ELECTRONIC ENGINEERING

VOL. 29 No. 357

NOVEMBER 1957

Commentary

THE electronic component industry of this country has expanded at an astonishing rate over the last decade and, since the end of the war, it has been doubling in size every five years. Much of this increase is due to the rapid growth of television and for some time past the entertainment equipment manufacturers have been the largest customers of the component industry.

But side by side with this supply has been a steadily increasing demand for components for inclusion in the many electronic devices which form part of the present day military machine and the position has now been reached where the Ministry of Supply, which acts as the purchasing agent for the three Armed Services, now absorbs by far the greater portion of the output of the component industry.

The extended use of electronic devices by the Services has given rise to a number of problems other than the mere quantitative one of expanding the manufacturing facilities of the component industry for, experience gained during the war, particularly in the Far East, has shown that few, if any, of the 'standard' components were suitable for use in Service equipment and the 18th century maxim, "A little neglect may breed mischief . . . for want of a nail the shoe was lost; for want of a shoe the horse was lost; and for want of a horse the rider was lost", was very forcibly driven home.

The introduction of such post-war developments as guided missiles and supersonic aircraft has accelerated the search for reliable components capable of operating over a wide range of adverse climatic conditions and severe mechanical shocks in Service equipment. According to the conditions encountered, ambient temperatures from $-55°C$ to $150°C$ are now likely to be met, with high relative humidities and possibly low air densities—down to pressures of 70mm or less. The conditions can be combined with forces due to vibrations up to 2kc/s, severe bumping up to $40g$ and angular accelerations up to $100g$.

Simultaneously the introduction of transistors, printed wiring and automatic assembly techniques has raised a number of fresh problems in component design. The use of the transistor demands components of comparable size and of lower voltage rating, while printed circuits and automatic assembly techniques require an entirely fresh approach to such aspects of component design as dimensions, methods of fixing and connecting.

Thus, for some time past, there has been a need to review component design on a broad basis and to discuss the various problems which present day requirements have created. Most timely, therefore, was the recent exhibition and symposium on electronic components organized by the Royal Radar Establishment of the Ministry of Supply and held at the Nelson Hall at R.R.E. Malvern from September 24-26—the first of its kind to be held in the United Kingdom.

The standard of the papers presented—some thirty in all—and the discussions which followed showed how great was the need of such a symposium and, at its conclusion, there was the general feeling that after a due period of time it should be repeated.

The highest praise is due to the Royal Radar Establishment for their efforts in launching the Malvern meeting which ran smoothly and efficiently, but we are sure they would be the first to admit that they are not the most suitable authority. They do not possess the organizational facilities on the required scale and they are hampered by security restrictions, nor does Malvern itself seem a suitable centre.

One suggestion which has much to recommend it is that the Radio and Electronic Component Manufacturers' Federation should take over the responsibility of holding future events of this kind. They have the requisite organizational facilities and they could, with advantage, arrange that future meetings run concurrently with the Components Exhibition which the R.E.C.M.F. holds each year in London.

* * *

Few scientific achievements have burst on the world with such dramatic suddenness as has the successful launching of the earth satellite by the Russians at the beginning of last month and much speculation has arisen in consequence.

Little is yet known regarding the satellite itself or the method of launching, but it would appear that it is larger and its orbit higher than anything planned by the Western World.

It is the first time that the Russians have been the first in the field with such a major technical achievement and it establishes, without a shadow of doubt, the remarkably high standard their scientists and engineers have attained in the post-war years.

WIND TUNNEL INSTRUMENTATION

By A. Tiffany*, B.Sc., A.M.I.E.E.

The design and operation of modern wind tunnels centres increasingly upon the instrumentation techniques available for collecting and processing the experimental data. Interest is growing in the application of digital methods in many fields of testing and control and this article outlines the development of a data handling system matched to the needs of a particular problem. The success of relatively simple and inexpensive electromechanical digital read-outs indicates increasing application of data handling techniques in the near future.

As modern wind tunnels have grown in size and complexity an urgent demand has arisen for expeditious methods of recording and reducing the flood of data that their operation produces. The advent of commercial digital computers coincided with this demand and both in this country and the United States computers have been installed by aircraft manufacturers and testing establishments to deal with this and many other numerical analysis problems arising as the tempo of research and development has quickened.

Schemes for tunnel data handling have varied widely to meet individual organizational needs and have ranged from the 'on-line' connexion of a single purpose computer, in which observations are fed directly into the machine as the testing proceeds and corrected results fed back continuously and virtually instantaneously[1], to arrangements in which tunnel results form only a small part of the work of a large general purpose computer, which may not even be near the tunnel site, corrected results being returned to the test crew after a delay of anything from an hour to a week.

The scheme described herein is a not untypical compromise in which punched cards are prepared, in a form suitable for direct entry into a computer, as the test progresses, enabling important conditions to be awarded priority in the event of computing delays so that additional or interpolative tests may be planned while the test set-up is still available.

All these schemes add to the already onerous task of the instrumentation engineers in providing stable and precise measurement of model behaviour a requirement for digital methods of data handling matching the precision of the primary measurements. The present article deals with one of the many schemes that have been devised to meet this new requirement: it has been in successful operation for some 18 months. Development is proceeding at a great many establishments with data handling installations for many other fields of testing, control and analysis and the installations currently operating can be regarded as interim measures or pilots for what is likely to be a considerable development in the next few years.

The present exposition of the problems and evolution of a comparatively simple installation may serve as an introduction to a field as yet inadequately documented.

An Outline of Instrumentation Requirements for a Wind Tunnel

The principal quantities of interest in the simplest type of wind tunnel testing are the three resultant forces and three resultant moments acting at the centre of gravity of the test model which is regarded as a rigid body having six degrees of freedom on the strain gauge 'balance' which connects the test model to the supporting member. (See Fig. 1.) The balance does a preliminary separation of the component forces and moments by providing members each primarily stressed by a particular component, e.g., normal force, rolling moment, while possessing a degree of rejection to the remaining five components.

Since such balances are necessarily compact, structural interactions are inevitable. Moreover the rejection of interaction stress signals often depends on the symmetry of placement and sensitivity of strain gauges so that imperfect positioning of gauge elements and variations of strain sensitivity between gauges, whether of internal or external origin, yield effective interaction terms in the calibration. In very compact balances evidence exists of second order interaction terms in addition to the linear set[2]. An interaction matrix is evaluated by dead weight loading of the balance before use in a model and the inversion of this matrix together with the conversion of the results from model axes to wind axes constitute the bulk of the data reduction programme carried out on each set of six components measured at a test condition. An accurate extraction of the corrected forces and moments necessitates strain measurements to about 0·1 per cent accuracy of a full scale value in the region of 200 microstrain. With conventional strain gauge equipment this requires a signal of about 2·5μV to be resolved. Digital methods apart, self-balancing equipment is desirable for this duty and servo-operated instruments of the Wheatstone bridge

* *Aircraft Research Association Ltd.*

The above photograph shows an aerial view of the Aircraft Research Association wind tunnel. Photograph reproduced by courtesy of 'Flight'.

variety have been developed to meet the required resolution and commercial potentiometric recorders and indicators have been successfully adapted for the purpose.

The remaining basic instrumentation is concerned with the model attitude in the windstream and the windstream speed. The model attitude is defined by a combination of two angles, commonly the root angle of the model support cantilever and the angle of roll of the support system about the longitudinal axis. These angles require correction for support deflexions in order to derive the true angles of attack, sideslip and roll. Synchro links are often used to obtain the basic angles to about 0·1° although direct counting methods, e.g., counting of Moire fringe patterns generated from circular diffraction gratings, have been attempted where space is at a premium.

The Mach number in the test or 'working" section is computed from the ratio of pressures in the working section and in the total head region ahead of the accelerating contraction or nozzle. These reference pressures are almost universally measured (in this country) by self-balancing capsule manometers of the weigh beam type which provide output and balancing actuation via step transmitters[3]. An accuracy of 0·01in Hg can be maintained over the range 0 to 72in Hg.

Some Special Problems of Measurement

The strain gauging precisions called for impose exacting requirements on both the recording instruments and the gauges themselves. Available gauges are far from satisfactory, in spite of widespread and prolonged development by both users and manufacturers. In particular the behaviour with temperature is unpredictable and in detail almost indeterminate. The cost of cooling plant for large tunnels constitutes an appreciable proportion of the total plant expenditure and attempts at exact control of air temperature would be prohibitive.

Gauges comprising a constantan wire grid embedded in a bakelite matrix have won popularity because of the stability of their strain sensitivity or 'gauge factor' since their first introduction as precision elements in load cell devices for automatic weighing. In order to minimize cross sensitivities wires of the order of 1 mil diameter have to be used so as to keep the winding angles small. The properties of fine wires and of the bonding of them into the bakelite matrix are ill defined and both these and the physics of the phenolic resin bond used to attach them to the metrical element combine to provide some very frustrating habits.

