some care in the filter design is necessary to prevent tilt on the square waves.

Only the latter amplifier stages are necessary when the equipment is used for conventional sweep tests.

Also included in this unit is a valve mixing a 50c/s signal with one of 20kc/s used for the testing of modulators. The sweep generator is not, of course, necessary for such a test.

Acknowledgment

The author wishes to thank Marconi Instruments Ltd for permission to publish the article.

Efficiency-Diode Scanning Circuits

(Part 1)

By K. G. Beauchamp*, C.G.I.A., A.M.Brit.I.R.E.

The post-war expansion of television coverage in this country, coupled with the demand for bigger and brighter pictures, has led to the development of new economical techniques in receiver design. One major factor contributing to success in the application of these techniques has been the evolution of 'resonant-return' or efficiency-diode scanning circuits. The basic theory behind this development is described and methods shown of solving problems arising out of the design of modern line-scanning circuits.

THE origin of the development of efficient line scanning circuits is now universally recognized as the publication of patents by Blumlein in 1932 giving details of a basic energy recovery circuit[1]. Little use was made of this at the time, however, as the cathode-ray tubes then in use were mainly designed for electrostatic deflexion, and those tubes requiring electro-magnetic deflexion of the electron beam operated at such low accelerating potentials (e.h.t.) and small scanning angles, that the increase in efficiency by the use of these methods was hardly considered justifiable. In the immediate post-war period, however, American receiver manufacturers began to make use of Blumlein's invention and with the appearance of the first British a.c./d.c. domestic television receiver[2] the system became quickly established in this country. Soon afterwards came the advent of large picture-area tubes having scanning angles of 70° and 90° and demanding e.h.t. potentials of the order of 12 to 20kV. These tubes rendered the use of an energy-recovery line scanning circuit essential if economical receiving equipment was to be produced, as may be seen from the following example.

Consider a typical modern receiver where a 14in diagonal, 70° scanning-angle tube is required to be scanned with an applied e.h.t. of 14kV. A scanning coil of 10mH inductance may be chosen as representative of current practice, and measurements with such a coil will show that a sawtooth scanning current of about 1A peak-to-peak is required in order to adequately scan the tube. The energy stored in the scanning coils at the end of the scan period is given as:—

$$W = \frac{L_y . I_y^2}{2} = 5\text{mJ} \quad \ldots\ldots\ldots\ldots \quad (1)$$

Now in the absence of any energy recovery arrangement this energy will be dissipated in a resistance network and supplied afresh with every cycle of operation. Thus the total power required P from the scanning circuit with the British 405 line system is:—

$$P = W . f = \frac{5 \times 10\,125}{1\,000} = 50\text{W} \quad \ldots\ldots\ldots \quad (2)$$

where:—

f = the frequency of sawtooth repetition in c/s.

Actually due to losses in the scanning transformer, valves, etc., the power requirements from the receiver h.t. supply would be greater than this and form a disproportionate part of the total receiver power consumption. It is worth pointing out at this juncture that using a.c./d.c. techniques, as is now almost universal, then an h.t. potential of only 170 to 200V will be available, and to develop 50W of scanning power from such a low voltage source would necessitate mean anode currents of about 300mA for the scanning valve(s)!

The major justification for the use of efficiency-diode circuits then, lies in their economy of power consumption. This is achieved by conserving the energy which would otherwise have to be dissipated at the end of each scanning cycle and returning this either to the supply source or to a storage capacitor for use during the subsequent scanning period. Thus two variations of the circuit may be considered, and as these are characterized by the manner of diode connexion as will be seen later, they may be differentiated by the prefix 'shunt' or 'series' applied to them.

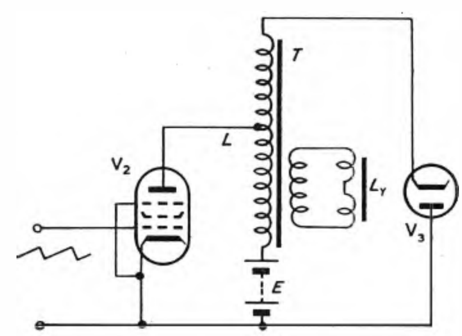


Fig. 1. Basic shunt efficiency-diode circuit

Shunt Efficiency-diode Connexion

Blumlein's original circuit falls into this category and a version of this is shown in Fig. 1, modified to include a matching transformer *T*. With this arrangement the pentode is rendered conductive for rather more than half the scanning period by the application of a suitable waveform to its grid. This may be of pulse waveform, but for reasons given later is usually a sawtooth having a peak-to-peak amplitude of about twice the valve grid-base. During the period of V_2 conduction the current rise in L_Y is a substantially linear one, due to the high ratio of inductance to resistance (L/R) reflected into the primary winding. This current rise ceases when V_2 is cut-off by the large negative potential applied to its grid. The high-Q resonant circuit formed by L and its associate stray capacitance then oscillates and the current in L_y quickly falls to zero, reverses in direction, and builds up to a maximum in the opposite phase.

The potential set up across the primary winding is now in such a direction as to cause diode V_3 to conduct; the current through the transformer, and hence L_Y, falling to zero in a linear manner controlled by the action of the diode. With the scanning current in L_y reduced to zero V_2 is again rendered conductive and the cycle recommences; the complete scanning period being shared by the almost

* *Associated Electrical Industries Ltd, formerly with General Electric Co. Ltd.*

equal consecutive conduction periods for the two valves.

In order that the potential across L_y may be maintained constant during the entire scan period and therefore a linear rise of current obtained, it is only necessary to adjust the transformer ratio between V_2 and V_3 so that equal peak potentials are developed around the two conduction paths.

During the period of diode conduction (the first part of the scan period) current will flow through the supply source E in the reverse direction to that normally supplied by such a source. This may be termed the 'charging current' and for a circuit having zero losses then this will be equal to the initial current supplied by E. It can be shown that a limiting condition exists where zero current is drawn from a source of finite potential in order to maintain a sawtooth current waveform in the scanning coils.

A practical circuit is given in Fig. 2 which is designed to scan the tube postulated earlier. This is a feedback circuit where the sawtooth controlling waveform for V_2 is obtained from an integrating circuit CR connected across the h.t. supply. The repetitive discharge of capacitor C, at a rate determined by C_1R_1, being effected by a low-impedance triode V_1, and rendered conductive when a positive pulse is applied to its grid from the output transformer T. Adequate scan can be obtained from this circuit with a total anode current for V_2 of only 10mA, supplied from a source potential of 400V.

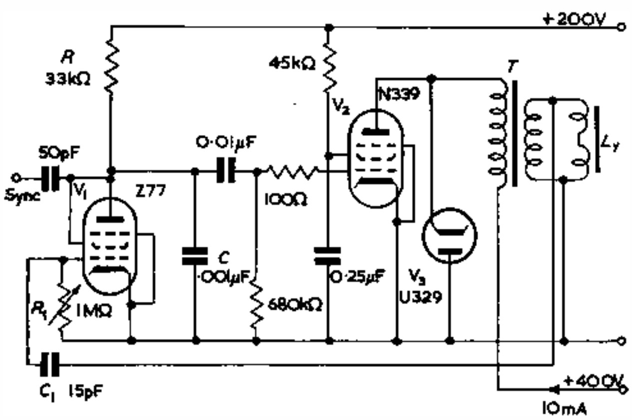


Fig. 2. A practical shunt efficiency-diode circuit

Series Efficiency-diode Connexion

If the energy remaining in the magnetic field at the collapse of scan is returned, not to the supply source as in the shunt circuit, but to a capacitor C_1, then the potential developed across this capacitor may be added to the h.t. supply for V_2. This gives the series condition shown in Fig. 3. It can be shown that in the limiting case of zero losses then a finite current will be drawn from a supply of zero potential in order to maintain a sawtooth current in the scanning coils. In a practical case some losses are, of course, experienced, but nevertheless this series arrangement does enable operation to be secured from a low value of h.t. potential. This is an important factor in the design of domestic a.c./d.c. television receivers and is responsible for the almost universal use of series efficiency-diode circuits in present-day receivers.

Tourshou 'Direct-drive' Circuit

With the Blumlein series circuit to be described in more detail later it will be seen that the coupling transformer L, shown in Fig. 3, is an essential feature of the arrangement. This matches diode and pentode currents enabling linear and stable operation to be secured. This limitation need not apply to all resonant-return systems however, and by adopting rather different methods of energy control Tourshou[3] has produced a simple circuit having some advantages over that due to Blumlein.

The basic Tourshou circuit is shown in Fig. 4. No coupling transformer is included and the scanning coils are inserted directly in the anode circuit of V_2, giving rise to

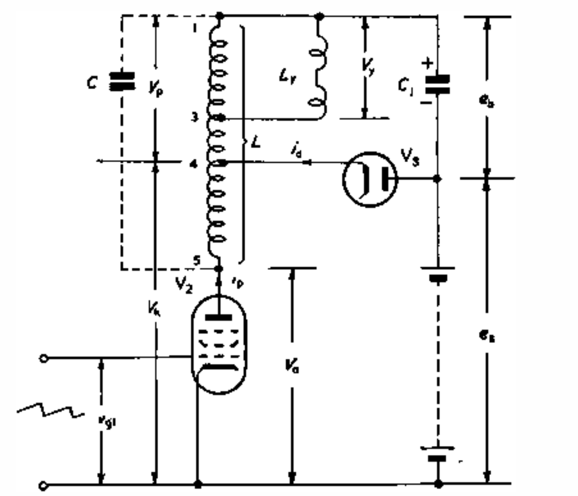


Fig. 3. Basic series efficiency-diode circuit

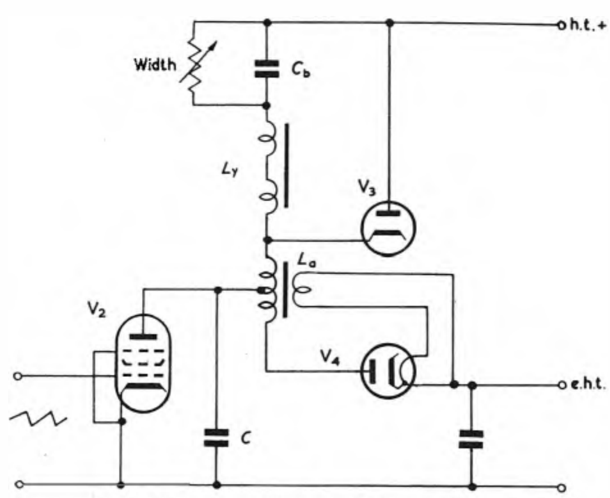


Fig. 4. Simplified direct-drive circuit

the nomenclature 'direct-drive' applied to this circuit. The chief difference between Figs. 3 and 4 lies in the inclusion of a coil L_a in the V_2 anode circuit. Its purpose is to store energy during the scanning period, which is given up to L_y during the retrace period. A complete treatment for this circuit has yet to be evolved but a useful description of its mode of operation is given by Jones and Martin[4] in which details are included of a practical design procedure. One major disadvantage of this circuit is that a high value of L_y is required for efficient operation. A value of some 34mH is quite common, and a large peak potential of 4 or 5kV is built up across the coils during the retrace period, leading to severe insulation problems.

Theory of Series Circuit

From the previous discussion of shunt and series circuits it has been noted that if zero losses are assumed then deflexion of the tube will require only 'wattless' power.

A simple equivalent circuit illustrating this has been given previously by Schade[5] (Fig. 5(a)). Here the scanning coils are shown as L,r with their associate stray capacitance as C. In order to commence the scanning cycle let switch S be closed. The source E will then cause an exponential rising current:—

$$i_1 = I\left[1 - \exp\left(\frac{-rt}{L}\right)\right] \quad \ldots\ldots\ldots\ldots \quad (3)$$

to flow through coil L, where t is the time from the scan commencement. Providing i_1 increases to only a small percentage of the maximum current $I = E/r$, then this rise will be substantially linear.

When the required value of i_1 has been reached the switch S is opened. The stored energy in the magnetic field $L.i_1^2{}_{(max)}$ Joules is then converted into potential energy causing a large potential to be developed across capacitor C. This energy, $\frac{C.E^2{}_{(max)}}{2}$ Joules in turn causes a negative-going current i_2 to flow through $L.r$. At the required

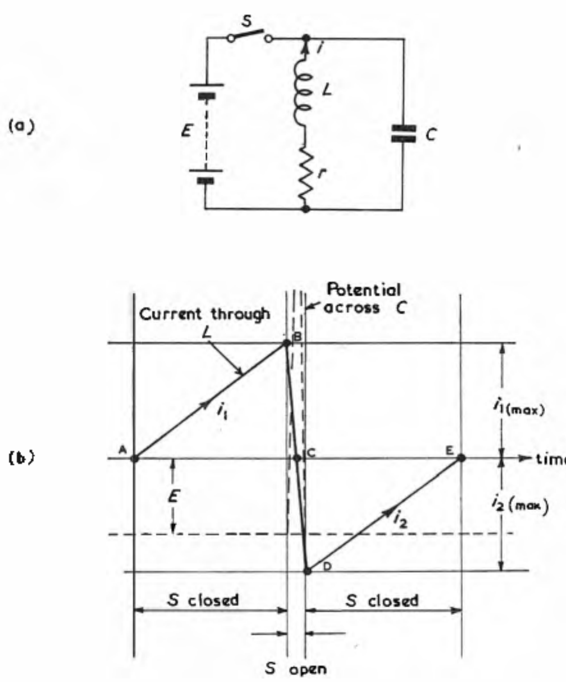


Fig. 5. Equivalent scanning circuit

instant when the potential across C is equal to the supply source E, switch S is again closed and this negative-going current i_2 decays exponentially to zero. This decay is also substantially linear and of the same slope as the rise of i_1. During the period of i_2 current decay the energy stored in the oscillatory circuit is returned to the battery, until $i_2=0$, at which time current is again drawn from the supply and the cycle recommences. A complete scanning cycle is given in Fig. 5(b) showing the variations of voltage and current during the one cycle just described.

In a practical circuit the switch of Fig. 5 is replaced by two valves, often a pentode or tetrode, and a diode. By applying a controlling potential waveform to the pentode grid the valves may be made to conduct alternately, performing the required switching action. One such circuit has already been given in Fig. 3. Consider first the pentode, V_2, to be cut-off by a large negative potential applied to its control grid, g_1. At the instant $t = 0$ let this potential $-v_{g1}$ be instantaneously reduced to zero. V_2 will conduct and its anode potential quickly reduced to the 'knee' of its i_a/v_a characteristic for the condition $v_{g1}=0$ volts. The current i_p which flows will be the sum of the magnetizing current i_m and the current through the scanning coils I_y, i.e., $i_p=i_m+I_y$. Now as the anode load is inductive a back e.m.f. will be set up across the transformer winding,

$$V = \frac{-L.d\ i_p}{dt} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

where, $L = L_y.(N_{15}/N_{13})^2$. As both V (the 'knee' potential) and L are constant then the rise of current di_p/dt will be a linear function with time.

The diode V_3 is tapped across the transformer at a point 4, such that the potential V_k is slightly in excess of the h.t. potential e_s during the conduction period for V_2, so that V_3 remains cut-off. After an interval, dependent on the controlling waveform, V_2 is again instantaneously cut-off by a large negative grid voltage $-v_{g1}$. The energy stored in the transformer magnetic field set up by the flow of current in L causes a current I to flow around the oscillatory circuit LCr, where r is the total circuit resistance. Initially this current is equal to i_p. Should V_2 remain cut-off and V_3 omitted from the circuit, then this tuned circuit would continue to oscillate and produce a train of damped oscillations at its natural resonant frequency, decaying exponentially as the circulating energy is dissipated in r.

With V_3 included, however, at the first current reversal point 4 becomes negative with respect to the h.t. potential e_s and V_3 is able to conduct. This valve has a low internal resistance and holds this point constant at about e_s volts. The energy not dissipated in the first half cycle causes current to flow through V_3, C_1 and L_{1-4} and the maximum value of this current can be shown to be—

$$i_d' = -i_p \,.\, \exp\left(\frac{-1{\cdot}65}{Q}\right) \ldots\ldots\ldots\ldots \quad (5)$$

where:—

$$Q = 1/r\ \sqrt{(L/C)}$$

During the period of V_2 conduction it was seen that V_3 was held cut-off by a cathode potential slightly in excess of its anode potential. Now that V_2 is cut-off and V_3 conducting, the operating conditions as far as the transformer and L_y are concerned, remain the same, i.e., a constant potential is maintained across transformer terminals 1-3 (L_y) and a linear falling current will flow through L_y. When this current has reduced to zero the grid potential for V_2 is again reduced to cathode potential and the cycle recommences.

The sequence of events is as given in Fig. 6 which shows the current and potential waveforms associated with the scanning circuit of Fig. 3. The complete scanning cycle may be divided into four sections, viz:—

(a) AB—second half of the scan period (V_2 conducting, V_3 cut-off)

(b) BC—first half of the retrace period (both valves cut-off, current flowing into circuit stray capacitance C)

(c) CD—second half of the retrace period (both valves cut-off, current flowing out of C)

(d) DE—first half of the scan period (V_2 cut-off, V_3 conducting).

This sequence of events may profitably be compared with that of Fig. 5.