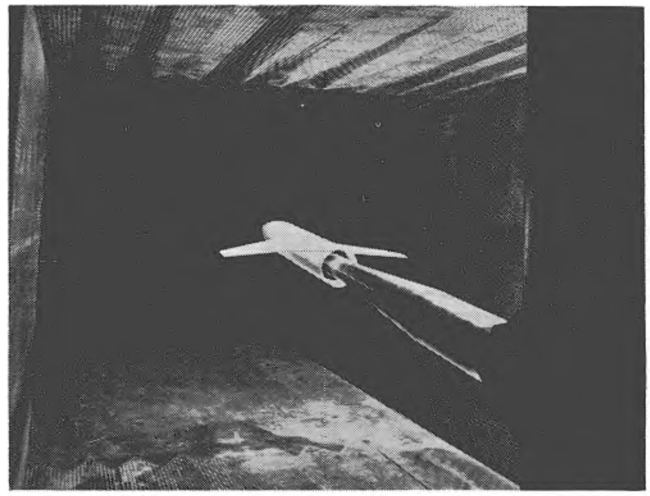


Fig. 1. A model mounted ready for measurements in the wind tunnel

A varying ambient temperature yields resistance changes in individual gauges amounting to many times the strain induced resistance changes to be measured. Combination of pairs of gauges disposed in adjacent arms of a bridge leaves drifts due to the disparity of ostensibly similar gauges which normally amount to 1 to 3 per cent of a 200 microstrain range (6 000lb/in^2 in steel), which is often accepted in other applications as the limit of what may be expected from strain gauges. Nevertheless it is found that by careful preselection of gauges into pairs prior to adhesion and *ad hoc* compensation of each complete bridge (normally four active gauges are used to form a complete bridge) zero drifts may be limited to 1 or 2 parts in 1 000 over a 30°C rise. Both the selection and the correction must at present be categorized as arts rather than sciences

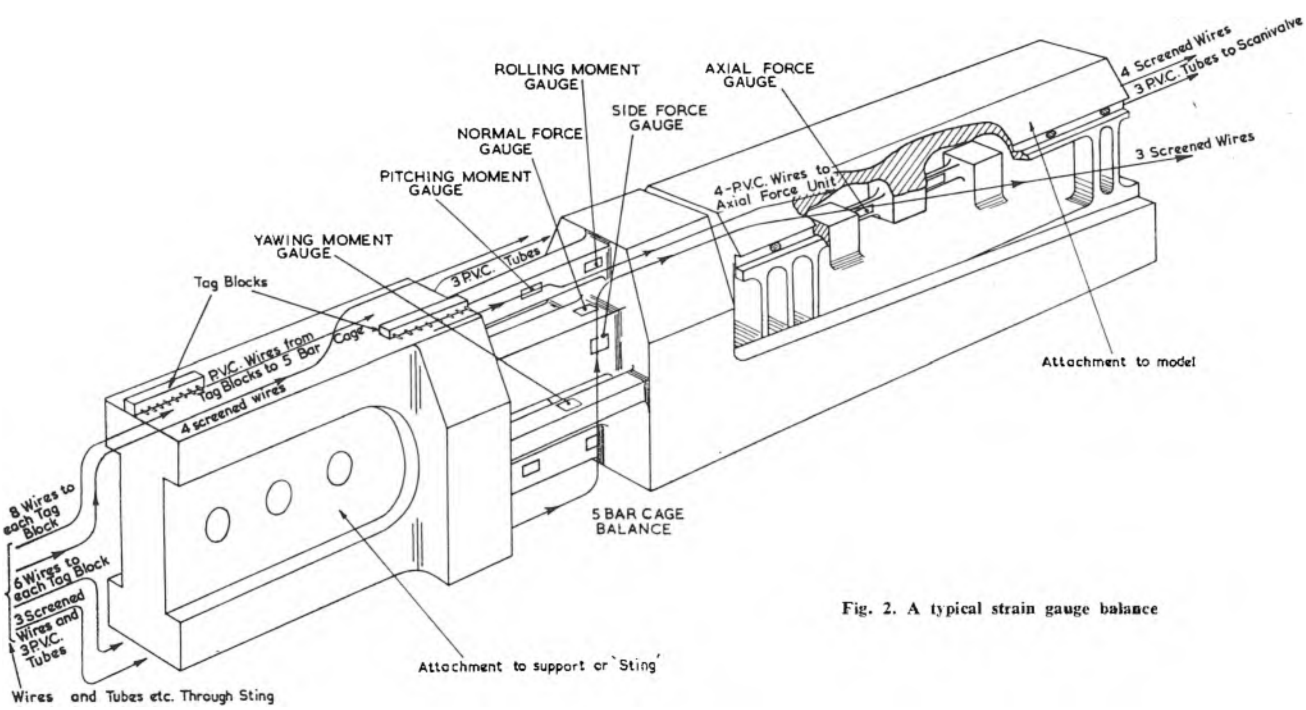


Fig. 2. A typical strain gauge balance

and little has in consequence been published. It is hoped to treat this and some of the associated circuit problems arising from attempts at precision more fully in a future article. MacFarland and Dimeff[4] have given a very full account of the factors limiting precision in strain gauge work. Fig. 2 shows a typical strain gauge balance.

Servo Operated Instruments for Strain Measurement

The self heating of strain gauges limits the permissible bridge excitation to about 6V. Assuming a gauge factor, K, of 2 a four arm bridge will give an unbalance signal of $dV = K.V..dl/l = 2 \times 6 \times 200 \times 10^{-6}$, or 2·4mV for 200 microstrain and a resolution of 1/1 000 of this is required. Alternatively a resistance change of $0{\cdot}4 \times 10^{-6}$ parts in a 120Ω gauge must be detected, i.e., 0·48μΩ. Opinions differ widely as to the relative merits of d.c. and a.c. bridge methods for these measurements but both have been successfully developed. Fig. 3 indicates the main components of one commercial instrument, utilizing a.c. bridge excitation in a manner ensuring insensitivity to the precise value of bridge voltage.

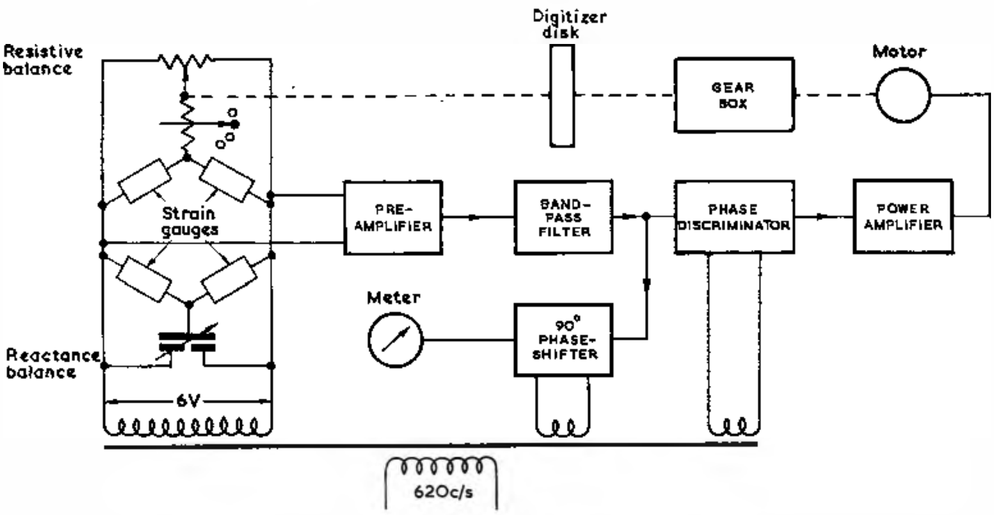


Fig. 3. Automatic strain gauge bridge

Such instruments suffer from the severe disadvantage of requiring reactive as well as resistive balancing, necessitating carefully balanced and mechanically stable leads to and from the bridge which may be 100ft or more from the instrument. Improper reactance balancing causes dead zones and cross couplings between channels which may be of the proportions of the mechanical interactions of a balance. Such instruments are also very unreliable in the presence of oscillatory components in the strain signal.

These objections are largely overcome if d.c. instruments are used and commercial potentiometer chart recorders offer a robust and sensitive instrument at modest cost. The dangers of thermo-electric effects are very real—constantan gauge wires yield an e.m.f. against copper of the order of 40μV/°C—although suitable interwiring between gauges can balance thermo-junctions against one another except in the presence of severe local temperature gradients. Successful schemes have been introduced in which the total thermo-electric output is isolated, by interrupting the gauge supply, and bucked out by a small series potential, automatically nulling, at intervals during operation[5]. Although the accuracy of chart records from commercial instruments is limited by many factors to some 0·3 per cent the precision of the angular travel of the main slidewire rotor is usually as good as 0·1 per cent providing the full scale range is not less than 2mV. Potentiometric methods are not apt for strain gauge work since the gauge supply has to be stabilized to better than the measuring accuracy. However substitution of a fraction of the gauge supply voltage for the usual standard cell will then give a measurement of fractional resistance change direct.