From the foregoing it may be seen that an extremely high peak potential $V_{p(max)}$ is developed at the pentode anode during the retrace period:—

$$V_{p(max)} = L_{1\text{-}5}\ i_p/t_r \text{ volts} \quad \ldots\ldots\ldots\ldots \quad (6)$$

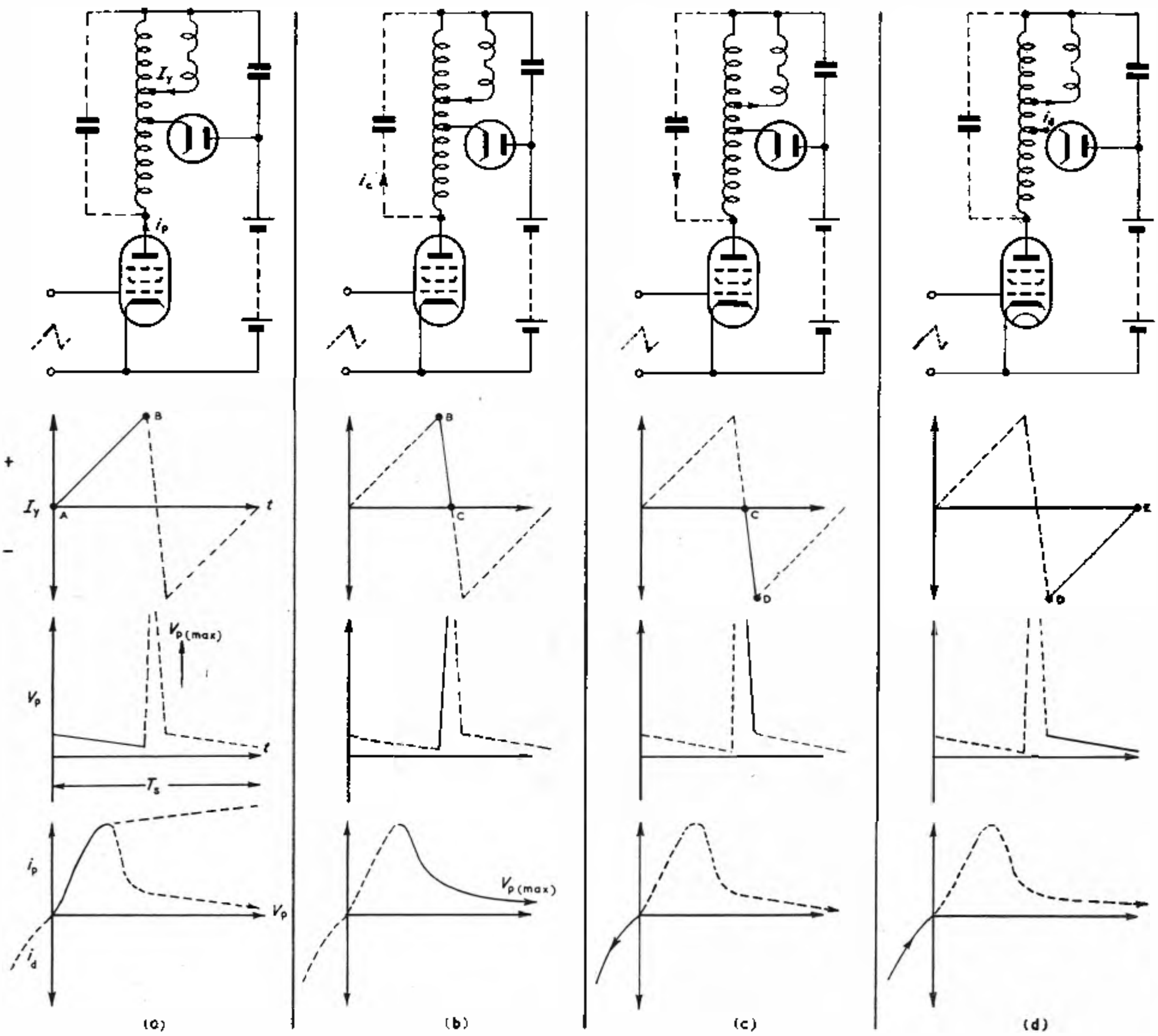


Fig. 6. Operation of efficiency-diode circuit

where:—

$L_{1\text{-}5}$ = inductance between terminals 1-5 in Henries

i_p = peak pentode anode current in Amperes

t_r = retrace period in seconds.

Substitution of practical values will show that peak potentials of between 4 and 7kV may be expected at this point. This has led to special design of line scanning valves provided with adequate anode insulation and, as will be seen later capable of supplying large peak anode currents at low anode potentials.

During the conduction period for V_3 a charging current i_d flows into the boost storage capacitor C_1. The charge acquired by C_1 may be added to the h.t. supply voltage e_s to provide a high h.t. rail supply for V_2, and may also be used for certain other sections of the receiver (e.g.: frame sawtooth oscillator). This 'boost' voltage e_b is often equal to or greater than e_s and is an important factor enabling efficient operation from a low h.t. supply to be realized.

Influence of Valve Characteristics on Performance

From the earlier simple equivalent circuit it was seen that a switch was postulated capable of operating at a high repetition rate and introducing no losses into the circuit. When this is replaced by two valves as previously described, then finite impedances are introduced which moreover are capable of varying during the scanning cycle. Consequently in order to appreciate the magnitude of the deviation from the desired linear sawtooth, it is necessary to interpretate the scanning cycle of Fig. 5(b) in terms of actual valve characteristics.

This may be done by using composite pairs of i_a/v_a characteristics such as are often used in the analysis of push-pull amplifier operation. A typical pair of curves is given in Fig. 7 for an Osram N339 pentode and U329 diode. Only the pentode characteristic for $v_{g1} = 0$ has been shown for the simple mode so far described, as the pentode is either non-conductive with a high value of negative bias applied to g_1 or this bias is removed, allowing i_a to increase along the characteristic shown.

A method of using this diagram for linearity assessment has been described[6] wherein a fictitious diode is assumed to be connected directly to the anode of V_2. This diode will have the characteristics of the diode actually used but with its scales of voltage and current adjusted by the transformer ratios N_{15}/N_{14} and N_{14}/N_{15} respectively, as shown by the dotted curve of Fig. 7. It will be observed that a region of severe non-linearity occurs at about zero transformer current (centre of picture) due to dissimilar slopes of valve characteristics OA and OB. This may be corrected by biasing the diode so that it will not conduct until the relative anode potential reaches a potential greater than that of point O, i.e., curve OA is moved to the right of the origin. An examination of the circuit of Fig. 3 will show that this can be obtained by adjustment of the diode tapping ratio N_{15}/N_{14}.

For no change of potential across L_y during the scan period the diode cathode must assume the boosted potential e_b during the diode conduction period, i.e., $V_{14}=e_b$. It may also be noted that the potential developed across the complete winding N_{15} is the difference between e_s+e_b and the minimum potential developed across V_2 (i.e.: the cathode potential of the fictitious diode) so that:—

$$V_{15}/V_{14} = N_{15}/N_{14} = \frac{e_b+e_s-v_a}{e_b} \qquad (7)$$

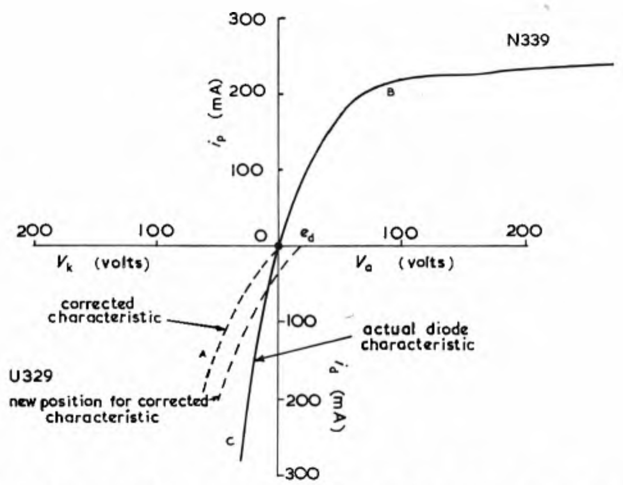


Fig. 7. Derivation of scanning circuit performance from valve characteristics

Thus for constant h.t. and boost potentials the commencing point for diode conduction may be altered by adjustment of turns ratio N_{15}/N_{14}.

Modes of Operation

Many other modes are possible other than the simplest one outlined above, and are dependent on the grid controlling waveform for V_2, position of diode tap, and value of load impedance. For convenience they may be classified in groups according to the pentode controlling grid waveform.

The simplest is that of a rectangular driving potential waveform where the grid potential is made to alternate instantaneously between two levels, that of zero voltage and a large negative potential sufficient to render the valve non-conducting.

Mode (A) in Fig. 8 has already been discussed. Other modes are dependent on the alternative position for the diode tap and two of these are also shown in the diagram. Curve (B) has a smaller diode ratio N_{15}/N_{14} and provides a larger total current swing. The diode now conducts over a longer fraction of the scan period; mean pentode current is increased and the efficiency is somewhat reduced. On further decreasing ratio N_{15}/N_{14} a position is reached, shown as (C), where severe cramping is experienced at the right-hand side of the picture. In a practical case a ratio of between 1·5 and 1·8 is chosen to obtain a compromise between the conflicting factors of efficiency, scan and linearity required, and maximum valve currents permitted.

Turning now to a sawtooth controlling waveform. With this applied to the grid via a *CR* network, grid rectification takes place causing the coupling capacitor to acquire a negative charge approximately equal to the peak positive excursion of the sawtooth waveform. In operation, therefore, the grid potential is varied linearly from about, or slightly above, zero potential to a negative potential dependent on the amplitude of the controlling waveform, thus fixing the period of pentode conduction.

As the change of grid potential is now no longer instantaneous, a family of i_a/v_a curves will have to be considered

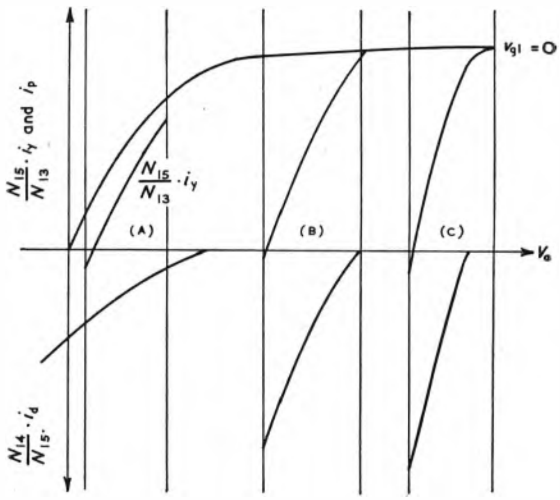


Fig. 8. Modes of operation—square-wave input

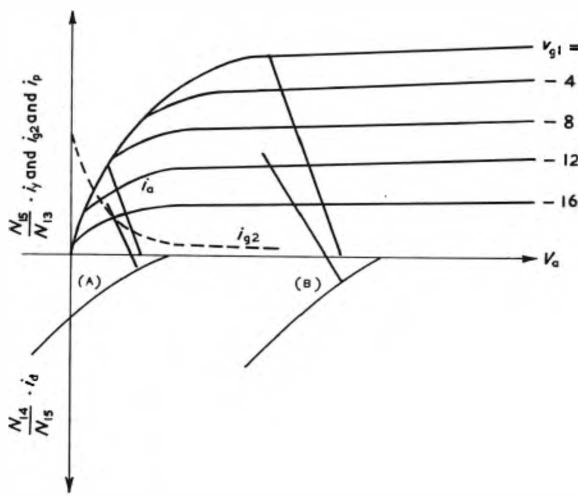


Fig. 9. Modes of operation—sawtooth waveform input

as shown in Fig. 9, where the anode current locus will vary along a dynamic load line drawn across these static characteristics. Operation with a high transformer ratio is shown as (A), and will cause V_2 to work over the lower portion of its characteristic. This will have the undesired effect of high g_2 current, and therefore excessive screen-grid dissipation, and as the g_2 current contributes nothing to the output it is usual to work V_2 over as high a portion of its 'knee' characteristic as possible consistent with allowable maximum anode dissipation. Such a position is shown as (B) in Fig. 9, where a lower N_{15}/N_{14} ratio is assumed.

In many practical designs the drive waveform often takes the form of Fig. 10. The flattening of the positive excursion of the sawtooth potential in (a) is due to the flow of grid current in V_2, while the negative pulse added to the waveform in (b) is often included to ensure that V_2 remains completely non-conductive when its anode reaches a high peak potential during the retrace period.

The inclusion of a modulating waveform in series with the diode for linearity correction, as described in the next section can also considerably modify the mode of operation for the scanning circuit. In particular pentode and diode current waveforms are affected and will depart from those given above.

Controlling Scan Linearity

From the preceding discussion it became evident that perfect linearity can only be obtained from the resonant-return circuit at the expense of either scan amplitude or

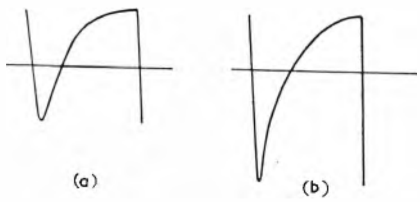


Fig. 10. Drive waveforms

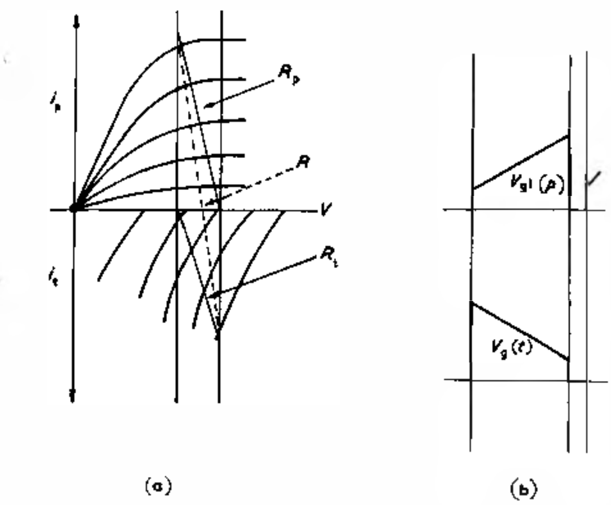


Fig. 11. Use of a controlled triode for V_s

excessive valve anode currents. Various methods have therefore been evolved to correct the linearity of the generated waveform for other, more convenient, modes. A simple method has been suggested whereby the scanning coils disposed on either side of the tube neck are rendered assymetric by their dissimilar shape and inductance value. The non-linear flux distribution across the tube neck can then be arranged to oppose an equal non-linearity in the scanning coil current. It will be obvious that this is far from being a flexible arrangement, as one requirement of a linearity-correction system is that it must be capable of simple adjustment to cater for variations met with in production and service. This requirement can be met in a number of ways, several of which are described in the following sections.

Use of a Controlled Triode

Perhaps the most flexible of these is the use of a low-impedance triode to replace the boosting diode. This system has been fully described elsewhere[5] wherein it is shown that by applying suitable controlling waveforms to both pentode and triode grids, then adequate contrbl of the rate of change for I_y is obtained over the entire scanning period.

It was seen earlier that by adjustment of the diode conduction period, the scan linearity may be altered. In Fig. 11(a) is shown the composite characteristics for the two valves, the single diode curve being replaced by a family of triode characteristics. By varying the triode grid potential during the scan period, the same effect may be achieved as if the diode tap N_{15}/N_{14} were to be altered during this period. By suitable choice of controlling waveforms, a combined load line R can be obtained from the individual load lines R_p and R_t to give a substantially linear rise of current. The shape of the grid waveforms is shown in Fig. 11(b), the triode input being derived from the pulse potentials developed across the transformer windings. This has been obtained by means of the network C_1, R_2, C_2, R_3 in the practical arrangement shown in Fig. 12.

Low impedance triodes suitable for this application are

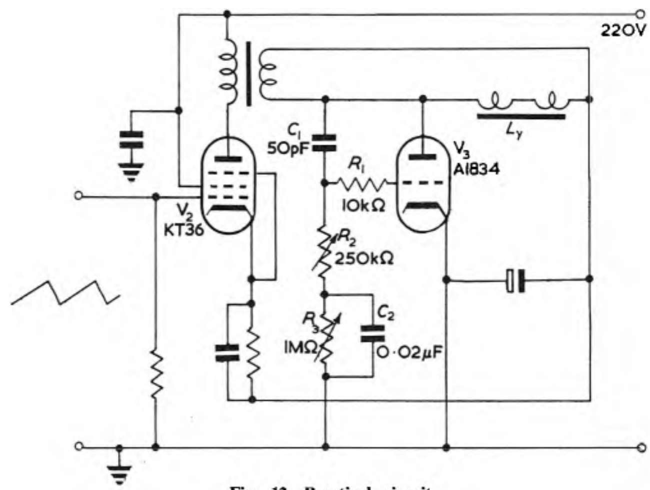


Fig. 12. Practical circuit

available (e.g. Osram A1834 or American 6AS7G) and find a use in monitor and television camera equipment where the additional cost involved is no deterent.

Use of a Controlled Diode

It is possible to simulate triode action by including in series with the diode circuit a low impedance generator supplying a correction waveform. To achieve this a damped tuned circuit, resonant at line frequency, is included in series with the diode circuit as shown in Fig. 13. This circuit is excited into resonance by the current transformed to it from the coupling winding 2-3. Adjustment of L and C values together with the degree of coupling M will alter the amplitude, phase, and frequency of the potential waveform developed across the tuned secondary 1-3. This adjustment is fairly critical and it is convenient to provide an adjustable ferrite or dust-iron core for the linearity transformer L to enable a variation in inductance to be obtained.

The reactive voltage developed across 2-3 will have to be considered when selecting transformer ratios and will lead to a somewhat lower ratio for N_{15}/N_{14} than is determined by calculation. This adjustment is best obtained empirically as the actual voltage is dependent on several inter-related quantities. A check for this may be made by adjusting the inductance value for L while at the same time monitoring the e.h.t. produced (an approximate criterion of circuit efficiency). Optimum linearity should then coincide with maximum e.h.t. obtained.

Saturated Reactor Circuit

A further method is dependent on the properties of a choke under partial core saturation conditions. The method has been used earlier by several workers mainly for the linearization of relaxation oscillators[7] and now finds wide use as a simple and fairly flexible method of scan-correction in efficiency-diode circuits.