The chart recorder also lends itself to adaptations of the chart drive arrangements to produce plots of test data against any chosen variable. Repeater motor and synchro links readily provide plotting against model attitude.

Multiple Pressure Measurements

The determination of pressure contours and distributions form a considerable part of the work in wind tunnel testing. Such experiments, generally termed 'pressure plotting', may require as many as 200 pressure observations to be performed at a single condition to accuracies of about 0·01in Hg. Much of this work is at present carried out by photographing manometer banks, but the laboriousness of retrieving the data from photographic records even with the aid of digital film readers, contrasts sharply with the facility with which force data is processed. Hence an urgent need exists for more direct methods of transferring pressure data in quantity into a computer; however, a requirement for rapid read-out is implicit if the expenditure of costly tunnel running time is not to exceed the costs of manual or semi-manual processing.

This aspect of data handling has received comparatively little attention as yet. Some current schemes involving high speed storage are costly—some £50 per manometer tube is estimated for two such schemes—and moreover carry with them all the disadvantages of manometer practice. For wind tunnel applications it is desirable that at least a proportion of the pressure instrumentation is capable of being housed in the model itself. Preliminary work at the author's establishment with rapid pressure sampling switches containing an integral flush diaphragm transducer to minimize the travelling volume has been extremely encouraging, results being obtained on high speed chart recorders (0·25sec balancing time) at the rate of about four per second. These recorders do not lend themselves to ready digitizing and it is thought that more promise is held in the application of digital voltmeters, now becoming available in several forms, to reading the output of the switch mounted transducers and the preparation of punched paper tape records, with possible simultaneous tabulation or plotting from the digital data. A 48-way pressure sampling switch capable of travelling at ten samples per second when mounted in a model is at present in use.

The Application of Digital Techniques in the Wind Tunnel

No single advantage of the many obviously accruing from the introduction of automatic techniques of data collecting, handling and reduction may be cited as engendering the current vogue. Apart from the advantage of having corrected results rapidly available during a trial the very heavy capital and running costs of large scale tunnels especially of the continuous flow variety make a high rate of testing prudent if not imperative. Driving powers of 10 000 to 60 000 h.p. may be required and maximum demand penalties and off-peak loading concessions make speed and flexibility of the testing programme well worth 'designing-in' from the outset. The installation described below forms an integral part of a facility designed round the availability of a digital computer for data handling. Tunnel access, model rigging and mounting arrangements were also designed to allow rapid and convenient interchange of test items. The tunnel was designed to serve as a common facility for a group of 14 aircraft and engine companies, to meet the special difficulties of testing in the speed range around Mach 1.

assimilable display of the trends of the uncorrected data; this monitoring facility needed only to be approximate since the corrections could not readily be applied with worthwhile precision. It was important that the time during which the instruments were locked up or 'frozen' for read-out be brief, in order that the next tunnel condition could be set up expeditiously while data handling proceeded—some short term memory had to be provided. Lastly it was important that the tunnel was never waiting for the instrumentation with operating costs in the hundreds of pounds an hour region going to waste. By the same token a heavy onus for reliability was put on the proposed installation since 30MW loads were not lightly to be shed and restored at frequent intervals. The system adopted is shown in outline in Fig. 4.

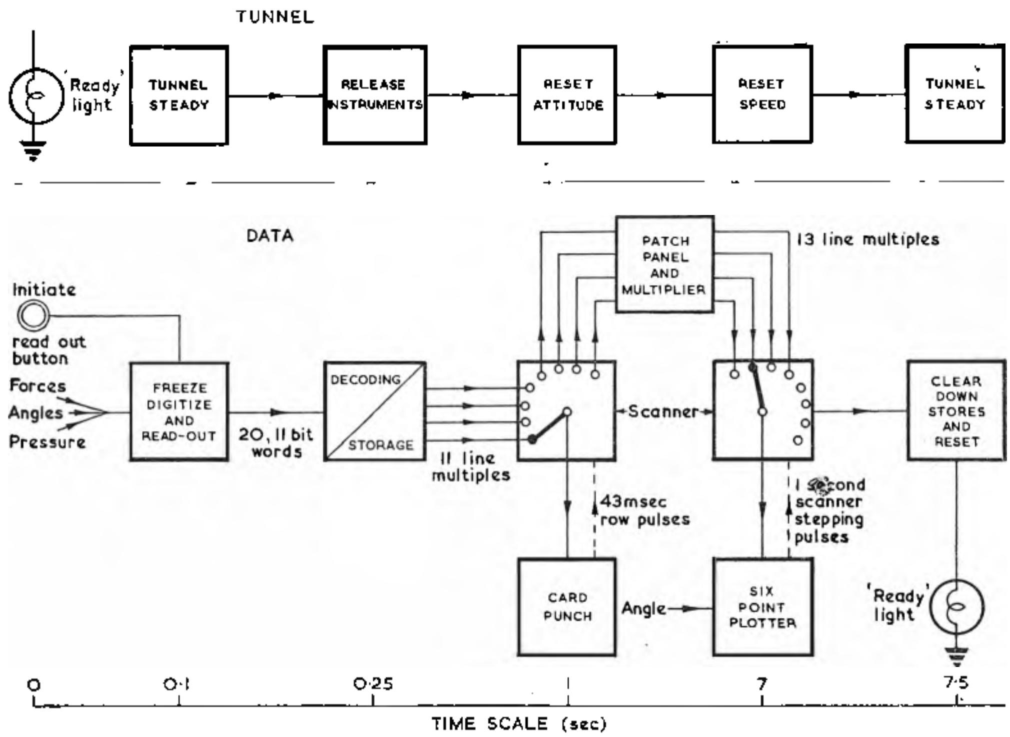


Fig. 4. Outline of digital data handling system for 9ft × 8ft transonic wind tunnel

Design of a Digital Data Accumulating System

The use of self-balancing servo instruments, synchro follow-ups and weigh beam manometers provide a common form of representation of the analogue quantities being measured, namely precise shaft rotations. By the use of suitable gearing all could be reduced to a single-turn-full-scale basis and this lent itself to the application of shaft position encoders or 'digitizers'. These provide, on a suitable number of lines, patterns of conduction and non-conduction uniquely representing any shaft angle in the range 0 to 360° quantized to a resolution matching the required accuracy. The pattern used in this instance provided 1 200 unambiguous coded outputs to the circle and this represents about the limit of what can be achieved on single turn encoders.

The choice of digitizer type and coding was influenced primarily by the use to which the data was to be put subsequent to read-out and by the time scale of the tunnel stability characteristics. The primary object was to prepare punched cards suitable for direct entry into an existing computer; conveniently but not essentially packaging the data from any one test point on a single card. A second objective was to provide the test conductor with a readily

Punch-Out Equipment

It will be seen that the longest operation is the monitor plotting which is carried out on a multi-point sequential plotting table. The plotting operation can, however, be omitted entirely and the complete punch-out cycle then occupies just under one second, enabling comparatively rapid sampling to be done under suitable conditions. In this form the installation has been repeated unaltered for a supersonic tunnel facility now building, being combined in this case with individual chart drive to a number of recorders to provide a parallel monitoring scheme, since the resetting time in the supersonic case is slight.

Relay storage of the digital data enabled the instruments to be released as soon as the stores have been loaded; parallel read-out into separate stores for each channel was practicable and cheap using relays. Coded commutator type shaft position digitizers were employed in which the conduction patterns are provided by light brushes passing over finely divided annular tracks on an engraved copper disk[6]. Such digitizers can pass, but may not commutate, relay currents so that the read-out consisted of passing an interrogating pulse while locking the servo mechanisms. Binary coding and row-by-row card punching sped the punch-out and enabled more information to be packed on to the card: the one card per read-out arrangement was achieved. The coding on the digitizer disks was specially formed to avoid ambiguities—but is not error detecting—so that storage was preceded by code conversion and results stored in simple 11 bit binary form. The code conversion is the simplest of all the 'unambiguous-to-plain-language' decoding logics which adds to the attractions of the binary form of coding. Decoding outside the computer simplifies the subsequent arithmetical operations in the plotting equipment. A typical data card is shown in Fig. 5.

Plotting Equipment

Selected variables can be patch-boarded across to the plotter and plotted against model attitude, generated by a linear potentiometer in the synchro follow ups. The plotter information is formed by summing the digits in a relay controlled divider network, an elaboration justified in practice as a check on the correct functioning of the digital equipment even though only gross errors are immediately observable. The adoption of binary coding allowed the problem of multi-ranged instruments to be accommodated conveniently by arranging sensitivities in the ratio 1:2:4, outputs from a single scale shaft position digitizer could then be allocated correct significance merely by shifting the patterns to left or right. This was done prior to plotting by means of a cross-wired multi-pole switch motorized to hunt to positions marked by individual range switches on each channel, thus avoiding discontinuities in plotted results.