The principle of operation is quite simple; a cored-choke *Ch* is included in series with the scanning coils as shown in Fig. 14. This core is subject to a constant applied flux so that the sawtooth current flowing through the coil can cause the core to become saturated when the value of this current exceeds a certain value. Under these conditions

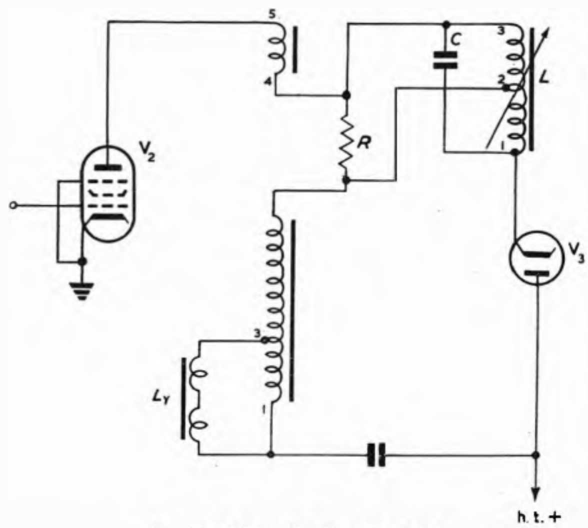


Fig. 13. Controlled linearity circuit

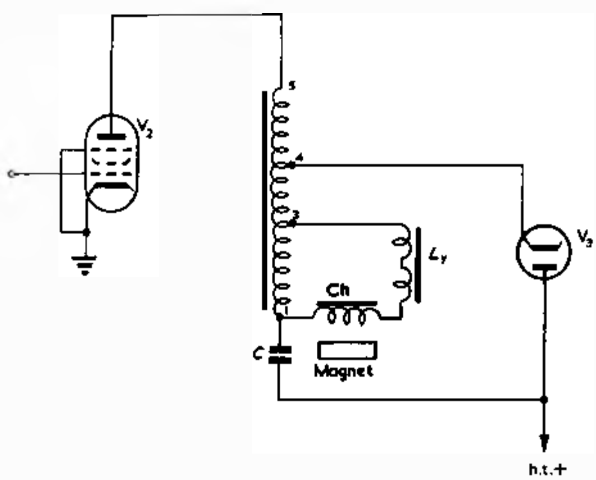


Fig. 14. Saturated reactor linearity circuit

the choke inductance falls, allowing an increase in scanning current to take place. The choke may be considered as having a negative voltage/current characteristic and be equivalent to inserting a negative resistance in series with the scanning circuit, whose value is controlled by the amplitude of sawtooth current flowing through it.

Ferrite material can be used for the choke core and during the scan period the material will be worked over a portion of its μ-H curve, the inductance of the choke being correspondingly varied. The extent of this inductance change will depend on the number of turns for the coil (governing section of curve used), fixed magnetic bias, and volume of core.

Two practical arrangements are possible, a closed magnetic circuit in which the fixed bias is obtained by a polarizing winding on one of the core limbs, and an open core across which a permanent magnet may be shunted.

Closed Magnetic Circuit

If it is proposed to include the linearity corrector in a completed design it is first necessary to obtain a curve of the uncorrected scanning current showing the departure from mean scan velocity at the screen centre. Such a curve is shown in Fig. 15(a) from which the actual and desired current waveforms can be derived (Fig. 15(b)). This gives the percentage of scan width which must be lost in order to correct the non-linearity of the scanning waveform. For the case shown this is quite large, about 21 per cent.

The value of I_Y is known so that the voltage drop across

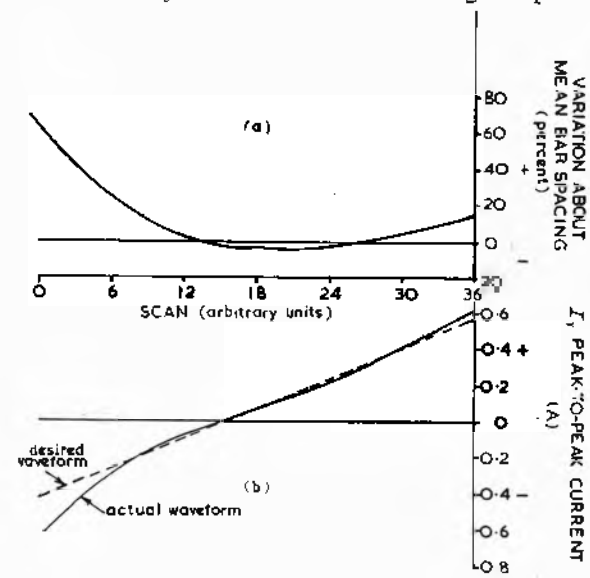


Fig. 15. Derivation of scanning linearity

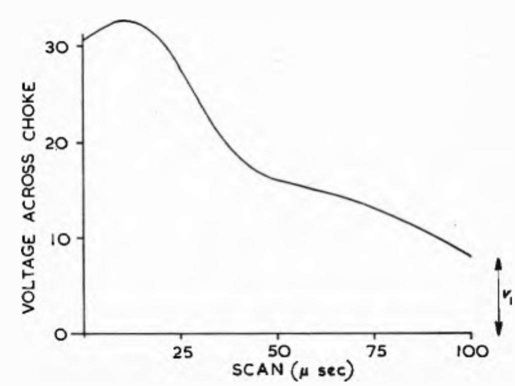


Fig. 16. Waveform across linearity choke

the scanning coils V_Y may be calculated. At the commencement of the scan period it is desired that the added linearity inductance L_1 will drop 21 per cent of this voltage in order to correct for severe acceleration distortion. Fig. 16 illustrates the case of a 90° scanning-angle circuit and with the tube operating at an applied e.h.t. of 16kV, the value of I_Y, given 10mH coils, will be about 1·22A. Therefore:—

$$V_Y = L_Y \; I_Y / T_s = 120\text{V}$$

and the correction voltage:—

$$V_c = \frac{0{\cdot}21 \; V_Y}{1 - 0{\cdot}21} = 32\text{V}$$

corresponding to an inductance figure for L_1 of 2·7mH.

Jones[6] gives a method of calculating core volume required based on the non-linearity derived from valve characteristics. The volume is however by no means critical and a trial with several core shapes at the calculated inductance figure will quickly determine the size required.

The polarizing d.c. supply needs to be made variable to provide a control over linearity, and an approximation for the ampere-turns of this winding may be derived from the linearity choke assuming a magnetic bias of one half the maximum change of flux obtained during the scan period.

i.e.: $-N.I_{d.c.} = \frac{N_{ch}.I_y}{2}$

The d.c. source must not be of too low an impedance in order to avoid undue damping of the linearity choke. Also due to the induced pulse potentials at line frequency it is desirable to derive this current from the line scanning circuit itself. The adjustment of the circuit is best carried out while observing the waveform developed across the choke. The waveform obtained at the correct amplitude can be compared with the curve actually required Fig. 15(a).

Open Magnetic Circuit

Instead of using a closed circuit and electromagnetic polarizing, a cored solenoid may be used and magnetic bias provided by a permanent magnet. The design procedure can be similar to that given above, the strength of the required polarizing field being determined by the known ampere-turns in the choke. Unless the core is well saturated however considerable scanning power may be dissipated within it. This can be assessed by observation of the choke waveform where the scan lost in this way can be estimated from the minimum potential level v_1 (Fig. 16).

Several mechanical methods of construction are possible—either the choke inductance can be adjusted by means of a sliding core within the solenoid, or alternatively the intensity and direction of the magnetic field can be varied. This latter method will give a closer coupling between the core and magnet enabling a smaller magnet to be used. Two suggested designs are shown in Fig. 17. In the first a small magnet is made capable of rotation between two soft-iron pole pieces affixed to the ferrite core upon which the choke is wound. The leakage magnetic field of a permanent magnet focusing device is utilized in the second design. This leakage field here can be quite large so that the variable inductance method is permissible without too much loss of scan width.

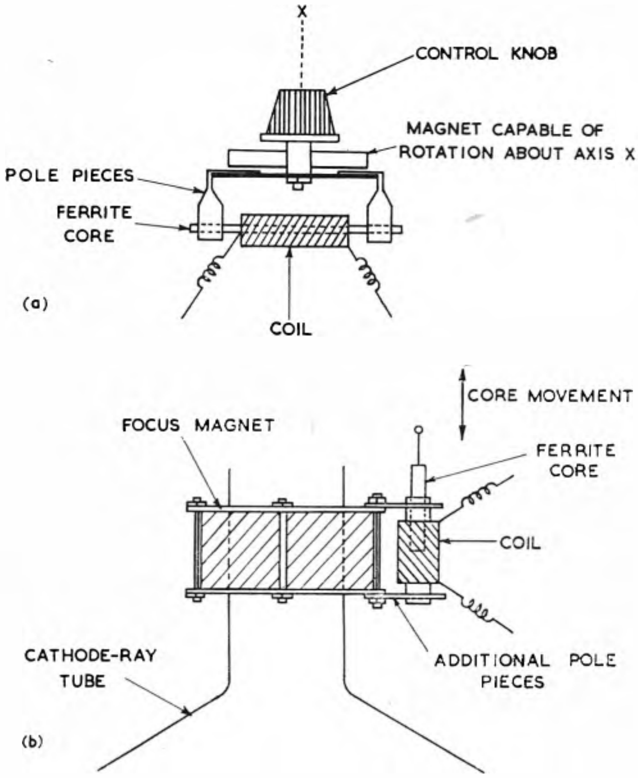


Fig. 17. Two methods of saturated-reactor assembly

Controlling Scan Width

As means will be required in any practical scanning circuit for adjusting the amplitude of sawtooth current generated, then a width control must be included in the completed design. A full analysis of this problem has been given previously[8] wherein it was shown that to avoid undue variations in e.h.t. generated by the scanning circuit then only one form of width control appears permissible. This has been termed the 'differential method' and consists of two coils whose inductances are simultaneously varied by a common adjustable core (Fig. 18), so that as the inductance of one coil is increased, the inductance of the other coil decreases. By connecting the two coils so that one is in series with scanning coils, while the other is in shunt with part of the transformer winding then the total inductance, measured across the scanning coil terminals, remains constant as the division of current between shunt and series paths is varied. To avoid resonances of the series coil with its associated self and stray capacitances it is necessary to shunt this coil with a series combination of resistance and capacitance. It may be shown that if the relationship $L_s = C.R^2$ is maintained then the circuit will be rendered non-resonant and modulation of the scanning waveform avoided.

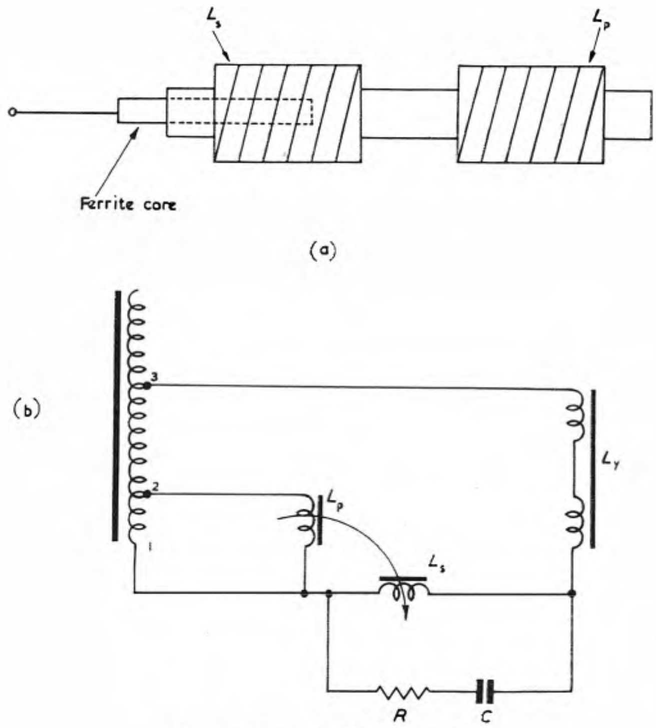


Fig. 18. Differential width coil system

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(To be continued)

Use of Scale Model Techniques in the Design of V.H.F. and U.H.F. Aerials

By F. J. H. Charman*, B.E.M., J. Thraves*, M.Sc., and E. F. Walker*, B.Sc.

In many cases due to the size or the nature of the aerial, it is not practicable to carry out measurements on a full-sized aerial and the use of a scale model is of considerable advantage both for design and assessment and for demonstration purposes. In this article the theory and practical methods of measuring radiation patterns of complex aerial systems by scale modelling are shown, and several examples illustrated.

IN some classes of work, for various reasons, it is not practicable to carry out measurements on a full sized aerial. For example the physical size of the aerial may prohibit the application of conventional methods of radiation pattern measurement. This is particularly so with certain aerials on aircraft, ships, missiles etc. in which currents induced by the aerial on the vehicle itself form the true source of radiation. Also when large aerials are being considered it may well be a very expensive process to build and test several full sized experimental models before the final dimensions of the aerial can be determined. Obviously both these tasks can be carried out much more quickly, cheaply and efficiently on a scale model of the aerial. Furthermore, to measure accurately the radiation pattern of, for example, an aircraft aerial, it is necessary to support and orientate the aircraft away from the ground so as to approximate free flight conditions. Obviously, the procedure is much simpler if a scale model aircraft and aerial can be used.

Scale modelling may also be used to advantage in determining the echoing areas of radar targets, since these are often large and cumbersome. Scale models also form a convenient aid to a lecturer. With a very simple set-up on the demonstration bench, important features concerning the radiating properties of aerials can readily be demonstrated. This is particularly so with "wire aerials" which are more often used at h.f. and v.h.f.

Scale modelling is also a very useful method in the laboratory as an aid to hastening the assessment of a new type of aerial. A simple model of the proposed design can be constructed, and its efficiency can be determined. Confidence can thus be found to proceed further with the design and construction of the actual aerial under consideration.

Theoretical Justification

That the experimental modelling of electromagnetic systems is justified can be seen from the principle of electrodynamic similitude[1,2,3]. The validity of the method arises out of the inherent linearity of the electromagnetic field equations. Consequently, all non-linear materials (e.g. certain ferromagnetic substances) must be excluded from considerations. Alternatively, the scaling conditions may be obtained by starting from the equations of wave continuity[4]. These are:

$$L^2f^2 = A/\mu\epsilon \quad \ldots\ldots\ldots\ldots \quad (1)$$

$$L^2f = B/\mu\sigma \quad \ldots\ldots\ldots\ldots \quad (2)$$

* E.M.I. Electronics Ltd.

Where L is a length or scale factor.

f is the frequency,

μ is the permeability,

ϵ is the permittivity,

σ is the conductivity,

A, B are constants.

Now $$1/\mu\epsilon = (\text{velocity})^2 \quad \ldots\ldots\ldots\ldots \quad (3)$$

and is independent of frequency. Thus for the scaling of a model to satisfy equation (1), the condition is

$$Lf = \text{constant} \quad \ldots\ldots\ldots\ldots \quad (4)$$

In equation (2), the conductivity is involved. This affects the skin depth of the currents and hence the effective size of current surfaces. However, since the skin depth for good conductors at v.h.f. and u.h.f. is not greater than 0·001in, the effect is negligible. Equation (2) may be written:

$$L^2f^2 = B/\mu\epsilon \cdot f\epsilon/\sigma$$

Remembering equations (3) and (4), this reduces to the requirement

$$f\epsilon/\sigma = \text{constant}$$

This shows that in lossy circuits, the conductivity must be scaled with the frequency; it is equivalent to varying R in $R\omega C = 1$, the equation for constant phase angle with frequency.

Thus to maintain similarity when the physical dimensions of the system are scaled down by N:1, both the frequency and the conductivity must be increased N times. In practice many aerial systems contain materials whose conductivity is either zero (lossless dielectrics) or very large (metals). Consequently, provided the model be made of high conductivity metal then the only parameter that need be accurately scaled is the frequency. This applies particularly to aerials mounted on present day aircraft since these are usually of an all metal construction. However, should the aerial system contain a lossy dielectric for instance, then the dielectric loss must be scaled as well as the frequency.

The conductor surface resistance of an aerial might have some effect on the radiating properties of the aerial since it could effect the distribution of current in the aerial. On the other hand, in all aerials which are at least half a wavelength in dimension, the loss due to conductor resistance, assuming good metallic conductors, is very small compared with the loss due to radiation. This latter, of course, determines the current distribution, but that is a function of the shape of the aerial, and is a property which scales with the aerial. Thus, one can say that for all efficient aerials, the radiating properties can be scaled.

Modelling Techniques

Before proceeding to describe in more detail several examples of the use of this, some mention must be made of the practical methods used in setting up model aerials.

The most important choice of all is the scale factor to be used. Primarily, this is such as to reduce the model aerial system to one of a convenient size for handling in a laboratory, and with which to take measurements. The availability of measuring equipment at the new frequency has also to be considered.

The reciprocity principle allows the choice between transmitting or receiving with the model, whatever the ultimate full-scale application, and the choice can be made in terms of practical considerations. In some circumstances, where the model size chosen is sufficiently large, a battery driven oscillator can be built into the model and its aerial system made to act as a transmitting aerial. This state of affairs is preferable since it means there are no leads running to the model which can interfere with its radiation pattern. However, in some cases the model is so small that it must be used as a receiving aerial; this has been the case in most of the work carried out by the authors, and will be assumed in the following illustrations.

Since scale models are primarily used in radiation pattern investigations, and the patterns are unaffected by the impedance matching, the model aerial system need not be a detailed copy of the proposed full size aerial. All that is necessary is to ensure that the model aerial carries substantially the same current distribution as the actual aerial, since this is the factor which controls the radiation pattern. However, to make best use of the available signal, the aerial is matched to the detector by means of a variable transformer or 'tuner' placed as close as possible to the aerial. It has been found very important to make the best use of the available signal because the presence of relatively large r.f. power with little output from the aerial exaggerates the effects of r.f. 'leakage' between transmitter and receiver. It must be remembered that r.f. cables, or joints, are often quite 'leaky,' and could, for example, allow unwanted r.f. signals to reach the detector. In such cases, the pattern obtained presents a large number of spurious lobes, far more than can be produced by the model with its limited size in wavelengths.

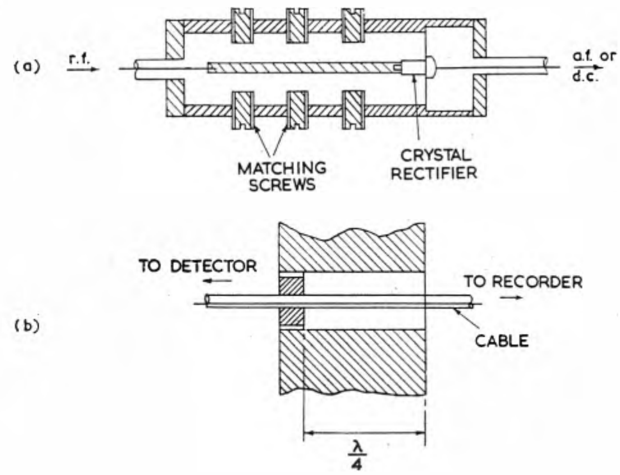

Fig. 1. Quarter-wave choke and tuner
(a) Tuner and detector
(b) Cable exit from model

The model can in general only throw up a number of undulations in its pattern equal to its dimensions in half-wavelength units (e.g. a long wave aerial). If there are site reflections the number of lobes or undulations appearing will be much greater—characteristic of the site rather than the model.

The type of tuner used varies with the frequency, but the one described below can be made to operate satisfactorily up to frequencies of 5 000Mc/s. This is convenient since it covers a large number of practical cases as will become apparent in later sections of this article. The tuner, shown diagrammatically in Fig. 1(a), consists of a length of coaxial line, with capacitance screws spaced $\frac{1}{8}$ of a wavelength apart, terminated in a crystal diode. The screws are adjusted, through holes in the model, for maximum output. The resulting d.c. signal (together with a.f. if a modulated system is used) is brought out of the model using a fine coaxial cable, such as Pyrotenax (UR137), which has an outside diameter of 0·125in. This cable is constructed as a drawn copper tube, and is quite rigid and conveniently small.

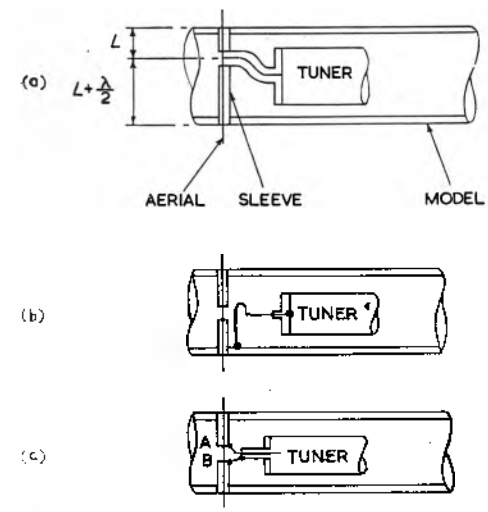

Fig. 2. Feeding arrangement for two unipoles

The presence of this cable, carrying the detected signal, introduces some difficulties. First, it must not carry away any r.f. current, and thus become part of the radiating system under examination. To avoid this, the cable leaves the model in a quarter wavelength choke as shown in Fig.1(b); it is thus isolated electrically from the model. Secondly, the lead must not disturb the field in the vicinity of the aerial. It has been found that if the vertical part of the lead is at least six wavelengths from the aerial, even if it is parallel to the electric field, there is very little distortion of the pattern.

The modelling of aerials and the feeding arrangements often requires ingenuity. In the case of a unipole, the inner conductor of the feeding cable may be used as the radiator, while the outer conductor could be terminated at the surface of the body. Methods of feeding two unipoles on a cylinder, diametrically opposite, are shown in Fig. 2. In general the aerials will be a balanced pair, although types (a) and (c) can be adapted for any phase difference.

In (a), the difference in length of the feeding cables gives the required phase difference. In (b), the coupling is electromagnetic. For greatest efficiency, the loop should be as close to the bared section as possible, this section being about ¼in long at a frequency of about 5 000Mc/s. A more complicated method is shown in (c). This arrangement depends on the fact that the 'short-circuit' distance between A and B via the circumference of the cylinder can be made an odd multiple of a half wavelength. ¼in Pyrotenax is again a suitable cable to use; it has a velocity factor of 0·525 and a characteristic impedance of 42Ω.

The above discussion has been limited to some methods which have been found useful in producing models of v.h.f. aerials and their feeding arrangements when placed on various structures such as cylinders and flat plates. Similar problems often rise with microwave aerials and it becomes necessary to determine their radiating properties when mounted on complicated structures. Modelling techniques are more difficult to apply in this case if the original microwave aerial is of a complicated shape such as a waveguide-fed slot array. This follows since the model aerial will need to be scaled with the same complexity. However, with more simple types of microwave aerials such as horns, etc., the scale models are also simple to make. The waveguide aerials may be scaled directly and the detector can be a conventional type of microwave crystal holder and associated tuner. Successful scalings of S band to Q band aerials have been carried out.

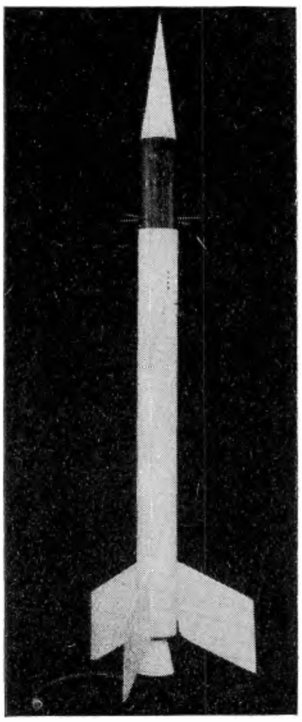

Fig. 3. Model of 'Skylark'

Mounting of Models

The models are usually rotated when taking radiation patterns, and it is most convenient to rotate them about a vertical axis. On the other hand, they must be in 'free space' as far as possible, and therefore well clear of the rotating gear. They can be mounted on the support in various attitudes as required for different sections of radiation pattern, the illuminating aerial being set up with vertical or horizontal polarization, as required.

Two types of support have been used: poles of s.r.b.f., or towers of foam dielectric. The poles should, of course, be as light as possible, and waxed thread stays may be necessary with large models. Foam dielectrics with a high degree of expansion, and dielectric constants as low as 1·05 and very low loss angles can be obtained; these are substantially transparent to r.f. waves. Towers several feet high can be constructed from slabs of such material, and are suitable for small models. It is difficult to detect any distortion of the pattern due to the small amount of material below the models.

The Illuminating Aerial

The requirements placed upon the illuminating aerial are common to those placed upon the transmitting aerial when taking conventional radiation pattern measurements.

Firstly, the illuminating aerial must be situated far enough away from the model system to ensure that the incident wave has essentially a plane wave front. This occurs when the separation R between the two aerials is such that:—

$$R_m \geqslant 2D_m^2/\lambda_m$$

and

$$R_f \geqslant 2D_f^2/\lambda_f$$

where the suffices m and f refer to model conditions and full sized conditions respectively.

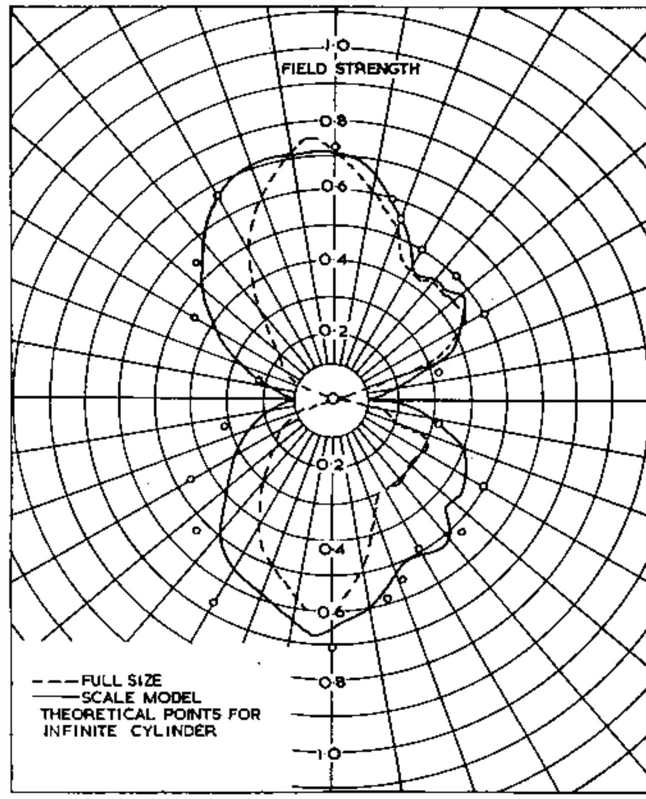

Fig. 4. Radiation pattern of telemetry spike on 'Skylark'

Since the relationship between D_m, D_f, λ_m and λ_f are already known from a previous part of this article it can be seen that

$$R_m = 1/N\ R_f$$

where N = the scale factor.

This brings out again a further advantage of scale modelling, i.e., that a smaller space is needed for radiation pattern measurements.

In these expressions, λ is the wavelength and D refers to appropriate dimensions related to the aerial. If the aerial is well isolated from the surface or structure on which it is mounted, D refers to the maximum aperture of the aerial. However, if the aerial is not isolated from its surroundings, D refers to the maximum effective dimension of the surroundings. For instance, with an aerial mounted on an aircraft the radiation pattern may well be influenced by reflection of energy from the aircraft wings and fuselage. Consequently D must refer to the aircraft dimensions.

When precise measurements are not needed, the range may be more than halved without grossly distorting the pattern. This is advantageous when site conditions and r.f. power are restricted.

Secondly, the wave incident upon the model must be reasonably uniform in amplitude over the dimension D. This means, in effect, that the beamwidth of radiation coming from the illuminating aerial must be greater than a certain minimum size depending on the values of D and R used.

Thirdly, the model aerial system should not receive energy from the illuminating aerial via reflections from surrounding objects, in particular from the ground. This controls the beamwidth in the vertical plane since too wide a beamwidth will illuminate the ground and cause unwanted ground reflections. The real solution to this problem is to work at a sufficient height above the ground.

These requirements lead to a kind of Thévenin's theorem, for scale model work: the illuminating aerial should be 'matched' to the model. It will be appreciated that the range formula above is reciprocal, and that since it is derived in terms of phase error across the aperture, it also determines the range in terms of the aperture of the illuminating aerial. Since it is desired to use the narrowest possible beam, in order to avoid site errors, the illuminating aerial should be of the largest possible aperture but on the other hand it need not be greater than the model, for this would only increase the range and there would be no further advantage.

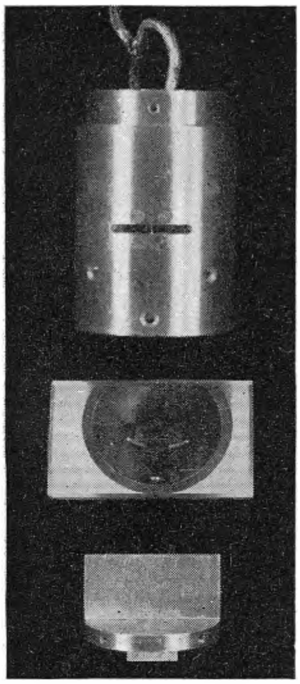

Fig. 5. Photograph of slot aerials

Practical Examples

This section will give some specific examples of scale model aerials dealt with by the authors, and possible solutions to other problems.

The aerial systems on the 'Skylark' upper atmosphere research rocket provide good subjects for scale modelling. This rocket is used as part of the British contribution to the scientific work associated with the I.G.Y., and contains several complete radio systems. One of these operates at the upper end of S-band and consists of two open-ended waveguides, diametrically opposed, though these are actually closed for aerodynamic and matching reasons by p.t.f.e. plugs. The photograph, Fig. 3, shows the model made to investigate its radiation pattern.

The scale factor chosen brought the frequency of operation to Q-band. The aerials themselves consisted of lengths of waveguide (size wg 22), closed by the plugs. A conventional Q-band crystal detector was placed in the centre, and the rectified signal led away by a cable through the rear of the model, as shown. Later, the polarization of the aerial was changed from linear to circular, by using circular waveguide and quarter wave plates. To find the new radiation patterns, these aerials were scaled to 0·25in diameter tubing. Two sets of radiation patterns were then found, by using a linearly polarized source and rotating the waveguide aerial and crystal detector, to find respectively the components parallel and perpendicular to the rocket axis.

Another system, the aerial of which consisted of a single, radial unipole, operated at about 500Mc/s. The aerial was constructed in the manner described above. The radiation patterns obtained could be compared with the results taken on a full size rocket. The patterns taken by revolving the rocket and the scale model about their longitudinal axes, together with theoretical points calculated for an aerial on an infinite cylinder[3], are shown in Fig. 4. It was not possible to find the radiation patterns of the full size rocket under ideal conditions; it is 25ft long and therefore it was difficult to fulfil all the conditions necessary to make accurate measurements. The figure shows the result, proving that a more realistic pattern is obtained by using a scale model.

The simulation of suppressed aerials presents fresh problems. This is also an example of finding whether a particular aerial system will satisfy certain requirements. A suppressed aerial on a missile might consist of a slot with an associated cavity behind it. Usually the space avail-

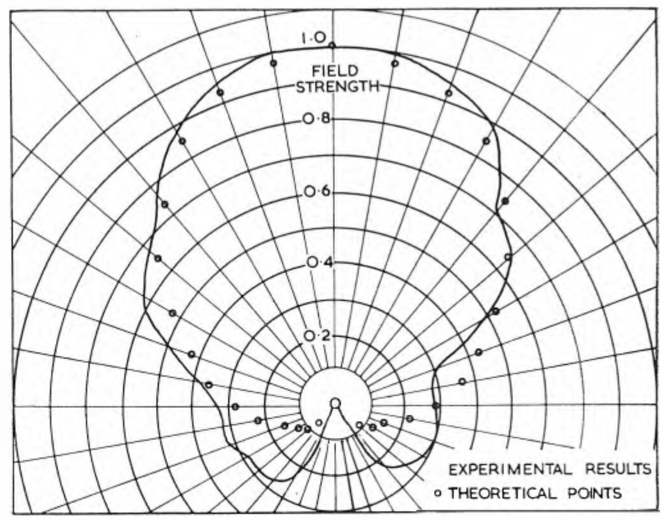

Fig. 6. Radiation pattern of slot aerials

able for such a construction is severely limited and the design is consequently complex. However, in finding its radiation pattern using a scale model, the design can be simplified, as the whole space inside the model is available. Fig. 5 illustrates a scale model of two circumferential slots, each backed by a semi-circular cavity. Each slot is fed at its centre by coaxial line, and in the case illustrated the two diametrically opposite slots are fed out of phase. Fig. 6 shows the radiation pattern of one circumferential slot, together with a calculated pattern[6] for the same aerial system on a infinite cylinder of the same diameter (2·1 wavelengths).

Acknowledgment

The authors would like to thank Miss J. M. Davies for the theoretical calculations and Mr. D. Smith for the experimental results. Acknowledgments are due to EMI Electronics Ltd. and the Ministry of Supply for permission to publish this article.

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The Accurate Measurement of a Time Interval

By A. E. Cawkell* and H. Ristlaid

A method of accurately calibrating the time interval between a triggering pulse and a delayed output pulse is described. A gated time marker generator is used, and the trigger and delayed pulses are displayed on an oscilloscope having a triggered time-base, mixed with time markers of crystal accuracy, the number of intervening markers being counted. The variable delay equipment used is capable of incremental adjustment of less than 10mμsec in a total delay of 10μsec. A technique is also described for the measurement to an accuracy of about ± 0·02 per cent of a long time interval delay, or long pulse in the millisecond range.

THE improved performance of a recently developed variable delay unit made it necessary to calibrate the time interval between a triggering pulse, and a delayed output pulse, more accurately than had hitherto been necessary.

A block diagram of the apparatus is shown in Fig. 1, and the relevant waveforms are shown in Fig. 2.

The time marker generator provides trains of time markers of crystal accuracy. In this application it is being gated by an internal gate generator of variable length and p.r.f.

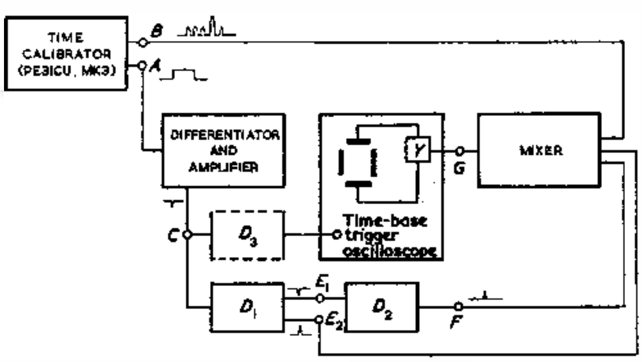

Fig. 1. The time measuring apparatus

Typically the display takes the form shown in Fig. 2 (*B*), in which the output from the 'combined' socket on the calibrator is shown, with 0·5, 1, 5 and 10μsec markers switched on, and appearing at different amplitudes for ease of identity. The markers last for the duration of the gate (*A*).

The gate output from the calibrator (also externally available), is differentiated (*C*), and is used to trigger the oscilloscope time-base, the sweep length being adjusted (*D*) to slightly exceed the gate length.

The variable delays D_1, D_2, are type PE31VAD (Mk 2), but could be of any similar type. These particular units produce delayed pulses in the range 1μsec to 1msec from an initiating trigger pulse, and consist of a 'Sanatron' circuit, pick-off comparator controlled by a helical potentiometer, and output pulses generated by blocking oscillators. The delayed output pulse has a rise-time of 0·15μsec, and a base length of 0·4μsec regardless of the delay setting. The time delay is smoothly adjustable by means of the helical potentiometer.

A delayed output pulse from D_1 (E_2), a further delayed pulse from D_2 (*F*) the unit to be calibrated, and the markers (*B*), are mixed, and fed to the Y amplifier or plates of the oscilloscope.

D_3 (shown dotted) is not in circuit.

* *Cawkell Research and Electronics Ltd.*

The resulting display takes the form shown at *G*.

D_1 is adjusted so that its delayed output pulse (E_2), is aligned with a convenient marker—in fact with a 10μsec marker in the diagram. This forms 'zero' time for the unit D_2 to be calibrated, as the latter is triggered from E_1, a pulse occurring at the same time instant.

The delayed pulse from D_2 appears further along the trace, and has been adjusted by D_2's delay control, to alignment with a 0·5μsec marker.