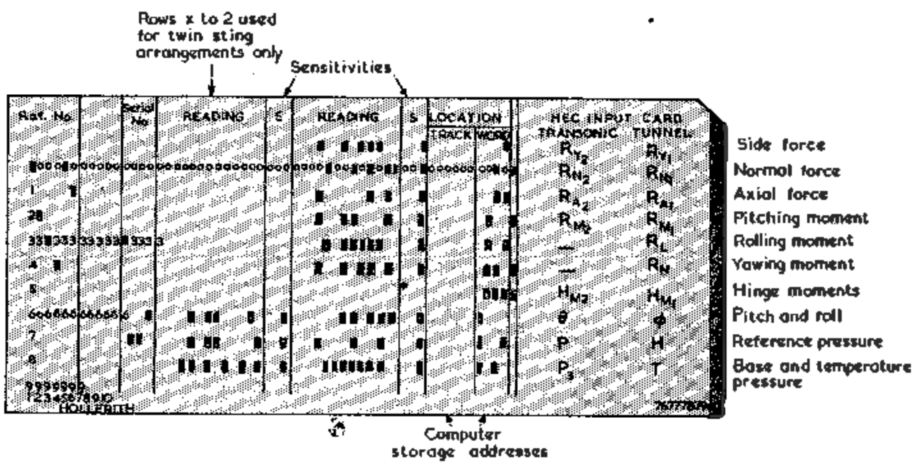


Fig. 5. A punched card for computer input

Operating Experience

The majority of the equipment of the final scheme was installed without opportunity for extended life testing. For this reason the design was aimed at absolute reliability rather than quick replacement on failure. The resulting installation is flexible in that various combinations of measuring equipment can be coupled into the system and sequences on to card and plotter with facility, but cumbersome in that such hooks-ups take some hours to cable through and check. Nevertheless, the experience of the first year and a half of operation confirms that the positive merits of reliability far outweigh the arguable demerits of relatively slow minor servicing. Moreover the usual pattern of testing in four hour runs, with intervening shut down for adjustment to model configuration of duration up to an hour, make elaborate packaging a needless expense. In the face of minor faults confined to a single channel faulty apparatus tends to be shed in order to make profitable use of the remaining tunnel time.

To date the tunnel has had to shut down due to total equipment failure on only one occasion. The use of well tried Post Office relays in the decoder and store equipment and derated heavy duty uniselectors in the scanning unit is felt to account for the very satisfactory reliability records of these items—three relay coil failures and two scanner timing failures due to running out of adjustment.

Apart from one catastrophic coil failure the commercial card punch equipment has an equally satisfactory record although minor difficulties arise due to faulty card handling and storage. Strict adherence to the manufacturers recommendations can minimize this risk.

The digitizers were at first thought to be the most likely source of errors but in practice have proved most reliable. Two brush failures have occurred in 18 months' continuous use and two engraved disks have been replaced as showing signs of wear but not failed.

Little routine or preventive maintenance is carried out but daily calibration checks are carried out using punched card rather than visual observations. The installation gives greatest reliability when used intensively and disturbed as little as possible. As far as practicable a basic set of equipment is kept permanently in place in the control room and variants added peripherally around it.

One practical disadvantage of the system installed is that all digital checking has to be done in binary and display lamp sets in the code require concentrated attention. Where the analogue measuring instrument has only an approximately divided scale the problem of absolute calibration checking becomes aggravated by the necessity to read observations off the digitizers. The relative simplicity of the data handling circuits in the writer's view more than compensates for this.

Conclusions

A fairly simple type of data handling system has been described in the context of the measurement problems peculiar to wind tunnel practice. The arguments leading to a system matched to the cyclic testing pattern are propounded in order to illustrate the principles upon which other such schemes might be based. Current schemes while satisfactorily accurate and reliable are felt to be precursors of more elaborate and generally applicable equipments of the near future.

Acknowledgments

The author is grateful to the Chief Executive of the Aircraft Research Association for encouragement to prepare and publish this material.

REFERENCES

1. Taylor, D. F. On-line Automatic Data Reduction at the Arnould Engineering Development Centre. AEDC Tech. Note 56-13.
2. Hansen, R. N. Mechanical Design and Fabrication of Strain Gauge Balances. AGARD Report 9.
3. Midwood, G. F., Hayward, R. W. An Automatic Self-balancing Capsule Manometer. R.A.F. Tech. Note AERO 2382.
4. McFarland, K. H., Dimeff, J. Problems Involved in Precision Measurements with Resistance Strain Gauges. AGARD Report 12.
5. Bassiere, M. Mesure et Enregistrement des Mesures sur les Balances Aerodynamiques a Jauges a Fil Resistant. AGARD Report 14 or R.A.F. Trans 666.
6. Petherick, E. J. Some Recently Developed Digital Encoding Devices. R.A.E. Tech. Note MS 21.

An Automatic Fog Warning Instrument

By J. W. Nichols*, B.Sc., A.M.I.E.E., and M. H. Westbrook†, B.Sc., A.M.I.E.E.

In this article an instrument which will give automatic warning of a decrease in visibility, through fog, is described. The instrument works on the back scattered light from a projected beam. The back-scatter is measured by a photocell, amplifier and RC integrator. The electronic features of the design are dealt with in detail.

FROM the time the first attempts were made to give warning to ships at sea of dangerous situations it has been necessary to make visual observations of visibility to determine when the warning is needed. The instrument described in this article has been developed with a view to avoiding the need to keep a visual look out for a reduction in visibility and to provide a means of starting the warning signal without manual intervention.

It is a common experience that in fog or mist a beam of light is scattered by the water droplets forming the mist and it is this principle which is utilized in this instrument. The design is to a large degree optical but, in this article, only the electronic features of the design will be dealt with in detail, the minimum optical and mechanical information being included.

Method of Operation

The projector unit is shown in Fig. 1 and a block schematic diagram showing the arrangement of the various parts forming the instrument is shown in Fig. 2. A projector unit consisting of a 20-in searchlight with a compact source mercury vapour lamp at its focus provides a narrow beam of modulated light. Above the searchlight with its axis parallel to the axis of the beam is mounted a telescope formed by a single lens. The angle of convergence of this telescope is controlled by the size of an aperture placed at the focus of the lens, the light passing the aperture falling on to a photo-electric cell. The electrical output of this photocell is amplified locally and then fed via further stages of amplification to a detector and *RC* integrator. When the output across the capacitor, which is a measure of the light from the searchlight scattered back into the telescope, reaches a pre-determined level a relay circuit is energized to provide the required alarm.

To provide for fog in a number of different directions the searchlight and telescope combination are mounted so that they may be rotated in azimuth, the use of slip rings being avoided by restricting the rotation to one single turn.

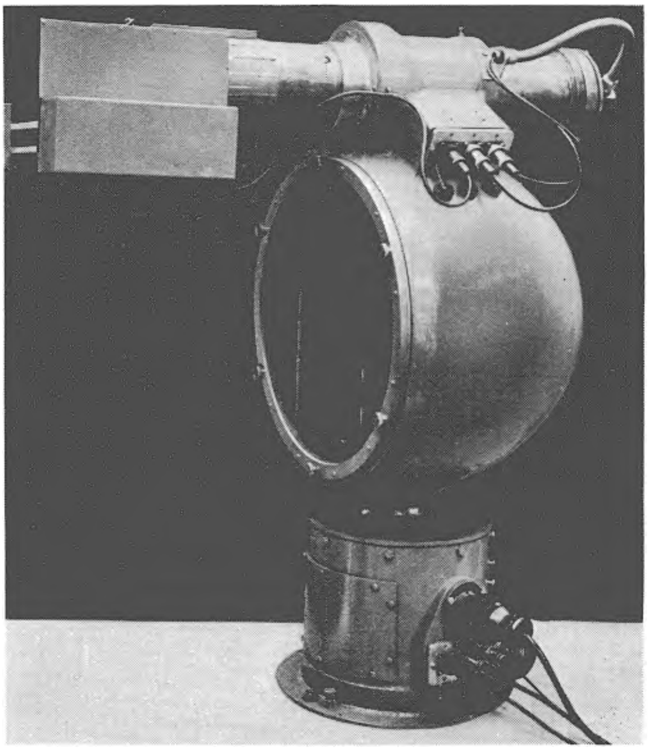


Fig. 1. The projector unit

Design Considerations

To obtain a beam of the highest intensity it is necessary to make use of a light source of very small dimensions. It is also necessary that the lamp chosen may be modulated electrically. These two facts have dictated the use of a high pressure mercury arc lamp. The modulation in such a lamp takes place at 100c/s, when connected to the 50c/s supply mains. Having chosen the lamp the choice of photocell is limited to one whose response is mainly in that part of the optical spectrum produced by the lamp.

The problem is now that of detecting the small changes in photocell current produced by the back scattered light. This is, of course, a typical example of the problem of detecting a signal in noise. The noise in this case arises from two causes, firstly the Johnson noise in the photocell load resistor and secondly, the shot noise in the photocell itself. The magnitude of the noise due to this second cause is a function of the unmodulated ambient light falling on the cell and correspondingly varies from day to night. However, a limit to the amount of ambient light permissible on the photocell cathode is set by the limit of the linear part of the photocell light/current characteristic. This factor is used to determine the aperture of the lens used.