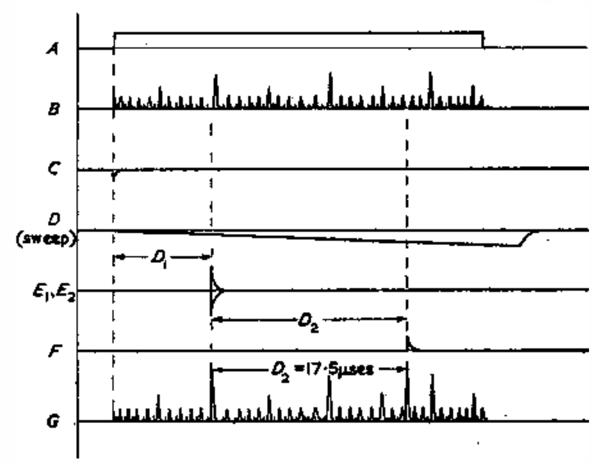

Fig. 2. The waveforms of Fig. 1

It will be observed that the delay due to D_2 may be easily counted as 17·5μsec.

One feature of the arrangement is that it has a nearly constant accuracy of perhaps ±0·2μsec, although if the sweep speed is such that the marker alignment procedure is easily resolved, it may be considerably better than this. On much longer delays ±0·2μsec would be a very small percentage of the total delay.

Using the above method it would however be difficult to adjust D_2 accurately to a very long delay—say, 995·5μsec as in this case the sweep might be set at 1msec, and it would be difficult to resolve the markers for the alignment procedure—an accuracy of about 1 to 2 per cent would be possible.

An accuracy of about ±0·03 per cent can however be obtained by the following method.

An additional variable delay, D_3, is introduced in the position shown. With D_3 set at zero delay, pulse E_2 is aligned with a suitable marker as before to establish 'zero' time. D_2 is set approximately to the desired delay—say 995·5μsec. The delay D_3 is now increased, and the trace will appear to move from right to left across the screen, as shown diagrammatically in Fig. 3.

It would be convenient to switch in 100μsec markers from the time calibrator to facilitate the count.

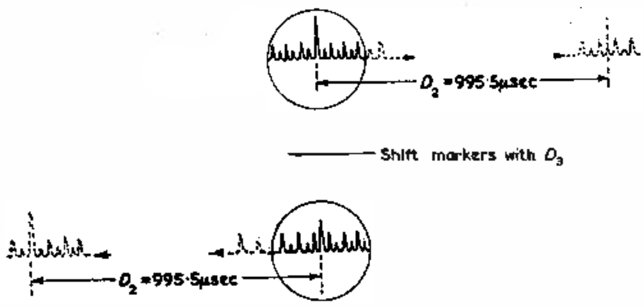

Fig. 3. Introduction of additional delay

As the trace is 'moved' a count is taken of the markers as they move across the screen, until the delayed pulse (F) appears. This is not a laborious procedure, as initially only the 100μsec markers are counted. Finally the delayed pulse from D_2 is aligned with the marker '995·5μsec', when D_2's delay will correspond to that figure. During the procedure the sweep is of course maintained at, say, 20μsec, so that the final alignment presents no difficulty.

One weakness of the arrangement is that there are three variable delays in circuit, all contributing their own jitter. With The PE31 VAD (Mk 2) instruments used the jitter per instrument is usually less than 1 part in 10 000: it is to be expected that the delayed pulse will jitter with respect to a time marker by about ±0·2μsec. It is usually possible to adjust the delay so that it jitters about a particular 0·5μsec marker, the accuracy approaching ±0·2μsec, which expressed as a percentage of the total delay is ±0·02 to 0·03 per cent. In practice accuracies of about ±0·015 per cent can be achieved. Better accuracy could be obtained if it were desired to calibrate equipment capable of generating a pulse of longer delay—probably better than 0·01 per cent in, say, 5msec.

A similar method may be used if it is required to measure accurately the length of a long pulse: the pulse is used to gate the time calibrator, the 'internal gate' being switched out of circuit.

In conclusion it may be mentioned that at the shorter time intervals excellent time 'resolution' may be obtained. Thus on the '10μsec' range, one division on the helical potentiometer corresponds to 10msec. Such small time increments cannot of course be obtained using equipment based on the 'crystal clock count-during-time-interval' technique, unless an extremely high frequency crystal (with associated counting difficulties) can be used. A further improvement is obtained if a 'Blose 3' type blocking oscillator is fitted as the delayed output pulse generator, as the rise-time is then less than 10mμsec.

International Planning Conference on Medical Electronics

The above conference was held at the Nouvelle Faculté de Médicine in Paris on 26 to 28 June, 1958, under the chairmanship of Professor Fessard (Prof. of Electroneurophysiology, Collège de France). The Vice-President and General Secretary was Dr. M. Marchal of the Laboratoire de Radiologie, L'Ecole des Hautes Etudes.

Some fifty delegates were present, of which the largest group (ten) came from the United Kingdom. The remainder were from France, Holland, Italy, Japan, Sweden, Switzerland, U.S.A., and W. Germany. Other countries, including the communist states, had been invited, but declined on account of the short notice given.

The proceedings were divided into two sections, one devoted to organizational matters (such as international co-operation and exchange of information), the other to the reading of thirty-three short scientific papers on subjects within the framework of Medical Electronics. This report is confined to the conclusions reached during the business meetings.

It was generally agreed that a successful full-scale scientific conference could not be arranged in time for next year, as originally planned. It was also agreed that giant conferences covering all aspects of diverse fields are often unsatisfactory, especially if held too frequently. Limited colloquia were considered to be more suitable for achieving the required results. Consequently, a second planning meeting will be held in Paris next summer, so that more definite arrangements for future conferences may be discussed. No attempt was made to strictly define the limits of 'medical electronics'.

Meanwhile, the objectives of international co-operation and co-ordination in medical electronics will be pursued by an interim committee composed of all those invited to attend the first Paris meeting. This committee elected the following interim Executive Committee, to co-ordinate their efforts and to engage in the detailed planning of future activities:—

President: Dr. V. K. Zworykin (U.S.A.), Vice-Presidents: Dr. M. Marchall (France) and Dr. C. N. Smyth (U.K.), Secretaries: Dr. C. Berkley (U.S.A.) and Dr. A. Rémond (France), Treasurer: Mr. B. Shackel (U.K.). The Executive Committee was authorized to appoint advisors as required, and chose immediately the following: Prof. O. Wyss (Switzerland), Dr. C. Sakamoto (Japan) and Dr. D. H. Bekkering (Holland).

With the assistance of the committee members in the various countries, the executive are to collect information on all organizations already related to this field, and will outline the scope of the planned organization. Interested persons and organizations in each country (including those not yet represented) will then be contacted and asked to meet and appoint fully representative secretariats on a regional basis. It is anticipated that a federation of these national groups will form the International Body visualized originally by Dr. V. K. Zworykin.

Copies of the new Bibliography on Medical Electronics (prepared by the Medical Electronics Centre of the Rockefeller Institute, and published by the Medical Electronics Group of the I.R.E. in New York—\$2.50) were issued to the full committee. They were requested to provide further appropriate references for inclusion in future editions, and to appoint co-ordinators (selected for language, geography and techniques) for this purpose. An offer by Dr. E. Van Tongeren on behalf of 'Excerpta Medica' to co-operate in developing a regular bibliography is to be explored. An abstracting service and a Journal of Medical Electronics will be kept in mind as a possible long-term development.

Another project to be initiated immediately, is the compiling of a list of experts, so that an index of advisors on the special techniques of medical electronics may be available. The listing of instrument manufacturers and of appropriate electronic equipment will also be considered.

Offers of help and facilities in connexion with future conferences were made by Dr. R. C. G. Williams and Mr. W. J. Perkins on behalf of the Institution of Electrical Engineers and the British Institution of Radio Engineers respectively.

An Improved Technique for Fast Multiplication on Serial Digital Computers

By M. Shimshoni*

It is suggested that, in multiplication on serial digital computers, a valuable saving of computing time can be achieved without using parallel multiplier circuits by considering only the significant digits of the multiplier. Types of calculation where this method may be applied with advantage are suggested and an example is given. The position of the binary point in the product of such a multiplication is considered and formulae for finding this position are given.

IN order to multiply two binary numbers, each of n digits, there are n conditional addition and shift operations that have to be performed[1]. The product of this multiplication will normally be obtained in two registers each n digits long; the register containing the most significant half of the product will be referred to as the A register, and the other as the B register.

If the n operations are carried out successively, in full serial manner, multiplication will take at least n word times. In this case multiplication will be a very slow operation compared with addition and other elementary operations that take only one or two word times. This type of multiplier is of very simple design, and is used in several computers, for example in the Stantec-Zebra and the Elliott 402.

As n is about 30 to 40, certain computers have 'built-in,' a parallel multiplier circuit, in which all the n operations of the multiplication are carried out simultaneously, and the whole multiplication takes only about three times as long as a simple addition. The resulting circuit is much more elaborate containing n adders, n delay circuits, n gates etc. This type of multiplier circuit is used in the Manchester University computer[2], in which $n = 40$.

Another form of multiplier circuit is used in a Ferranti computer. There the multiplier is divided into two parts, each of which multiplies the multiplicand in a parallel manner; the partial products are then added. In comparison with the fully parallel multiplier circuit, this method halves the number of adders required but approximately doubles the multiplication time[3].

It will now be shown how in certain cases the multiplication time can be cut down in serial multipliers, without the introduction of additional equipment.

In many practical cases neither multiplier nor multiplicand has n significant digits. If it is known that, in certain types of multiplication, the multiplier has never more than m ($m<n$) significant digits, the whole multiplication could then be carried out in only m word times. Moreover if one is not interested in more than n digits of the product, the process can be further simplified.

By suitable shifting let the multiplier occupy the m least significant digits of a word, and the multiplicand of length r ($r<n$) digits, occupy the r most significant digits of its word. If now multiplication proceeds normally for m steps instead of n, all the n most significant digits of the product will be in the A register, although the binary point of the product may be in an unusual position. At the end of the whole computation the binary point can be brought back to a more usual position by some shifting operations. Although the digits of the multiplier should preferably be concentrated at the least significant end, those of the multiplicand are not always kept at the most significant end; their position is mostly fixed by the desired position of the binary point in the product. The way the position of the binary point in the product is dependent on the positions of the binary points in its factors is shown in the appendix.

This type of faster multiplication has been used in a programme for the calculation of crystal structure factors on the Stantec-Zebra, with an overall saving of time of about 50 per cent. This calculation was specially suitable for this method, because the most frequent calculation was of the type $(hx+ky+lz)$, where h,k,l, were small integers that could be represented fully by 8 digits (including the sign digit), and x,y,z, were proper fractions that could be represented with 12 digits. h,k,l, were chosen as multipliers, so that the whole multiplication took only 8 word times compared with the 33 word times of a full multiplication.

Other obvious types of computation where this method might be used are matrix calculations. For the multiplication of two $r \times r$ matrices r^3 multiplications are needed. If during input the $2r^2$ factors of the matrices are stored after proper scaling, the overall multiplication time might be considerably reduced. This applies only if the order of magnitude of the factors is known. If this order of magnitude is not known, floating point techniques have to be used, and the whole computation becomes much lengthier. In general wherever the floating point method is used, the fast multiplication cannot be applied with advantage.

In the two computers with serial multipliers that have been mentioned above, fast multiplication instructions are as simple to write as normal multiplication ones. In other computers with a serial multiplier circuit which have no facility for fast multiplication, it should not be too difficult to modify them so that shorter multiplication instructions will be possible.

Acknowledgment

Acknowledgment is due to Dr. D. Rogers of University College, Cardiff, for some useful discussions.

APPENDIX

The first digit of a number serves usually as a sign digit. Therefore if two numbers, n, digits long, are multiplied one obtains a product $(2n-1)$ digits long, because one sign digit will serve the whole double-length number. As the two registers A and B have $2n$ digits, one of the digits on the B register (either the first or the last) is always 0. The B register will thus be regarded as containing only $(n-1)$ digits.

Consider two numbers a and b, with their respective binary points p and q places from their least significant ends. If a and b are fully multiplied by all the n operations, the binary point in the product will be

$$s'(B) = p+q$$

places from the least significant end of the B register, or

* *The Weizmann Institute of Science, Israel, formerly University College, Cardiff.*

$$s'(A) = p+q-(n-1)$$

places from the least significant end of the A register.

If a and b are multiplied by a shortened multiplication, in which only m digits of the multiplier are considered, there will be $(n-m)$ less shifts to the right, and the binary point in the product will be

$$s(B) = p+q+(n-m)$$

places from the least significant end of the B register, or

$$s(A) = p+q+(n-m)-(n-1) = p+q-m+1$$

places from the least significant end of A.

In the calculation of

$$t = hx+ky+lz$$

mentioned above, t was required to have its binary point 10 places from the least significant end. h,k,l, were integers 8 digits long, their binary point being 0 places from the least significant end. Thus in order to find the position p of the binary point in x,y,z, one has $s(A) = 10$; $q = 0$; $m = 8$. Hence:

$$10 = p+0-8+1;$$

$$p = 17.$$

The values of x,y,z, have to be stored with their binary points 17 places from their least significant ends. As x,y,z, can be represented with full accuracy by 12 digits after the binary point, these 17 places are more than sufficient for their storage.

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2. Robinson, A. A. Multiplication in the Manchester University High-Speed Digital Computer. *Electronic Engng. 25*, 6 (1953).
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An Analogue Computer for the British Transport Commission

The British Transport Commission have installed a special purpose analogue computer in the office of the Chief Electrical Engineer, British Railways Central Staff, for use in pre-determining train performance with the new forms of motive power which are being introduced under the railway Modernization Programme. The advance information obtained will be used to compile revised timetables, designed for the maximum utilization of the new equipment.

To carry out such a revision the information is required long before the trains enter service and is usually predetermined at the design stage, before the equipment is even manufactured. This is made possible by the precise characteristics to which modern traction equipment can be built, and by the knowledge that all the units produced of a particular type will reach a standard performance.

From known data of a type of locomotive or train, and details of a particular line—gradients, speed restrictions, and stopping places—it is a straightforward process to work out the time taken to travel from one place to another. The production of the information, however, can be substantially speeded-up by the use of a computer, and in particular by an analogue computer of the type which employs angular rotation or voltage to represent the magnitude of the physical quantities concerned—accelerating or decelerating effort, speed, time, and distance.

The analogue computer which is in use in the office of the Chief Electrical Engineer, British Railways Central Staff, was developed and built under the direction of Professor E. Bradshaw of Manchester University, in association with the Chief Electrical Engineer. In this computer voltages are used to represent the physical quantities concerned.

The operator with a train performance problem to solve draws out the tractive effort/speed characteristic of the locomotive or train, selects a suitable train resistance curve and a punched tape of the gradients of the route. The positions of all stations, speed restrictions, etc., are marked on a 24in chart on a distance base and braking curves drawn backwards from each station at which a stop is to be made, and from each speed restriction.

The results are displayed as a continuous record of speed against distance with an independent time trace.

Modified watt hour meters are employed to solve electrically the two basic equations:—

$$v = \int a.dt.$$

$$s = \int v.dt.$$

A stabilized voltage is applied to their voltage coils and to the current coils are fed voltages representing the two variables.

In the case of the first integrating meter the voltage applied to the current coil, representing net tractive effort, is derived from three sources:—

(a) A Variac type auto-transformer driven by the curve following mechanism of the tractive effort/speed curve, alternatively an auto-transformer adjusted to the appropriate braking rate.

(b) A Variac type auto-transformer driven by the curve following mechanism of the train resistance/speed curve.

(c) A transformer having a secondary winding tapped to provide turns ratios of 1: 2: 4: 8: 16: 32: 64.
This is the information on gradients which is punched on a tape in a binary coded form and picked up by knife edged roller contacts. Relays controlled by the roller contacts select the appropriate tappings, a total of 127 different combinations being possible using the seven sections of the winding, and auxiliary relays, also controlled by roller contacts, control the sign of the gradient voltage.

The angular velocity of the disk in this first integrating meter is a function of acceleration and its displacement a function of speed. The edge of the disk is painted in alternate black and white segments and light from a lamp built into the meter is reflected from the white sectors on to a photocell.

Signals from the photocells pass through a phase discriminator, in the case of the first integrating unit only, via shaping and differentiating circuits to uniselectors which drive the drum carrying the curves of speed dependent functions; this in turn drives the pen recording speed and imparts an angular rotation to the Variac type auto-transformer, the output voltage of which represents speed.

The voltage applied to the current coil of the second integrating meter is derived from the 'speed' auto-transformer and the rotation of this meter disk through a servo mechanism, similar to the one described above in relation to the first integrating meter, drives the distance chart forward together with the punched tape carrying the information on gradients.

A small synchronous motor is arranged to provide a time trace of sawtooth form by means of a cam following mechanism and pen.

One of the more interesting features of this computer is the automatic curve following device which operates as follows:—

A probe unit comprising two probes approximately 2mm apart straddles each curve of a speed dependent function—tractive effort, train resistance, motor current; a broad band of conducting, colloidal graphite is painted immediately below the curve and connected to earth.

The probe unit driving motor, a permanent magnet d.c. machine, which also drives the Variac type auto-transformer associated with that curve, is fed via two relays according to signals received from the probes. If both probes are on the graphite, i.e., below the curve, the motor winds them up and if both are off the graphite, i.e., above the curve, the motor winds them down. In this way the probes follow the curve as it is rotated by the uniselectors and in their normal position, astride the curve, the driving motor is disconnected from the supply.

As mentioned above, provision is made on the main chart drum for a third speed dependent function, motor current in the case of an electrically hauled train, and an automatic curve following mechanism drives a Variac type auto-transformer which provides a feed to an unmodified kWH meter. This effects the integration of I with respect to time and records the result on the usual decimal type counter. The product of current consumption and line voltage provides a continuous record of energy consumption which is of particular value to the railway operator.

Simultaneous Display of a Rectified Waveform and a Time Scale Using a Single Spot Oscillograph

By J. K. Grierson*, M.Sc.

THE use of a cathode-ray oscillograph in conjunction with 35mm film to photograph transient phenomena is standard practice. In this method the spot is deflected by the signal in the vertical (y) direction, from the horizontal (z) tube axis, and then photographed on 35 mm film which moves horizontally in the direction (x) at right-angles to the tube axis. It is sometimes desirable to photograph, simultaneously, a signal and a series of time marker pulses, using a single spot oscillograph. The apparatus to be described enables this to be done. In this, the incoming signal is rectified and the trace is returned to the baseline for short periods at any time when a time marker pulse is due. The time marker pulses are then added to the modified signal.