In the design of the amplifier and detector stages precautions are taken to ensure that the signal-to-noise ratio is not worsened and the final detector system limits the noise bandwidth to 0·0625c/s. This, of course, implies that no alarm of fog will be given until 16sec after its incidence. For the application for which the instrument was designed this is no disadvantage. This limitation has been used in determining the speed of sweep of the system so that an arc of about 20° is covered in the 16sec. Unless ships are extremely close to the station it can easily be arranged that reflections from these will not give rise to false signals.

Circuit Details

A schematic diagram of the amplifiers and detector circuit is shown in Fig. 3. This diagram has been simplified

* *Research and Development Section, Trinity House.*

† *Formerly with Trinity House, now with the Rank Organization.*

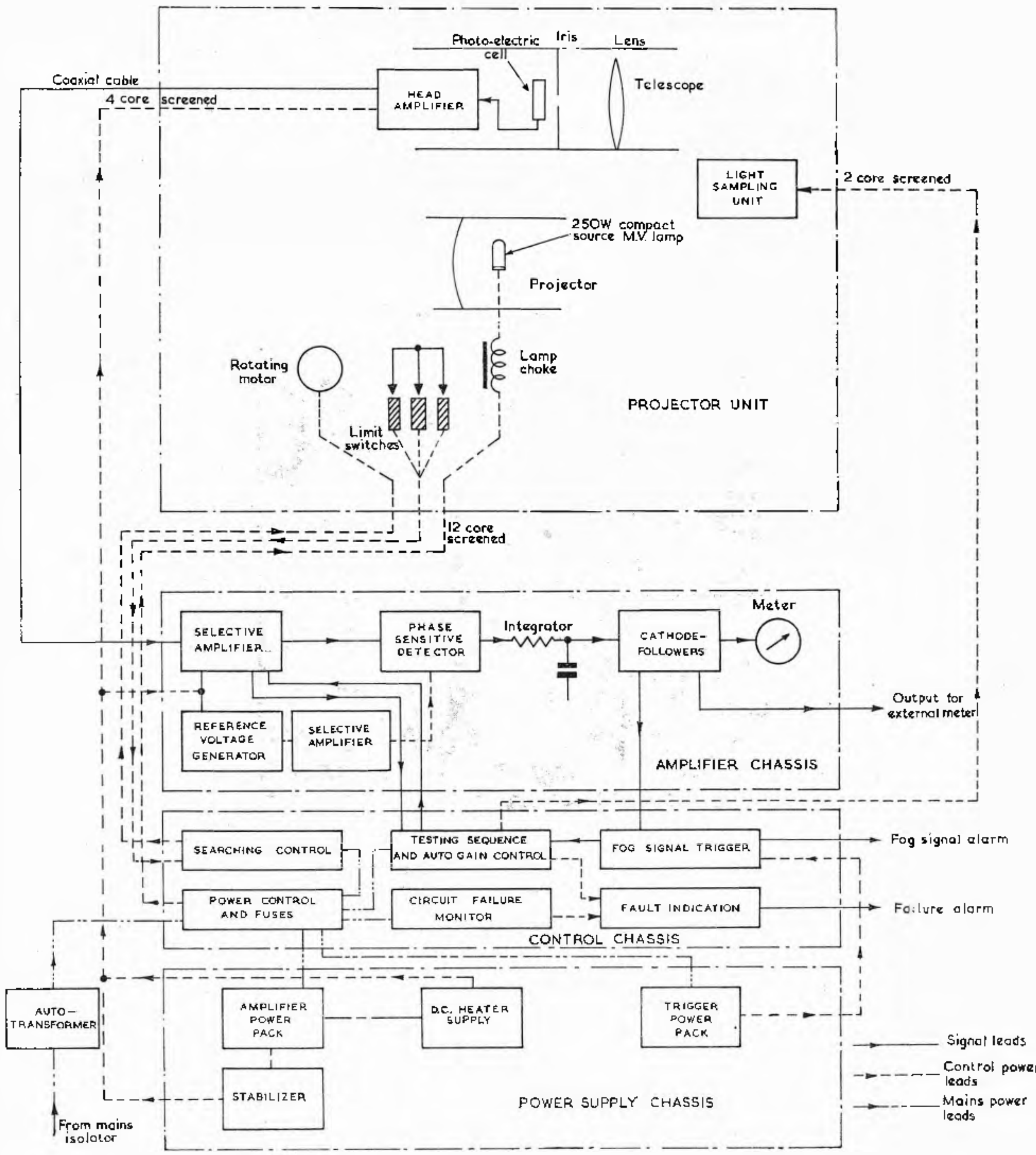


Fig. 2. General arrangement of the fog detector unit

by the omission of the h.t. decoupling which in this system is extensive, since the overall gain of the amplifiers is about 120dB at 100c/s and besides the possibility of oscillation unwanted 100c/s signals must be kept to a minimum.

The first two stages of amplification are mounted on the telescope itself adjacent to the photocell in order both to which restricts the high frequency response making an early limitation to the noise spectrum handled.

Valve microphony is a problem in these early stages which has been solved in this case by mounting the valve in a large block of metal which completely surrounds the glass envelope, the valve being of the 'wired in' type. The

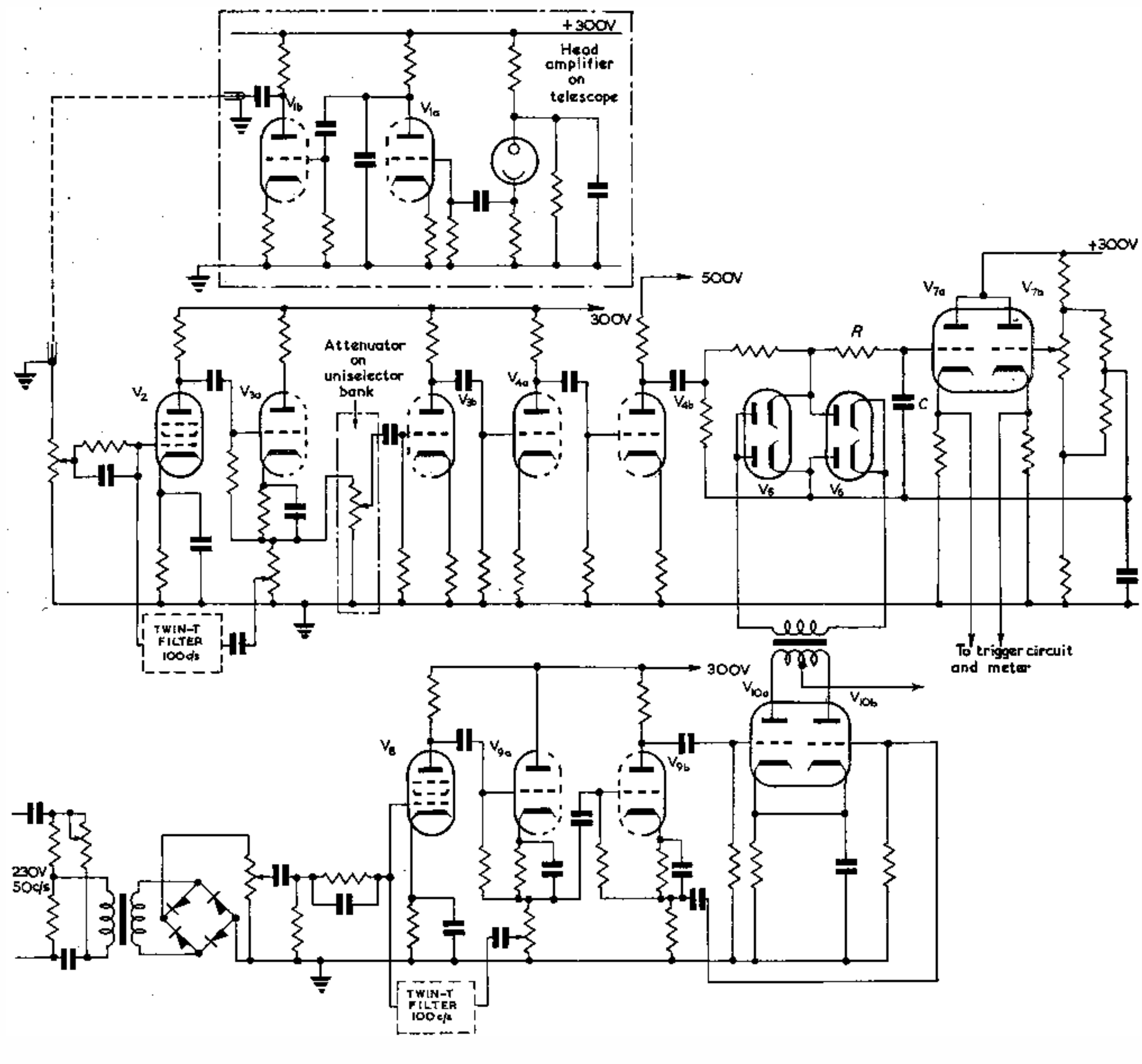


Fig. 3. The amplifier and detector stages

reduce the output impedance and raise the signal level before transmission from the rotating mechanism to the main panels. These two stages formed by V_{1a} and V_{1b} are of conventional design, it being noted that the photocell output is a.c. coupled to the grid of V_{1a} to avoid changes in bias level which would otherwise arise from different ambient lighting levels. One other point is the introduction of a shunt capacitor to the anode of V_{1a} heater circuit of valve V_1 is fed with d.c. very heavily smoothed to avoid hum pick up.

The signal from the relatively low impedance anode load of V_{1b} is fed over a screened cable to the remaining stages which are housed in the main equipment panel. It has been found that with separations of the main panel from the telescope of up to 100 yards the equipment will perform satisfactorily.

In the main amplifier the signal is first applied to a stepped attenuator by means of which the overall gain may be set to the nearest 3dB. The output of this attenuator feeds the selective amplifier formed by V_2 and V_{3a}. The frequency selective characteristic is obtained by the provision of a twin-T band-stop filter[1] centred on 100c/s in the feedback loop between the cathode of V_{3a} and the grid of V_2. The bandwidth of the amplifier is made variable by a potentiometer in the cathode of V_{3a}. The bandwidth is reduced in this amplifier to about 10c/s.

The output of V_{3a} is also used to feed a further stepped attenuator formed by resistors connected to the bank contacts of a uniselector which forms a vital part of the automatic gain setting system described later.

From here the signal is fed via conventional voltage amplifiers, V_{3a} and V_{4a}, to V_{4b} which forms the driving stage for the coherent detector system. To enable a large signal voltage to be handled in V_{4b} the anode is fed from 500V.

Before discussing the detector system it is worth noting the noise limiting devices used up to this stage and the circumstances which dictate their necessity. These consist of the shunt capacitance on the anode of V_{1a} and the selective amplifier, both of which are introduced as early as possible into the amplifier chain. It is necessary to do this to eliminate further increases in noise due to cross-modulation products were the level of noise allowed to exceed the grid base of the valves used.

The coherent or synchronous detector[2] formed by V_5 and V_6 is fed from the anode of V_{4b} via an isolating capacitor and resistor. An integrating circuit with a long time-constant is formed by the following resistance capacitance combination. The detector cannot in this case be returned to earth as coherent signals at 100c/s may arise from various sources such as heater current, h.t. supplies, etc. The effect of such coherent signals is to give rise to a standing voltage on the capacitor C which may be either positive or negative. To provide for this without the introduction of a stabilized negative voltage for the subsequent cathode-follower V_{7a}, the detector has been returned to a positive point across a potential divider chain. A further cathode-follower V_{7b} has its grid connected to a similar potential divider chain. By adjustment of the grid potential of V_{7b} it is possible to arrange for the two cathodes of V_{7a} and V_{7b} to be at the same voltage. Wanted signals will now give rise to a positive voltage on the cathode of V_{7a} with respect to that on the cathode V_{7b}. A meter connected between these two cathodes may be used as a measure of visibility, the two cathodes also being fed to the trigger circuits.

The coherent detector must, however, be fed with a voltage which bears the correct phase relationship to the wanted signal waveform at the anode of V_{4b}. This is derived from the circuit shown in the lower part of Fig. 2. A supply of 230V 50c/s from the same source as the lamp supply is fed via a phase shifting network to a transformer, to the secondary of which is connected a full-wave bridge rectifier. The output of this rectifier contains a large proportion of 100c/s signal. This signal is fed to a selective amplifier V_8 and V_{9a} which is identical to that formed by V_2 and V_{3a} in the signal circuit. By making these two selective amplifiers identical, independence of variation of supply frequency is insured since the phase shift introduced in one channel will be met by an equal phase shift in the other. It has been suggested, although not observed, that the light output of the mercury vapour lamp may not bear a constant phase relationship with the supply voltage. If this is proved to be the case the input to V_8 will be taken from a photocell placed in the main lamp enclosure.

The output of V_{9a} drives the phase-splitting stage V_{9b} which, in turn, drives the output stages V_{10a} and V_{10b} in push-pull. Transformer coupling to the detector bridge provides a convenient method of impedance matching.

Functional Check and Automatic Gain Setting

The output between the two cathodes of V_{7a} and V_{7b} is fed to a Schmitt trigger circuit which has variable controls for both trigger level and recovery level. It is necessary, however, to supply h.t. power to this circuit from a floating supply. A relay in one half of the Schmitt trigger provides the circuit for sounding the fog signal. As, however, the equipment is rotated to search for fog over an arc up to 360° it is necessary to provide additional relays which ensure that after the trigger relay has operated it cannot be released until the search has completed one full turn in the opposite direction to that in which the original operation took place. By means of the variation in operate level and recovery level it is possible to ensure that although the relay may operate when the visibility falls to, say, 2 miles, it will not de-energize until the visibility has improved to, say, 3 miles. This avoids having the fog signal spasmodically started and stopped during conditions of varying visibility.

In an equipment of this nature where performance depends on maintaining the electrical gain of the amplifiers within close limits, it is necessary to introduce some form of automatic gain control. The system employed here also makes corrections if the light output of the lamp changes or the glass surfaces of the searchlight or lens become dirty. In addition, it is desirable to check that the setting of the trigger level is correct.

To provide all these features it is arranged that a sample of modulated light from the projector taken outside all the surfaces likely to become dirty is introduced into the telescope. At the same time any light entering the telescope from back scatter in the atmosphere is cut off from the telescope. This checking system operates once for each complete traverse of the rotating mechanism. The amount of light entering the telescope is adjustable by a mechanical system of holes in two plates which slide over one another and may be locked in a pre-set position. In this way one may arrange that under the test conditions the photocell output is the same as would be obtained were the visibility, say, 2 miles. In order that the check and gain setting may take into account trigger sensitivity the operation or otherwise of this circuit is used to adjust the gain of the amplifier.

The sequence of operations is then as follows:—

(1) The search is stopped at a predetermined position.

(2) Any charge on the integrating capacitor following the synchronous detector is removed by short-circuiting it.

(3) Light sample is admitted to the telescope and short-circuit removed from capacitor. At the same time the time-constant of the integrator is reduced to speed up the test.

(4) After an interval equal to this time-constant a test is made to check whether the trigger has operated. During the test the operation of the relay is naturally isolated from the fog signal sounding system.

(5) If the trigger has operated, a uniselector controlling the gain of the amplifier is stepped so as to reduce the gain by one increment, after which the capacitor is discharged, the light sample is removed and the rotation re-started.

(6) If, on the other hand, the trigger has not operated, the uniselector is stepped to increase the gain by one increment and after a further interval equal to the integrator time-constant the check is repeated. This process continues until either the trigger operates or the uniselector reaches maximum gain at which point a fault alarm signal is given.

It is not proposed to describe in full the operation of the relay circuits but it may be worthy of note that stepping of the uniselector, which may be stepped forward or backwards, is secured by the discharge of a capacitor charged to 300V thus avoiding a separate low voltage supply.

The result of this testing is to ensure that the equipment retains its trigger setting around the visibility level at which it is decided that a fog signal should be sounded.

Results

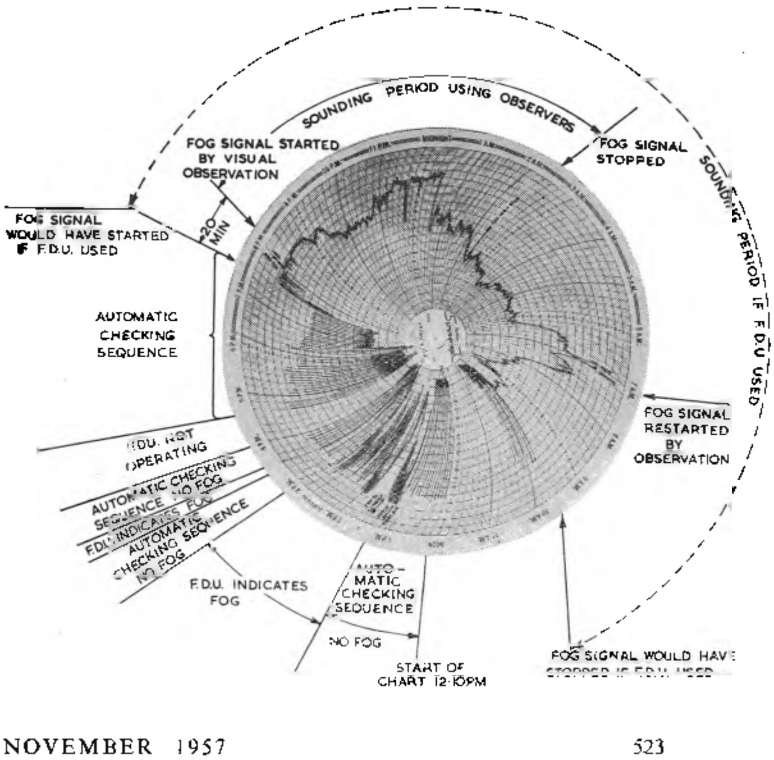
Equipments of laboratory design involving the principles stated above have been in operation over a period of two years, during which a large number of readings of output voltage and observable visibility have been taken. The observed visibility was assessed by noting the number of black and white chequered boards spaced at known ranges which were visible. The results have shown a high degree of consistency and have agreed fairly well with a somewhat crude theoretical assessment. Possibly, however, of more significance is that whenever the equipment was operational on no occasion has a fog condition been reported by the lighthouse keepers when it was not also recorded by the instrument. Furthermore, the performance of the instrument is consistent between day and night, a condition which is nearly impossible with human observers.