In the circuit of the apparatus (Fig. 1) the pulses (Fig. 2(a)), which are 'negative going', are fed to the pulse shaping and attenuating network in the grid circuit of V_1. V_1, V_2, and V_3 constitute a feedback amplifier. The time scale (now large and positive going, Fig. 2(b)) is prevented from over-shooting negatively by the crystal diode, MR_2. If the adding circuit, which consists of the 100kΩ resistor that connects the cathode of MR_2 and the anode of MR_3, and the 100kΩ resistor that connects the anode of MR_3 and the signal input terminal are considered, it is seen that:

(1) If no time scale pulse is present, a rectified signal waveform will appear at the anode of MR_3 to provide a baseline from which amplitude measurements may be taken.

(2) If a time scale is present, the signal will be swamped and the voltage at the anode of MR_3 will be that of the baseline (Fig. 2(c). The signal is shown in Fig. 2(d).

This behaviour is due to the fact that MR_3 will never allow the signal to go positive, hence, when a time pulse is present, the high amplitude of the pulse will overcome the signal even when the signal is at its maximum negative excursion. This brings the voltage back to the baseline, MR_3 preventing the overshoot.

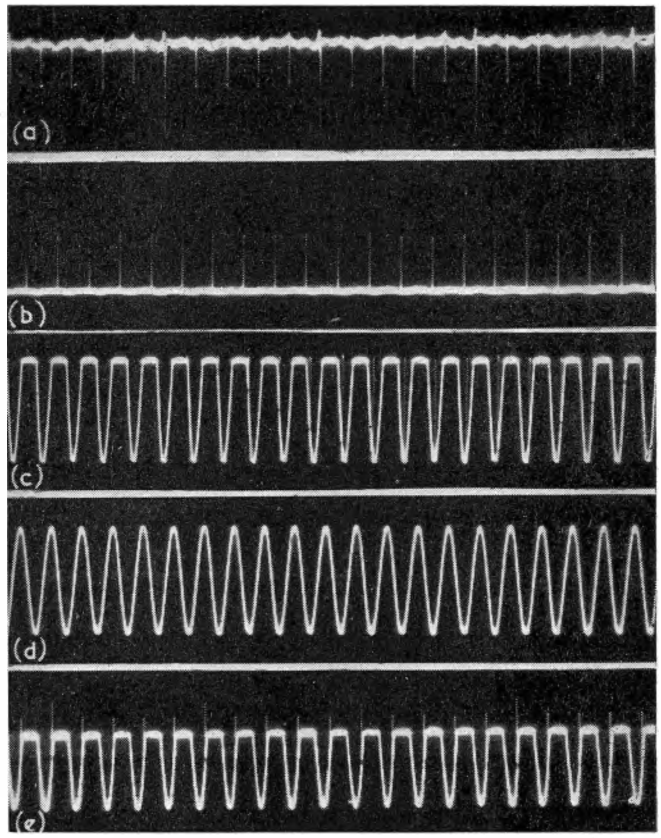


Fig. 2. The waveforms of Fig. 1

The cathode-followers, V_4 and V_5, together with the 10kΩ cathode resistors, constitute an adding circuit, the time scale being added to the output from the anode of MR_3 (Fig. 2(e)).

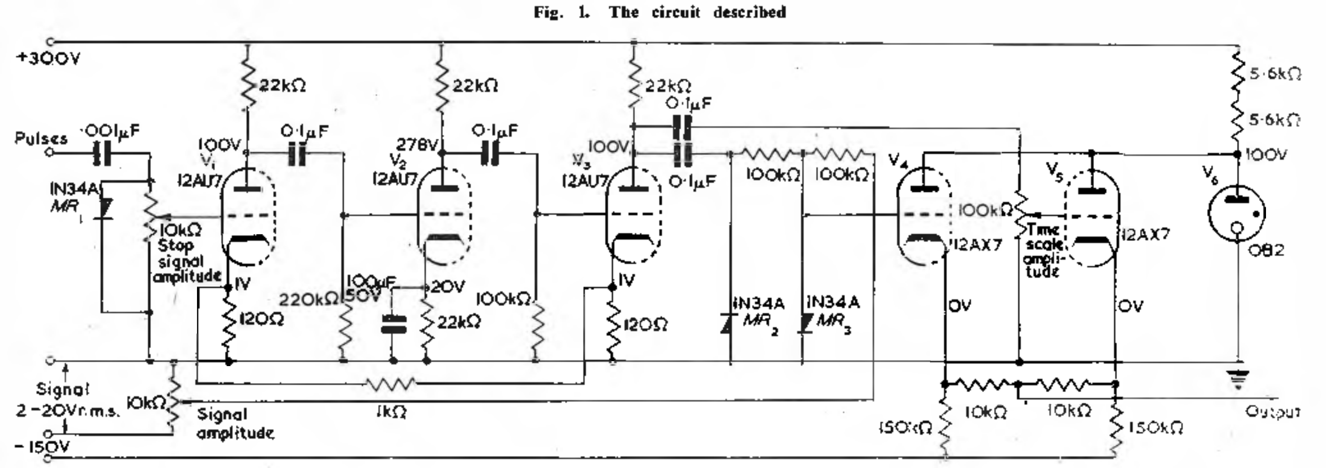
The final output is obtained from the electrical centre of the two 10kΩ resistors in the cathodes of V_4 and V_5.

Fig. 2 shows actual traces obtained at various parts of the apparatus, which is thus seen to work satisfactorily. The three controls, 'stop signal amplitude', 'signal amplitude' and 'time scale amplitude' enable the signal and time scale amplitudes to be varied independently without affecting the baseline.

The space between markers was 10msec and the frequency of the sine wave was 100c/s in the test photographs, but the apparatus is suitable for the whole range of audio frequencies.

* *Defence Research Board, Canada.*

Fig. 1. The circuit described



Short News Items

Following the formation of the Data Processing Section of the Electronic Engineering Association (formerly the Radio Communication and Electronic Engineering Association), conversations have been held with other associations interested in this field, either as manufacturers or users. The aim is to provide a clearing house for the exchange of information as a step towards co-ordinating technical committee work, research, development and application of the new techniques in industry. It is also hoped to provide a channel for the presentation of the British industry point of view in international discussions concerned with electronic automation and computation. So far, discussions have been entirely on a technical level. Meanwhile, the name Electronic Forum for Industry has been adopted by those attending the meetings. Organizations taking part so far, in addition to the Electronic Engineering Association, who were the conveners, are: B.E.A.M.A. (British Electrical and Allied Manufacturers' Association); B.R.V.M.A. (British Radio Valve Manufacturers' Association); M.T.T.A. (Machine Tool Trades Association); O.A.B.E.T.A. (Office Appliance and Business Equipment Trades Association); R.E.C.M.F. (Radio and Electronic Component Manufacturers' Federation); S.B.A.C. (Society of British Aircraft Constructors); S.I.M.A. (Scientific Instruments Manufacturers' Asssociation); T.E.M.A. (Telecommunications Engineering and Manufacturing Association). "The Electronic Forum for Industry will not assume the work or responsibilities of the individual associations," said Mr. C. Metcalfe, Chairman of the Data Processing Section of the Electronic Engineering Association, who presided at the first discussions. "Nor will there be any impingement on the work of the professional institutions. The Forum will be a co-ordinating body to help to guide industries through the growing complexity of electronics." Secretaries of the Forum are the Electronic Engineering Association, 11 Green Street, London, W.1.

The next standard course at the Harwell Reactor School on which places are available starts on 10 November and ends on 6 March 1959. The fee for the course is £250 exclusive of accommodation. Applications will be considered from overseas as well as the United Kingdom and this will be the seventeenth standard course at the school, where the 1 000 students who have attended to date include 231 from 31 other countries. Application forms and further details can be obtained from the Principal, Reactor School, Atomic Energy Research Establishment, Harwell, Didcot, Berkshire.

An international team of 21 Scientific Secretaries from 13 countries has been appointed for the Second United Nations International Conference on the Peaceful Uses of Atomic Energy, to be held in Geneva from September 1-13. All have arrived at U.N. Headquarters. The scientific secretaries are specialists in physics, chemistry, biology, biochemistry, medicine, radiology and other fields related to the use of nuclear energy. They are from Argentina, Belgium, Canada, Czechoslovakia, France, Israel, Italy, Japan, Spain, the U.S.S.R., the United Arab Republic, the United Kingdom and the United States. They have been assigned to work at U.N. Headquarters, and later at Geneva, on specific subjects that will receive major attention at the Conference: nuclear fusion; fission reactor engineering; physics; biology and isotopes; and raw materials, mining and chemistry. The titles of 2 247 papers to be presented to the Conference by 41 Governments have been announced. It is expected that other titles will be announced in the near future. The lists containing these titles may be consulted in the Library of the United Nations Information Centre, 14-15 Stratford Place, London, W.1.

The British Amateur Television Club will be holding its fourth Amateur Television Convention on Saturday, 6 September, from 10 a.m. to 7 p.m. in the Conway Hall, Red Lion Square, Holborn, London, W.C.1. There will be displays of amateur-built television equipment in operation, and among the attractions will be pictures received from the Club's outside broadcast van, as well as demonstrations of colour television. Both members and non-members of the Club will be welcome: the charges for non-members are as follows: All day 5s., admission after 2 p.m. 2s. 6d.

Over 2 000 members and guests of the Institution of Electrical Engineers attended a Conversazione at the Royal Festival Hall at the end of June. They were received by Mr. T. E. Goldup, C.B.E., President, and Mrs. Goldup, assisted by Mr. S. E. Goodall, President-Elect, and Mrs. Goodall, and members of the Council of the Institution. About 70 exhibits and demonstrations of advances in science and engineering and their applications were shown. Noteworthy among these was a demonstration of closed circuit colour television. Demonstrations included the location of lightning flashes by cathode-ray direction finding, a model of *Sceptre* in which success was achieved in stabilizing pinched toroidal discharge, and there were a number of exhibits relating to the generation of electrical power by nuclear reactors.

The Fifth National Symposium on Reliability and Quality Control in Electronics will be held from 12-14 January 1959 at the Bellevue-Stratford Hotel, Philadelphia, Pa., U.S.A. Although this annual event has hitherto contained papers almost exclusively of U.S.A. origin, the organizing Committee is anxious to make it clear that papers from outside the U.S.A. will be most welcome. Mr. R. Brewer of the Research Laboratories, The General Electric Co Ltd, Wembley, Middlesex, has been appointed Publicity Area Chairman for United Kingdom and Western Europe and can supply further information about the Symposium and the procedure for submitting papers.

Ferranti Ltd have announced plans to consolidate, within the next few months, the operations of Ferranti Electric Ltd, Toronto, and Packard Electric Co Ltd, St. Catharines, under the name of Ferranti-Packard Electric Ltd with headquarters in Toronto. The new company will manufacture all products now made by both companies. Mr. T. Edmondson, now President and General Manager of Packard Electric Co Ltd, will be appointed President and Chief Executive Officer. Dr. J. M. Thomson, now President and General Manager of Ferranti Electric Ltd, will be Chairman of the Board of Directors of the new company. The manufacture of power transformers, the production capacity for which is equal to that of Ferranti Ltd, will be concentrated in Toronto while instrument transformers, meters and relays will be made in St. Catharines. The electronics and ordnance operations now conducted by Ferranti Electric will continue in Toronto and the manufacture of distribution transformers and meters will continue at the Packard Electric plants in St. Catharines and Three Rivers.

The Central Office of Information has recently produced a film entitled *The Inquisitive Giant* which tells the story of the Jodrell Bank radio telescope. This is available for hire from the Central Film Library, Government Building, Bromyard Avenue, Acton, London, W.3, the Scottish Central Film Library, 16-17 Woodside Terrace, Charing Cross, Glasgow, C.3, and the Central Film Library of Wales, 42 Park Place, Cardiff.

Welwyn Electrical Laboratories Ltd of Bedlington, Northumberland, recently held an Open Day to celebrate the 21st anniversary of the Company.

LETTERS TO THE EDITOR

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

A Transistor d.c. Convertor

DEAR SIR,—In a recent article[1] in ELECTRONIC ENGINEERING I was interested to see the description of a transistor d.c. convertor. While reference was made to the circuits published by Uchrin and Taylor, and Royer, the circuit shown was in important respects different from these and not unlike one described by the writer[2,3].

Your authors have used a common-base circuit. The common-emitter circuit described by Royer is probably used more commonly, and there seems to be in the literature inadequate appreciation of the advantages which can accrue from the use of the common-base configuration especially when the circuit is arranged to take full advantage of its properties. Indeed, when the common-base circuit is mentioned its use is often deprecated, and, the equation for its oscillation frequency is either vaguely or incorrectly stated.

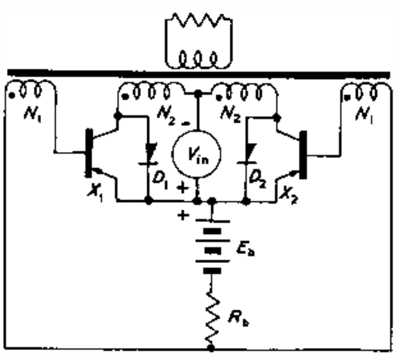


Fig. 1. Modified Royer common-emitter convertor

The usual common-emitter circuit is shown in Fig. 1. The elements E_B and R_B together constitute a current bias source which is necessary to ensure oscillation and to prevent overdrive. If the circuit is used as an invertor for power-supply applications the source V_{in} will be fixed in voltage and it can fill the role of E_B as well. In other applications V_{in} may be a signal source and therefore a separate bias source E_B may be desirable. In this circuit each transistor alternates between a conducting and a cut-off state. When one is cut-off the other is conducting and vice versa. The conducting transistor, if its collector-emitter voltage drop is neglected, causes the voltage V_{in} to be applied across the N_2 turns of one-half the collector winding of the transformer. Hence $d\phi/dt = V_{in}/N_2$. From this it follows that the oscillation frequency is given by $f = V_{in}/(4N_2\phi_m)$, where ϕ_m is the saturation flux of the core. The emitter windings of N_1 turns play no part in determining the frequency although, of course, they are essential to the circuit operation. The diodes D_1 and D_2 may be included to clip transient peaks from the leading edges of the output square wave.

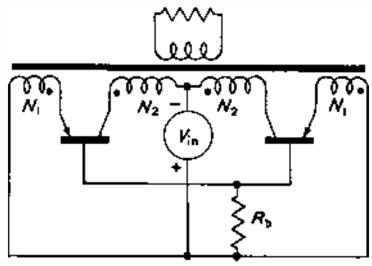


Fig. 2. Differential common-base convertor

The common-base circuit of Fig. 2 differs from that of Fig. 1 in several important ways. When a transistor is conducting in this circuit the source voltage V_{in} appears across N_2 and N_1 turns in series. However, in order that the windings of N_1 turns may serve their purpose of holding one transistor in the conducting state and the other in the cut-off state, the N_2 and N_1 turn windings must be differentially connected in the series circuit through the collector and emitter of the conducting transistor. Thus the source V_{in} is connected effectively across N_2-N_1 turns and the frequency of oscillation is given by $f = V_{in}/(4\ (N_2\text{-}N_1)\ \phi_m)$. A higher frequency of oscillation is obtained using the same components as in Fig. 1. This may be advantageous in some applications; but perhaps more important is the fact that in this circuit it is possible to limit the drive applied to the conducting transistor at high values of V_{in} without affecting adversely the square output waveform. The resistor R_B is included for this purpose. Resistors cannot be inserted in series with the bases of the transistors in Fig. 1 without interfering with the output waveform. Experimental tests on identical units show that the differential circuit of Fig. 2 with a suitable value for R_B will operate quite satisfactorily at values of V_{in} which cause failure of the transistors due to overheating when they are used in the circuit of Fig. 1. Reliable oscillation is obtained without resort to the use of a current-bias source such as is shown in Fig. 1.

I believe that these remarks may serve to clarify the description given by your authors, particularly in regard to the frequency equation which they have given. In this equation the symbols V and N are not adequately defined in relation to the convertor circuit.

Yours faithfully,

C. H. R. CAMPLING

Department of Electrical Engineering,
Queens University,
Kingston, Ontario.

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2. CAMPLING, C. H. R. Differential Magnetic Multivibrators. AIEE Conference Paper No. 57-769 (1957).
3. CAMPLING, C. H. R. Magnetic Inverter Uses Tubes or Transistors. *Electronics 31*, 158 (Nov. 1958).

The Author's Reply:

DEAR SIR,—We are indebted to your correspondent, Mr. Campling, for presenting in greater detail the general formula quoted for the oscillator frequency in common base connexion. An alternative approximate expression to the one he quotes is $f = \dfrac{V + dV}{4\,\phi_s N}$ where dV is the voltage fed back into the emitter circuit. However, since ratio V/dV as determined by the transformer winding ratio is usually large, the resulting change in frequency is small and for a convertor would be unimportant to a first order. In quoting the general formula, therefore, the authors chose to ignore second order effects, such as loss of voltage across the switching transistor and the one above, and for this reason have defined V as the applied voltage and not the supply voltage.

However, we acknowledge that the formula for f has been incorrectly given and should be a direct inversion of the expression stated for the period of oscillation. The expression below, therefore, represents a more accurate form of the formula.

$$f = \frac{V \times 10^8}{4B_s\ N.A.}$$

where f is the frequency (c/s)

B_s is the saturation flux density (lines/cm^2)

A is the core area (cm^2)

We regret we were unable to quote references to Mr. Campling's papers in our article, but these were not to hand when the script was prepared.

Yours faithfully,

J. B. WARMAN

Telecommunications Section,
Research Laboratories,
Siemens Edison Swan Ltd.

Voltage Stabilizer Principle

DEAR SIR,—With reference to the correspondence on the above subject in your April issue I should like to make the following comments.

(1) Messrs. Billington and Chakanovskis ascribe the original suggestion as to the use of positive feedback in electronic stabilizer circuits to the M.I.T. volume "Vacuum Tube Amplifiers" edited by Valley and Wallman, and published in 1948. This may, in fact, be the first published reference to the subject; I should, however, like to mention that the patent covering the positive feedback circuit described in Mr. Pacak's reference 2 was applied for in 1945.