A typical 24-hour record chart is shown in Fig. 4.

Prototype manufactured models to a detailed design have been made by Messrs. Stone Chance Limited under a licence granted by the National Research Development Corporation who, in conjunction with Trinity House are pursuing the patent applications.

The electronic control cabinet is shown in Fig. 5.

Fig. 4. Typical record chart showing variations in visibility recorded from fog detector unit 19-20 July 1956



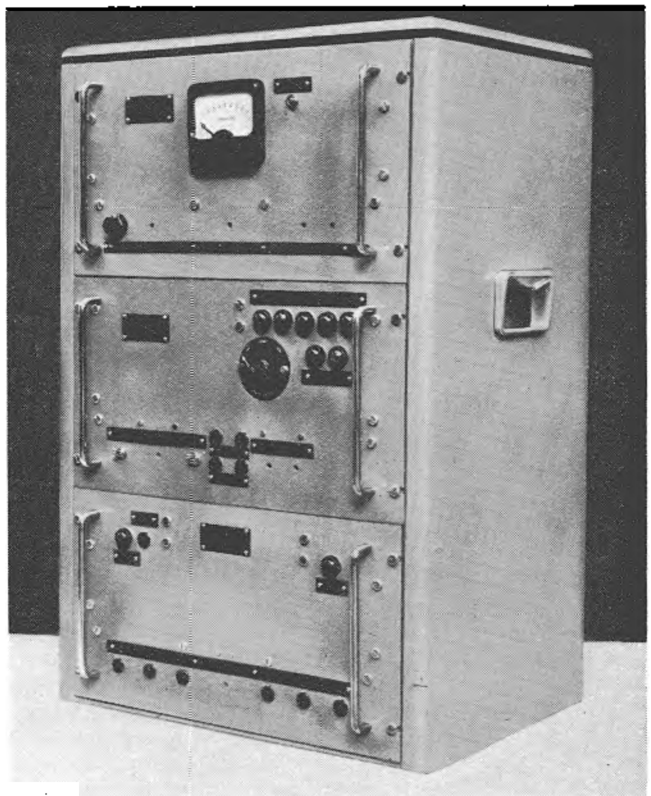


Fig. 5. The electronic control cabinet

Future Development

The equipment described with a rotating beam is suitable only for shore sites or rock lighthouses, and would not lend itself to installation on a light vessel. To overcome these difficulties a second version making use of a drum lens giving a beam of a few degrees in the vertical plane with 360° in the horizontal plane is being developed. A number of lenses are used to provide about eight telescope systems which are focused into four photocells. The outputs of these are added and fed to amplifiers of similar design to those described. Because, however, of the lower optical sensitivity higher amplifier gain is required and to obtain similar signal-to-noise ratios in this equipment the overall bandwidth is reduced to about 1/500sec by the provision of a very long integrating time-constant.

It is hoped that such a system may be stabilized for use on light vessels. It could also be usefully employed on other vessels in order to indicate that the visibility had fallen and it would clearly be possible to connect the output voltage from such an instrument with a radar set to indicate by a range ring the existing visibility. Such a system might provide some help in reducing the possibility of collisions at sea.

REFERENCES

1. Hillan, A. B. Parallel T-Bridge Amplifier. *J. Instn. Elect. Engrs.* *94*, Pt. 3, 27 (1947).
2. Tucker, D. G. The Synchrodyne Receiver. *Electronic Engng.* *19*, 241 and 276 (1947).

A Design Method for

Wide Band Balanced and Screened Transformers

in the Range 0·1 — 200 Mc/s

By M. M. Maddox*, B.Sc., and J. D. Storer*, A.M.I.E.E., A.M.Brit.I.R.E.

A method for the design and construction of wide band transformers is described in which the physical characteristics of the transformer are determined so that it appears as an ideal transformer together with a symmetrical low-pass filter with the appropriate iterative impedance. The winding, which may be encapsulated, is screened and has a form which achieves a high degree of balance. Examples are given of transformers so designed having an insertion loss of about 0·5dB together with a good balance and impedance match over a wide frequency range.

IN a survey of wide band transformers currently used in aerial feeder systems at radio receiving stations a requirement was seen for a wide band transformer having an improved performance and which was at the same time cheap and easy to produce; it was this requirement which stimulated the work of which the present article is the subject. Although wide band transformers have applications in many fields, e.g. video amplifiers, measuring apparatus, it should be borne in mind throughout that the particular application to aerial feeder systems is the underlying theme.

H.F. aerials are commonly erected as far as is practicable from the receiving station and transmission lines up to a mile long are used to feed the signals to the receivers. The transmission line may be either coaxial cable, in which case a balanced to unbalanced transformer is used to match the aerial to the cable, or a combination of an open wire balanced feeder and coaxial cable, when two transformers are required, the first, a balanced to balanced, to match the aerial to the open wire feeder, and the second, a balanced to unbalanced, to match the open wire feeder to the coaxial cable. At 20Mc/s the loss of open wire feeders is about 1dB/mile and that of coaxial cable, 0·2dB/100ft. It is therefore apparent that the insertion loss of the matching transformers must be extremely low.

* Government Communications Headquarters.

List of Symbols

L_p	= Primary inductance
C_{sp}	= Electrostatic capacitance between screen and primary
C_{cp}	= Electrostatic capacitance between core and primary
C_{wp}	= Electrostatic capacitance of primary winding
C_{ss}	= Electrostatic capacitance between screen and secondary
C_p	= Effective primary capacitance
C_s	= Effective secondary capacitance
L_L	= Leakage inductance } referred to the primary
R_w	= Core losses } referred to the primary
R_L	= Load resistance
Q	= Normalized susceptive component of primary admittance for given r.
f_1	= Lower frequency limit of the transformer
f_2	= Upper frequency limit of the transformer
f_c	= Cut-off frequency of equivalent low-pass filter
Z_o	= Iterative impedance of equivalent low-pass filter
Z_2	= Impedance of equivalent filter at f_2
N	= Turns per μH factor for core
n	= Turns ratio
n_p	= Primary turns
n_s	= Secondary turns
r	= Standing-wave ratio

Losses, other than transformer core losses, will be introduced if there is mismatch between aerial and feeder or between feeder and coaxial cable. The standing-wave ratio (s.w.r.) is a convenient measure of impedance match and this must therefore be as close to unity as possible. This is particularly important in the case of an aerial, e.g. the heptagon, which must be accurately terminated. An s.w.r. of 1·25 over the band is considered to be the worst mismatch which can be tolerated in such instances—this has been taken as a guiding figure in making transformer specifications.

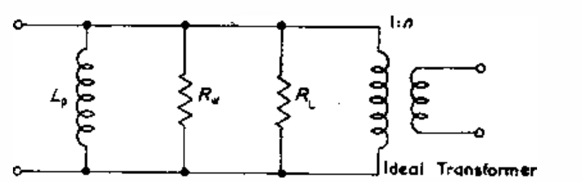


Fig. 1. Low frequency equivalent circuit of a transformer referred to the primary

Further, any unbalance in the transformers will result in a greatly increased noise pick-up and crosstalk in the feeders and hence a high degree of balance, defined for the purposes of this article as twenty times the logarithmic ratio of the voltage across the balanced winding to the out-of-balance voltage, is required.

General Considerations

It is well known that a physical transformer is replaceable by an ideal transformer coupled with a network, whose form depends on whether high or low frequencies are under consideration.

At low frequencies the transformer is equivalent to the circuit of Fig. 1 where L_p is the inductance of the primary winding and R_w is a resistive shunt which represents the core losses referred to the primary. The insertion loss, and hence s.w.r., depends on the admittance of this parallel combination—the lower this admittance the better the performance of the transformer. R_w and L_p are determined by the efficiency of the core and the number of turns on the primary winding; thus for any specified transformer it is necessary that there should be sufficient turns to provide a maximum shunt admittance at the lowest frequency.

In the high frequency case when the admittance of the R_wL_p combination may be neglected the equivalent circuit referred to the primary becomes that shown in Fig. 2. The primary and secondary capacitances C_p and C_s, and the total leakage inductance L_L become the controlling factors and form a π section low-pass filter. The technique is to so arrange the winding form that C_p and n^2C_s are equal and of such a value that when combined with L_L the filter becomes a constant K π section with an iterative impedance equal to that of the primary impedance of the required transformer.