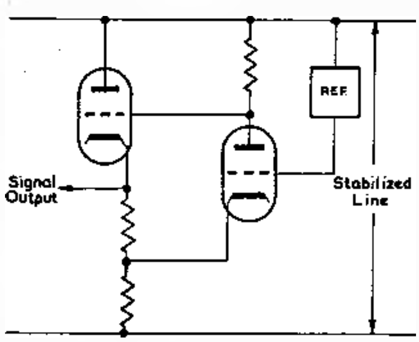


Fig. A. Simplified drawing of Pacak's circuit

This circuit is drawn in a simplified form in the accompanying Fig. A. It will be observed that the positive feedback circuit is almost identical to that of Messrs. Billington and Chakanovskis' Fig. 2.

Connexion of the cathode-follower anode to the positive stabilized line, however, makes the positive feedback independent of variations in unstabilized input voltage; the feedback can therefore be adjusted to give exact compensation for changes in input voltage and load current.

Output voltage is adjustable by altering the value of the common cathode resistance, which is analogous to the 'foot' resistance discussed in Messrs. Billington and Chakanovskis' article; to maintain exact compensation feedback must be adjusted at the same time. The common cathode resistance can be extended indefinitely; many practical difficulties which otherwise arise in stabilization of potentials above about 1kV are thus avoided (the circuit has been used experimentally up to 30kV).

With ideal components, there would be no point in adding a further amplifying stage and, in practice, an output dynamic range of about 20V can be obtained for variation in stabilized voltage of the order of 1 in 10^4. The addition of an amplifier as recommended by Messrs. Billington and Chakanovskis, between the positive feedback stage and the series control valve, would make the design and adjustment of the former less critical, although possibly introducing its own problems.

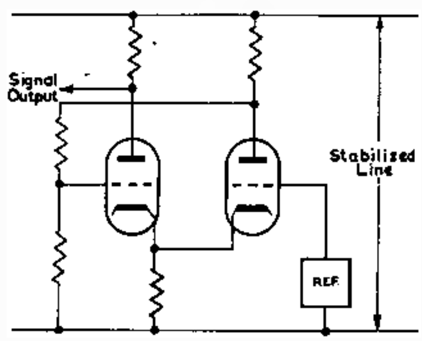


Fig. B. Simplified drawing of the Valley-Wallman circuit

(2) It is not clear why Messrs. Billington and Chakanovskis state in their article that the Valley and Wallman circuit does not use positive feedback, since the only essential difference between it and the double triode circuit of Mr. Pacak's Fig. 8 is that in the latter case the output is fed from the left-hand grid to a separate amplifier whereas in the former a similar amplification is performed in the left-hand triode. This can perhaps be made clearer by redrawing the basic elements of the two circuits as shown in the accompanying Figs. B and C.

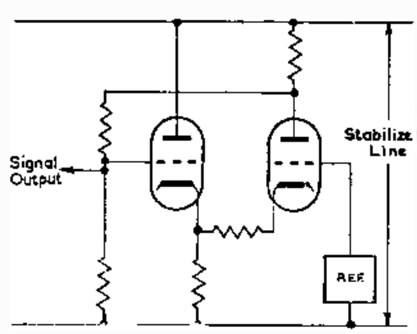


Fig. C. Simplified drawing of Pacak's Fig. 8

Furthermore, if the double-triode portion of Messrs. Billington and Chakanovskis' Fig. 1 is redrawn in a similar manner to facilitate comparison (Fig. D), it will, I think, be apparent that, if the gain is appreciable and second order effects due to variations in h.t. potential can therefore be neglected, the action of their circuit is essentially the same as that of Figs. B. and C.

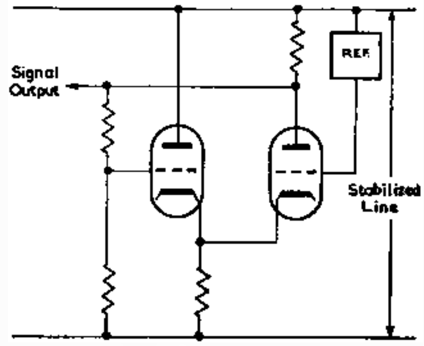


Fig. D. Simplified drawing of the Billington and Chakanovskis circuit

(3) With reference to Mr. Pacak's discussion of the apparently paradoxical effect whereby a corrective action takes place with no permanent change in stabilized voltage, I would suggest that his explanation would hold good even if no physical capacitance were present in the circuit, the time constant which he postulates then being realized by electron transit time in the valves.

Yours faithfully,

R. B. MACKENZIE,
Richmond, Surrey.

The Authors' Reply:

DEAR SIR,—We would like to thank you for your clear and fair appraisal of the matter. We must apologize for not including a reference to your contribution in our original paper or even in the correspondence, though Mr. Pacak corrected our error. While in no way excusing ourselves, you will appreciate that these situations can easily arise when the work is part of a technology which is not the chief concern of the authors. Our reason for publishing was stated in the second paragraph of our letter. The work was done at a time when reasonably priced commercial equipment was hard to come by in Australia, and it has led to a unit which is being made in this country.

In acknowledging each of your points (particularly our misinterpretation of Valley and Wallman's Fig. 11.58) we would mention only one matter. Connexion of the cathode-follower to the unstabilized line was made for economic reasons and is discussed at the beginning of the succeeding paper. Wider operating range is obtained at some sacrifice of performance.

Yours faithfully,

C. BILLINGTON

E. CHAKANOVSKIS

Division of Industrial Chemistry,
Commonwealth Scientific and Industrial Research Organization,
Melbourne.

BOOK REVIEWS

Electron-Tube Circuits

By Samuel Seely. 2nd Edition. 695 pp. 250 figs. Demy 8vo. McGraw Hill Publishing Co. Ltd. 1958. Price 81s. 6d.

THE second edition of this well-known and useful book has grown considerably as compared with the first edition in quality of production, in size, in range of subjects treated and in price. Much has been rewritten and several entirely new sections have been introduced, particularly on circuits using semiconductor devices.

As in the first edition, the basic physics of components is dealt with extremely briefly. What is treated as fundamental, is the operation in circuits of components with certain specified characteristics, rather than the origin of the characteristics themselves. This would not appeal to the physicist but is not intended to do so: it is entirely desirable for the radio- and electrical-engineering students for whom it is designed.

The general plan of the book has not been much changed. As before, about a third of it is devoted to amplifiers and a quarter to oscillators, including two chapters on sweep generators. The excellent sections on rectifiers and power supplies have not been altered, but have been brought forward to Chapters VI and VII instead of XII and XIII—a definite improvement.

The curious omission of push-pull amplifiers in the first volume has been remedied and complete new chapters on untuned power amplifiers, tuned potential amplifiers and tuned power amplifiers have been written. The section on oscillators has been partially rewritten with some gain to the student's ease of comprehension and an extensive section added on methods of modulation.

Apart from the larger changes, much trouble has been taken to improve the general layout. Even where there has been little or no rewriting of the text, new blocks have been made to give improved figures.

My only real criticism is of the sections explaining the physics of semiconducting devices, which seem to me to be far from clear. The "explanation" of the action of the point-contact diode on pages 36-37 in particular, is likely to be meaningless to someone meeting semiconductors for the first time. A less serious complaint is that, while I have myself often said "a large d.c. current" it does not look well in print. Neither of these points is likely to impair the value of the book to its intended readership. The new edition is well worth the great care and trouble which have evidently gone into it, and considering its size and quality the increased price is exceedingly reasonable.

J. H. FREMLIN

Television in Science and Industry

By V. K. Zworykin, E. G. Ramberg and L. E. Flory. 300 pp. 80 figs. Demy 8vo. John Wiley & Sons Inc. Chapman & Hall Ltd., London. 1958. Price 80s.

THE adaptation of closed-circuit television to industry is now becoming widespread, and it is surprising that we have had to wait so long for a book of this type.

The book is divided into five main chapters. The first of these comprises a historical record covering the last 100 years or so, while the last chapter boldly forecasts the effects of television on the lives of future generations. Although neither of these chapters will be of immediate use, either to television engineers or to prospective users seeking to educate themselves on a new medium, the average reader will find much to interest him in these pages.

The remaining three chapters are devoted to Applications, Equipment and Achievements. The chapter on Applications gives a comprehensive picture covering the majority of potential uses to which closed-circuit television may be put. The subjects include safety, elimination of undesirable or dangerous occupations, economy, co-ordination, inspection, film-making, medicine, education, advertising, security and military uses.

Chapter III describes the various items of equipment commonly used in closed-circuit applications, black and white, colour, and stereoscopic apparatus all being covered. While the authors have wisely devoted little space to receivers, a considerable section has been included on camera circuits and on the principles of various camera tubes. It is difficult to see the particular reader at which these pages are aimed. While clear to the television engineer, the few paragraphs describing the N.T.S.C. colour system for instance, or the brief description of interlaced scanning systems in Chapter I, can hardly augment the prior knowledge of the less specialized reader. Apart from a few such cases of overcondensed descriptions, all of which are amply covered in contemporary publications, chapter III contains much valuable information not readily available elsewhere. In particular, the sections on microscopy and transmission are most useful and in the latter case even more material, particularly on switching methods, could have been included with advantage. A comprehensive list of equipment manufacturers is given in this chapter and should prove of value to potential users.

Chapter IV devotes over 100 pages to an impressive collection of achievements in the field of closed-circuit television. The authors must have devoted considerable time and effort to assembling the material for this chapter. The subjects covered include gauge and furnace monitoring, security applications, underwater cameras and aeronautics, banking, mining, education, medicine, and the surveillance of processes in factories ranging from steel mills to transistor plants.

An appendix giving a detailed comparison of Vidicon and flying spot techniques as applied to microscopy will prove useful to equipment designers and others interested in this field.

A major omission for the reader with no electronic background is the absence of any information on the degree of skill required by operating staff or on the problem of maintenance.

This book should not be overlooked, either by television engineers, or by those who wish to obtain a general background in order that they can assess the value of this new medium in relation to their own particular problems. The number and quality of the illustrations are above average.

N. N. PARKER SMITH

Magnetic Tape Recording

By H. G. M. Spratt. 319 pp. 50 figs. Demy 8vo. Heywood & Co. Ltd. 1958. Price 55s.

Magnetic Recording Handbook

By R. E. B. Hickman. 176 pp. 111 figs. Demy 8vo. 2nd Edition. Geo. Newnes Ltd. 1958. Price 21s.

Techniques of Magnetic Recording

By Joel Tall. 472 pp. 60 figs. Demy 8vo. Macmillan & Co. 1958. Price 55s. 6d.

HERE we have three books dealing with the same subject. Scanning the many references in these works, one gains the impression that a critical survey of current papers on this art would have resulted in a single book of greater interest than follows repetition of already well-trodden ground. Mr. Spratt says that S. J. Begun's book (*Magnetic Recording*, Murray Hill Books Inc., 1948) "sets a standard in this field difficult to meet," and one must therefore conclude that these more recent works are written with the object of covering applications, techniques and equipment not known at that time.

In Mr. Spratt's case, all that S. J. Begun has said is said once again, in a somewhat more precise manner so far as exactitude can be assigned to an art in which the fundamental materials are subject to constant change. The book has many sections of great interest, especially those on the manufacture and testing of tapes, data recording and storage, machine tool control, recording of television signals and other modern developments based on magnetic recording methods. Although some continental machines are described, it is a pity that no details are given of the outstanding American Ampex recorders. To fully cover the ground, one might expect the best products of each country to have been included. The bibliography is quite adequate and the appendices dealing with BSI standards and

CCIR recommendations are useful. If the reader is an engineer, Mr. Spratt's rather expensive book can be strongly recommended. It is excellently produced and illustrated, written in a scholarly style, and completely authentic. In particular, the basic principles of magnetism and the performance of magnetic tapes are so completely presented that this work may well come to be regarded as the standard text on magnetic recording in this country.

Turning to the book by R. E. B. Hickman, now in its second edition, this is just the answer for the enthusiastic amateur. The contents are deceptively skilfully condensed so that for 21s. the reader receives excellent value. The emphasis is on practical details of construction, maintenance and usage, many complete circuits being shown. Fault analysis and servicing are accorded a very comprehensive section, and the bibliography is good. Few mathematics are employed and for an exposition of the art in a very readable yet satisfying form, this book is extremely good value at the price.

We are told that Mr. Joel Tall is a "consultant in the detection of falsified recordings" and has taught tape recording and editing. We find therefore that a great part of his book is given to editing, recording and re-recording techniques. In fact, most of the contents are devoted to television recording, information, medical and legal recording, as well as broadcast practice using tapes. The book is then more of a guide to these less-understood facets of the art, and in these respects the production and especially the profusion of illustrations is most creditable. Mr. Tall illustrates everything; on page 104, there is a photograph of Prof. Kellog recording frog songs. Perhaps frogs do sing in the United States. Näive though some of the ideas are, this book is full of sound practical advice and can be rated as a real acquisition to the literature of magnetic recording practices. Chapters VII to XX bristle with hints and tips, the historical introduction and bibliography are most detailed and Mr. Tall is careful to give credit where due regardless of nationality. An unusual book and one to be recommended; indeed, much of the information is not to be obtained elsewhere.

ALAN DOUGLAS

Programming for an Automatic Digital Calculator

By K. H. V. Booth. 238 pp. Demy 8vo. Academic Press Inc., New York. Butterworths Scientific Publications, London. 1958. Price 42s.

THE rapid development of electronic computers has resulted in a great variety of machines now being on the market. This in turn has created a demand for information on programming techniques for the benefit not only of potential users of these machines but also of consultants of firms considering the installation of computing equipment. The ideal book would be a kind of "comparative anatomy" setting out the essential features common to all machine codes as well as those particular to any one machine, and discussing their merits from the point of view of intended applications. This book has yet to be written, and readers have to be satisfied with the second best thing—a series of books explaining the code and programming technique for special (real or imaginary) machines. Under the circumstances, any new member of this series must be welcome, and it is in this context that Mrs. Kathleen Booth's *Programming for an automatic digital calculator* is of value. Essentially, the book is a Programming Manual for the APEXC Computer installed at Birkbeck College, London University, and it contains a large number of programmes and subroutines written for this machine in full detail. Apart from information on the technique of programming this particular machine, many hints on numerical analysis are given which will be of value to the mathematician who would like to know more about the actual techniques which are commonly used in computers. The techniques described include methods for calculating reciprocals, square roots, trigonometrical functions and the solutions of linear simultaneous equations.

A section of particular interest describes the application of computer methods to language translation, a field in which the Booths are pioneers. The actual programme given is for translating French into English.

Several misprints were noticed by the reviewer. For example on page 8 a form of the Greek letter ν has been substituted for the letter v in 6 places. Again, on page 25, a subscript r has been omitted in a mathematical expression, the result being rather confusing.

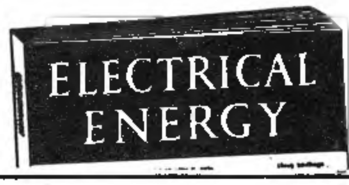
These minor blemishes do not detract from the value of the book which is clearly and carefully written and should find a permanent place in the library of any organization concerned, or expecting to be concerned, with the design or application of electronic digital computers.

C. F. GRADWELL

Universal Decimal Classification-Abridged Trilingual Edition

516 pp. Demy 4to. British Standards Institution 1958. Price £6 6s.

THE three language (German, English, French) edition of the Universal Decimal Classification, which is to rank henceforth as the International Standard Abridged Edition of U.D.C., has been published in this country as B.S. 1000B: 1958. This trilingual edition, with its outline of U.D.C. schedules corresponding closely to B.S. 1000A is intended primarily for classifiers in all branches of documentation and library work, but it will also, no doubt, prove useful to technical translators who do not always find the ordinary dictionary adequate for their wants.



ELECTRONIC EQUIPMENT

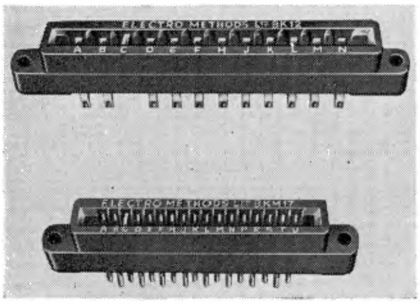
A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

PRINTED CIRCUIT CONNECTORS

(Illustrated below)

Electro Methods Ltd, Caxton Way, Stevenage, Hertfordshire

ELECTRO METHODS have now extended their range of precision printed-circuit connectors by the introduction of series 'BK' and 'BKM'. These have been produced to meet the demand for a connector which will accommodate a printed-circuit card based on 0·1in grid. The series 'BK' with contact-centres 0·2in are available with 8 and 12 contacts. These connectors, which are interchangeable, have end-fixing and polarizing-pins which can be fitted in any position. A choice of contact terminal is available.



In addition to the new series 'BK' and 'BKM', the series 'K' with 6, 12, 18 and 22 contacts (contact-centres ·156in) and the series 'KM' with 14, 21, 31 and 37 contacts (contact-centres ·078in) are also available.

All mouldings are mineral-filled melamine, conforming to B.S.S. 1322, and provide high arc-resistance and high dielectric and mechanical strength. Pin-contacts are made of brass, and socket-contacts of spring-tempered phosphor-bronze. Both are precision-machined and gold-plated for low contact-resistance, for the prevention of corrosion and for ease of soldering.

HYDROGEN BELL FURNACES

(Illustrated next column)

Royce Electric Furnaces Ltd, Walton-on-Thames, Surrey

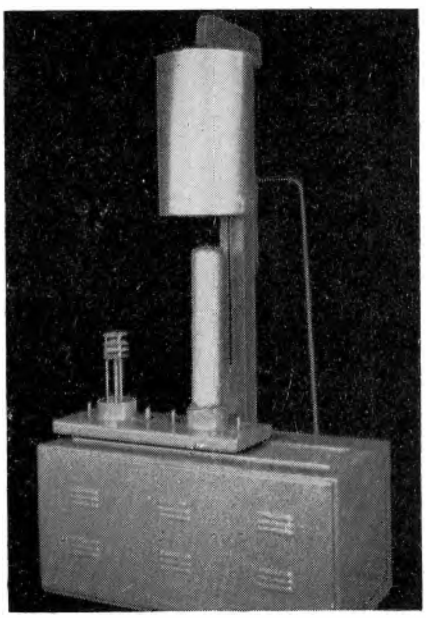
TYPICAL of the number of small furnaces specially designed and manufactured for individual requirements in the electronic and allied industries is the hydrogen bell furnace supplied by Royce Electric Furnaces Ltd.

The furnace illustrated has a work space of 5in diameter by 13in high and is suitable for operation at temperatures up to 1 000°C. Heating elements of 80/20 nickel chromium surround the heating chamber giving a uniform heat transfer throughout. To give a long service life the heating elements operate from the low voltage secondary of a stepdown transformer permitting the use of a heavy section element. The body is insulated within the cylindrical casing reducing heat losses to a minimum.

For operation with protective atmospheres, heat-resisting steel gas-tight retorts are provided, complete with sealing rings to engage in the base. Gas is introduced through the base, leading to the top of the retort to completely purge oxygen from the work chamber. Two identical bases are mounted upon a traversing trolley, each base complete with gas seals, heat-resisting steel heat baffles, protective gas inlet and outlet.

The charge to be treated is stacked upon the base and the gas-tight retort placed in position. When the chamber is purged of oxygen the heated bell furnace is lowered over the retort. During the heating cycle, the second base may be loaded up and purged. When the heating cycle of the first base is completed the heated bell is lifted and the base trolley traversed to bring the second base to the heating position. The bell, retaining its sensible heat, is then lowered over the second retort while the first charge is allowed to cool in the protective atmosphere. Thus a continuous operation may be maintained and as the furnace remains at temperature power consumption is minimized.

Hoisting gear for the furnace is counterbalanced and operated by a double acting air cylinder supplied complete with operating valves and speed control. For temperature control, two fully automatic temperature controllers are provided with a thermocouple in the heated bell and one in each base. A changeover switch is fitted for use with either base. The whole is mounted upon a steel stand having louvred removable panels providing convenient accommodation for the switchgear and transformers.



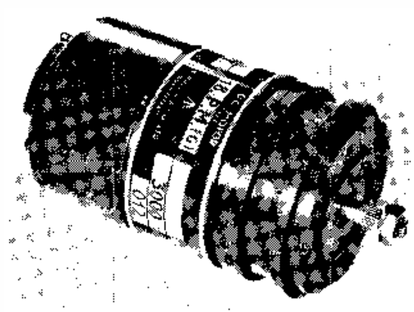
Larger bell furnaces with chamber sizes up to 3ft diameter by 5ft high are also supplied as standard for operation at temperatures up to 1 150°C in protective atmospheres.

PERMANENT MAGNET D.C. MOTORS

(Illustrated below)

R. B. Pullin & Co. Ltd, Phoenix Works, Great West Road, Brentford, Middlesex

THIS new range of permanent magnet d.c. motors is made to the international size 18 and is known as the type PM.18. It is available for 12V or 28V operation governed or ungoverned. As



they are specially insulated, the units are capable of withstanding total temperature up to 170°C and the materials are chosen to ensure a high degree of corrosion resistance.

A version of this motor, known as type 18PMT, is also available and is a highly sensitive, low ripple d.c. tachometer generator with a rated output of 16·7V per 1 000rev/min and a maximum linearity error at 5 000rev/min of 0·1 per second.

THERMALLY COMPENSATED OSCILLATOR

(Illustrated above right)

Automatic Telephone and Electric Co. Ltd, Strowger House, 8 Arundel Street, London, W.C.2

A NEW type of thermally compensated crystal oscillator has recently been patented by the Automatic Telephone and Electric Co. Ltd. The oscillator is at present available for operation on any spot frequency between 4 and 16Mc/s and achieves a stability of ±5 parts per million over a temperature range of −20°C to +70°C with respect to any specified frequency.

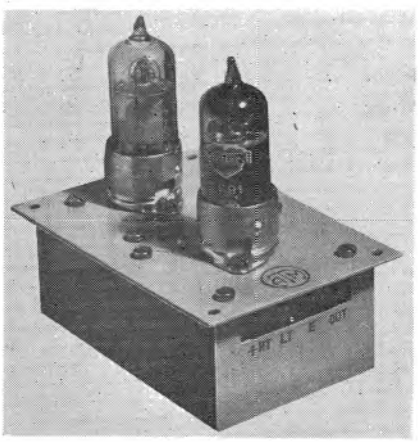
The high degree of stability is achieved—without resource to crystal ovens, thermostats or other current-consuming devices—by a unique method whereby the oscillator, which is essentially of a conventional type such as the Pierce, also has a temperature-compensating network controlled by a thermistor mounted inside the crystal envelope.

The complete oscillator unit is available as a small sub-assembly which can easily be incorporated in a wide variety of equipment during the design stage. The current consumption (8mA at 230V d.c. and 0·3A at 6·3V) is no more than that of any other type of oscillator.

The advantages of this type of oscillator are expected to appeal particularly to designers of mobile v.h.f. radio equipment where increasingly higher stabilities over wide temperature ranges are required due to closer channel spacings.

The compactness and stability of the unit are also expected to interest manufacturers of test equipment, and with this in view A.T. & E. are now investigating the possibility of applying the same techniques to oscillators operating on lower frequencies, as well as to those using battery type valves and transistors.

These units are ready for operation within a minute or two from switching on.



SERVO ELEMENTS

(Illustrated above right)

Muirhead & Co. Ltd, Beckenham, Kent

TO meet the increasing demand for low weight, small size components, Muirhead have produced the size 08 servomotor. Conforming to the standard 08 frame, this miniature motor has an overall diameter of 0·750in, a length of 0·963in from front face to ends of connexion tags, and a shaft extension of 0·156in. The shaft is splined in the form of an involute pinion of 13 teeth 120 d.p., 0·1245/0·1240in o.d., 0·1083/0·1078in p.c.d., 20° pressure angle. The body material is black dichromate finished stainless steel. The motor is ideal for use in packaged systems and miniaturized servomechanisms. It has a no-load speed of 6 500rev/min and a stalled torque of 0·1oz in and is designed for 26V 400c/s operation.

Another new product is the size 11 Linvar. This small size, low weight instrument, housed in a standard Bu-Ord size 11 synchro frame, has a voltage output proportional to rotor angle rising to 42·5V. The linear relationship of $\pm\frac{1}{4}$ per cent extends over the range ±85° from the zero output position, and is achieved by a special patented design of the rotor and stator laminations which gives a smooth variation of output voltage devoid of any fluctuations due to tooth ripple. The

Linvar can be applied to satisfy the requirements of analogue computation, remote control or indication, and may be operated by mechanical measuring devices, e.g. strain and deflexion gauges, weighing machines etc., where an a.c. output voltage varying linearly with deflexion is required.

V.H.F. VALVE-VOLTMETER

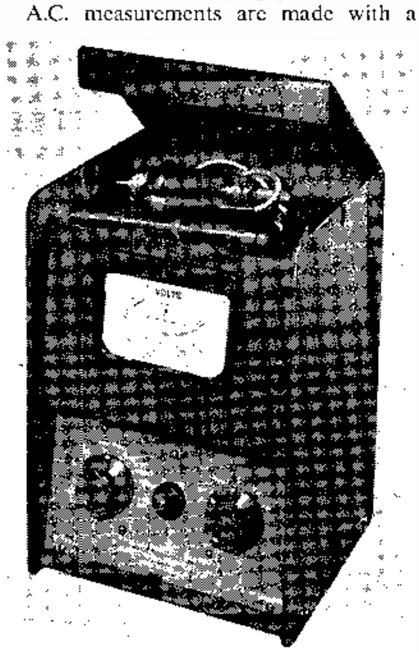
(Illustrated below)

Marconi Instruments Ltd, St. Albans, Hertfordshire

THE a.c. measurement range of the TF1300 is 0·1 to 100V, 20c/s to 300Mc/s. It measures d.c. up to 300V and resistance up to 5MΩ. The indicating meter is direct-reading on a.c., d.c., and ohms ranges, and no correction factors are necessary. Zero stability is of a high order with respect to both time and mains variation, and only one zero setting is required for all a.c. or d.c ranges. Both a.c. and d.c. inputs are isolated from chassis.

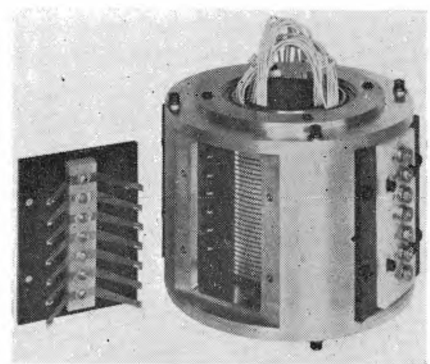
A radically new concept in mechanical simplicity has been introduced in this instrument, enabling costs to be cut while maintaining a high standard of performance. There is no conventional chassis, and all components are mounted on the rear of the sloping front panel, which hinges upwards for ease of servicing. Valve-base wiring and preset controls for calibration are concealed behind a quick-release meter surround. Operational presets are accessible in a case-top recess which is also used for lead stowage. The recess lid hinges back to form a support that allows the instrument to be used, if required, in a 45° sloping position.

A.C. measurements are made with a



light-weight, cylindrical probe allowing direct contact with the circuit under test. Small, compact, and fitted with a non-slip p.v.c. sheath and safety guard, the probe is ideal for use in confined spaces. The rectified output from the probe is applied to one arm of a valve bridge comprising two double triodes; a second arm of the bridge is coupled to a splash-current balancing network including a potential divided across the h.t. supply and a 3V stabilizer element. The indicating meter is automatically protected from overload.

For d.c. measurement, the input is applied directly to the bridge via the d.c. leads, and the diode and splash-current network are automatically disconnected. For resistance measurement, the unknown resistor is connected, by means of the d.c. leads, in series with a standard resistor and the 3V stabilized supply that is used for splash-current balancing in a.c. measurement. The meter responds to the voltage developed across the unknown resistor.



PRECISION SLIP-RINGS AND COMMUTATORS

(Illustrated above)

I.D.M. Electronics Ltd, Whitley Kiln, Basingstoke Road, Reading, Berkshire

I.D.M. Electronics have introduced a specialist service for the manufacture, to customers' specifications, of precision slip ring assemblies and commutators by the Electro Tec Process.

This method of production, although new to Britain, is well established in the United States of America.

By this process a plastic blank is moulded around lead wires. Accurate machining reduces this blank to the proper preliminary shape. Gold or silver is deposited into the grooves by electro plating, thus making conducting rings an integral part of the unit. Final machining ensures perfect concentricity and true dimensional accuracy. Thus errors due to accumulated tolerances, normally associated with conventional farbicated assemblies, are eliminated. The conducting rings are finally polished to a finish of four micro inches or better. The special surface finish, thus obtained, minimizes friction, wear and noise.

This process is very versatile; diameters of slip rings may vary from 0·035in to 36·0in, either cylindrical or flat. Cross-section of the rings may range from 0·005in on small diameter assemblies to 0·080in or more, on large power carrying assemblies. The extreme miniaturization made possible by the Electro Tec technique results, not only in saving of weight, and space, but in minimum frictional torque and very low noise levels. This process also eliminates expensive tooling charges, and thus frequently results in lower costs than are possible with other methods of manufacture.

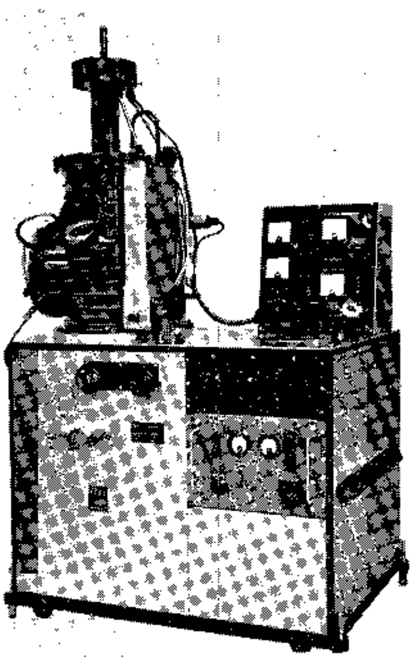
Insulation resistance, between circuits is greater than 2 000 000MΩ, and high temperature ratings up to 260°C are attainable.

SILICON CRYSTAL PULLING UNIT

(Illustrated below)

Edwards High Vacuum Ltd, Manor Royal, Crawley, Sussex

THE 'Speedivac' silicon crystal pulling unit is basically a flexible unit and where necessary it can be modified



by the user with the minimum effort and expense to suit individual requirements peculiar to the art of crystal pulling. It is eminently suitable for both development work and full-scale production.

The pumping system comprises a 6in oil diffusion pump capable of a final vacuum of better than 5×10^{-6}mm Hg, and having a baffled speed of more than 300 litres per second. A 9LB6 baffle and isolation valve is mounted above the diffusion pump which is backed by a 15ft^3/min rotary pump. The rotary pump outfit is mounted separately on the floor and connected to the diffusion pump by flexible connexion to minimize vibration. A small 1in oil diffusion pump is also fitted to the equipment for evacuating the 'screen tube' surrounding the crystal pulling rods, thus ensuring that no oil or grease contaminates the growing chamber. The whole pumping system is complete with isolation and control valves and protective devices. Vacuum measurement in the plant is provided by Pirani and Penning type vacuum gauges which are mounted on the top of the cabinet housing in a complete instrument panel. The same panel also contains the instrumentation for the other electrical equipment which is used on the plant.

The remainder of the equipment comprises a crystal growing chamber with hoist and rotating mechanism for raising and rotating the crystal during its growth. The growing chamber is approximately 2ft high and its provided with pumping and loading ports and viewing windows. The hoist and rotating mechanism is contained in a flanged metal casing approximately 16in high and this is mounted directly on to the cover of the main vacuum chamber. The crystal pulling rod is arranged for variable rates of rotation and pull. The entire apparatus is housed in a stove enamelled metal cabinet, the dimensions of which vary depending on whether the rotary pump is mounted externally or internally. The cabinet is 3ft 6in high by 3ft 8in wide by 3ft deep and the overall height of the plant to the top of the crystal mechanism is almost 7ft.

GENERAL PURPOSE OSCILLOSCOPE

E.M.I. Electronics Ltd, Hayes, Middlesex.

In addition to meeting the needs of general laboratory electronic work, this new instrument—known as the type WM7—can also be used for specialized work on radar, television, computers and millimicrosecond oscillography, normally catered for only by the more specialized type of oscilloscope.

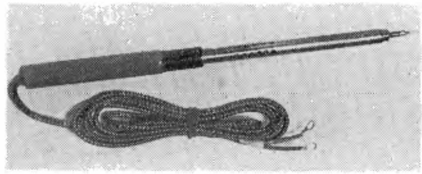
The instrument's Y amplifier bandwidth extends from d.c. to 50Mc/s (−3dB), when a plug-in wide band amplifier having a vertical sensitivity of 100mV/cm is used. An alternative dual-channel beam switch plug-in unit gives a bandwidth of d.c. to 25Mc/s, with a vertical sensitivity of 50mV/cm. Details of other plug-in units, which are an important feature of the equipment, will be available shortly. These include differential input and high-gain wide bandwidth units.

Among the many new facilities incorporated in the WM7 is a push-button controlled cathode-ray tube beam locating system which operates so that if the spot is deflected off the screen depression of the push-button confines the beam within a limited area on the screen, so that the normal shift controls can be used to centralize the display.

The instrument also employs a unique cathode-ray tube display feature whereby parallax errors are automatically eliminated by the use of a specially illuminated reflecting filter and graticule. A new type of 5in high sensitivity cathode-ray tube is used with an accelerating potential of 10kV.

Two time-bases are built into the unit, giving normal and delaying sweeps. The normal sweep speed is continuously variable over a range from 12·5m/μsec to 0·5sec/cm. Time and voltage measurements to an accuracy of 2 per cent are available under all conditions of expansion.

High accuracy measurements can be made rapidly either by direct readings from dials, or by the volts/cm, time/cm methods in conjunction with the special parallax corrected graticule. All supplies to the instrument are fully stabilized and d.c. coupling is used between all main functional circuits.



LIGHTWEIGHT SOLDERING IRON

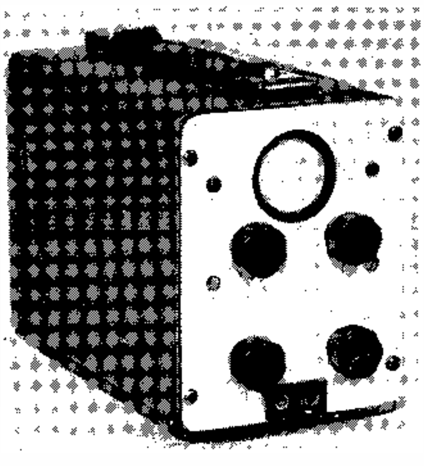
(Illustrated above)

Browning's Electric Co. Ltd, Boleyn Castle, Green Street, Upton Park, London, E.13

DESIGNED for instrument and electronic work the Browleco pencil iron weighs only 1¼oz, has a barrel diameter 5/16in and an overall length of 8in.

The element is wound on to an insulated copper alloy bit holder designed to act as a heat storer, into which various size copper bits can be fitted. This design gives excellent conductivity to the working tip and takes approximately 2min to heat up.

The standard voltage is 6 or 12V but the elements can be wound specially up to 50V. The irons are suitable for working from a battery or a double wound transformer.



MINIATURE OSCILLOSCOPE

(Illustrated above)

Bourne Engineering Co. Ltd, Abbey Mill, St. Albans, Hertfordshire

THE Bourne Engineering Uniscope is available in two forms—firstly as an independent oscilloscope for the laboratory, workshop, and maintenance engineer, and secondly as a versatile monitor for building into larger electronic equipments such as computers, racks, production test gear and so on.

The Y amplifier sensitivity is up to 0·1V/cm from 5c/s to 350kc/s and the time-base is continuously variable from 5c/s to 50kc/s.