Standard filter equations may be applied so that if Z_o is the iterative impedance and f_c the cut-off frequency,

$$Z_o = (L_L/C)^{\frac{1}{2}} \text{ and } f_c = \frac{1}{\pi(L_LC)^{\frac{1}{2}}}$$

where $C = 2C_p = 2n^2C_s$.

The cut-off frequency is thus determined by the leakage inductance and the shunt capacitances. In practice the latter can be kept small and need not increase with the number of turns. The leakage inductance, however, increases with the number of turns, i.e. f_c decreases, and it is apparent that the upper and lower frequency limits are related and have conflicting requirements. A core must therefore be used

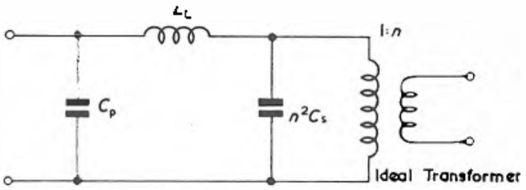


Fig. 2. High frequency equivalent circuit of a transformer referred to the primary

which will permit the maximum shunt admittance to be obtained with the fewest possible turns.

In order to produce a transformer which has a good balance ratio it can be shown that the winding must be of such a form as to have equal coupling between the primary and each half of the secondary and equal winding capacitances for each half of the secondary. Hence in the particular case of a balanced to unbalanced transformer this further requirement must be met.

Two constructions are widely used to meet these requirements. The first is a two winding transformer toroidally wound on a ferrite or mumetal ring core, with a copper screen separating the windings. With this method the individual elements of capacitance and leakage inductance are difficult to control and external compensating circuits are usually included. Winding capacitances tend to be high and restrict the cut-off frequency. The second is an auto-transformer with a concentric spiral winding of copper tape, again on a ring core of ferrite or mumetal. The auto-transformer has a greatly reduced leakage inductance and quite good bandwidths can be obtained as a result, but it suffers from an inherently bad balance ratio because the coupling between the primary and each half of the balanced winding are not equal. The effect depends on the number of turns and becomes worse with increasing turns ratio.

Neither of the transformers quoted above satisfies the conditions for optimum performance. A construction is required in which the leakages and winding capacitances can be accurately and independently controlled and which enables a high degree of winding balance to be attained. It will further be of advantage if the transformer is compact and can be wound on a pot type core which will itself provide adequate screening without external additions.

The present design provides a construction which complies closely with these requirements and which enables a greatly improved performance to be obtained.

Constructional Details

The transformers can be constructed using either a single layer of wire or a concentric spiral of tape for the primary winding. The latter has the disadvantage that the winding capacitance and hence the total primary capacitance increases rapidly as the number of turns is reduced. The winding capacitance of the layer winding represents such a small part of the total primary capacitance, that no such effect is present; however, the leakage inductance of this type of winding is greater than that of the spiral, and for most cases the spiral winding is preferred.

The primary is wound on a thin former built up by winding polystyrene or other suitable tape round the centre rod of a Ferroxcube type LA1 pot core assembly.

This core was chosen because of its high initial permeability, and is preferred to an 'E' type core as the winding is totally enclosed in a high permeability shell reducing stray flux to a negligible value. There is therefore no need for a screening can and the assembly may be potted in a suitable casting resin, or contained in moulded polythene as required. The main requirement of the core is that it should permit the required shunt admittance to be obtained with the minimum number of turns, the ring piece is therefore ground to reduce the gap as far as is practical resulting in a considerable increase in shunt inductance for a given number of turns. This process is controlled by testing each core with a standard winding.

For a balanced primary winding of either type a single unshorted turn of copper tape is wound over the completed primary to form an electrostatic screen, and connected to the core. The thickness of insulation between winding and core, and winding and screen, is arranged to ensure equal winding-core and winding-screen capacitances. In the case of an unbalanced layer wound primary, one end of the winding is connected to the screen, however, it should be noted that for reasons detailed later it is not possible to take a transformer designed for balanced working and convert it to unbalanced by simply earthing one side of the primary. When a spiral winding is used for an unbalanced primary, the last turn is connected to the core and acts as a screen.

The material and thickness of the winding to core and winding to screen insulation together with the width of the winding (and in the case of the spiral, the material and thickness of the inter-leaving material), are all controllable and determine the primary shunt capacitance.

A single layer wire secondary is wound over the screen or the last turn of the spiral as the case may be to provide a symmetrically placed winding, the gauge of wire is chosen so that a winding length appropriate to the required capacitance is obtained. This ensures equal coupling between the primary and each half of the secondary, also equal winding capacitance from each half of the screen. The leakage inductance can be controlled by the thickness of insulation between the screen and the secondary winding, as can the secondary capacitance. The latter is also controlled by the material used for this layer, and by the secondary winding length, i.e. the wire gauge used. For some applications, e.g. aerial transformers, a static leak is required. The secondary is then centre-tapped and the tap connected to the core. As this point is at core potential due to the balanced condition of the winding, making the connexion has no effect on the performance.

A further variation is to use a single layer wire primary and a concentric spiral tape secondary. In this case the first turn of the secondary spiral is earthed and acts as a screen, this form of winding is physically more difficult to achieve, and is only advantageous in certain circumstances where a small secondary capacitance is required. It will therefore not be discussed in detail, but the arguments relating to the other forms of winding are applicable.

The Components of the Equivalent Circuits

Primary Inductances

The amount of primary inductance L_p that is required for a given s.w.r. or insertion loss is dependent on the admittance of the combination of L_p, R_w, and R_L in Fig. 1. It has been pointed out previously that these considerations are only valid at low frequencies, and it is therefore necessary to make L_p sufficiently large to meet the requirements of s.w.r. at the lower limit, f_1, of the frequency range specified. A simplified method of calculating L_p sufficiently accurate for practical applications is as follows.

Consider the parallel combination of the transferred load R_L and the primary shunt inductance L_p, and let the s.w.r. be limited to r. The normalized admittance of R_L and L_p must fall at a point on a Smith Chart within a circle of radius r. The normalized conductive component will remain constant at 1·0 and the range of the suspective component may be found as the intersection of the unity constant conductance circle and the circle of s.w.r. equal to r. Let this value be Q, then

$$Q = (R_L/2\pi f L_p) \text{ or } L_p = (R_L/2\pi f Q)$$

Thus if the chosen s.w.r. is 1·25, f_1 is 1Mc/s and R_L is 80Ω, from the Smith Chart $Q = 0{\cdot}225$ and hence

$$L_p = \frac{80 \times 10^6}{2\pi \times 10^6 \times 0{\cdot}225} = 56{\cdot}6\mu\text{H}$$

If N is the turns per μH factor for the core, the primary turns required for an inductance L_p are given by $n_p = N(L_p)^{\frac{1}{2}}$. The minimum attainable leakage inductance, L_L, increases with primary turns, and as low values for L_L are necessary at high frequencies it is desirable to make N small. As described previously, N is decreased by grinding the ring piece of the pot core assembly. Using LA1 material it is found convenient to make N unity, and hence $n_p = (L_p)^{\frac{1}{2}}$; thus 7 turns will result in 40μH, 8 turns in 64μH etc.

Turns Ratio

The turns ratio, n, is the square root of the required impedance ratio, and the secondary number of turns n_s is given by $n \cdot n_p$.

Leakage and Lead Inductance

For a given winding form the minimum value for leakage inductance depends on the number of primary turns, but values above this minimum can be obtained by reducing the coupling factor. This is achieved by increasing the thickness of insulation between the last turn of the spiral winding and the layer winding.

In order to make practical use of these two variables in the design of a specific transformer it is advantageous to plot a number of curves as in Graph D. This is seen to consist of a family of curves of leakage inductance against thickness for various numbers of primary turns. Their use will be apparent from the examples of transformer design given later and here will only be explained the method of producing them.

Any convenient number of turns may be wound on a core and a secondary winding of the desired number of turns of suitable wire gauge is chosen as above. The leakage inductance is then measured by a short-circuit method for varying thicknesses of interwinding insulation. In Graph D an 8-turn primary winding was chosen and L_L was measured for interwinding thicknesses of ·003in. and ·006in. This is represented diagrammatically in Fig. 3. The points A, B, C . . . can then be plotted. The curve through A, B, C . . . then represents the relation leakage inductance/thickness of interwinding insulation for an 8-turn primary winding. At each point such as A, the number of primary turns per μH of leakage, N_L, may be obtained and then using the relation $n_p = N_L(L_L)^{\frac{1}{2}}$ the leakage may be calculated for various values of primary turns. Points A′, A″ . . . are then plotted corresponding to these values and the family of curves A′ B′ C′ . . . , A″ B″ C″ drawn. The series of curves shown in Graph A was plotted for a tape width of 0·25in. Theoretically a series should be plotted for each tape width used, but in practice, the variations of leakage with tape width has been found to be small.

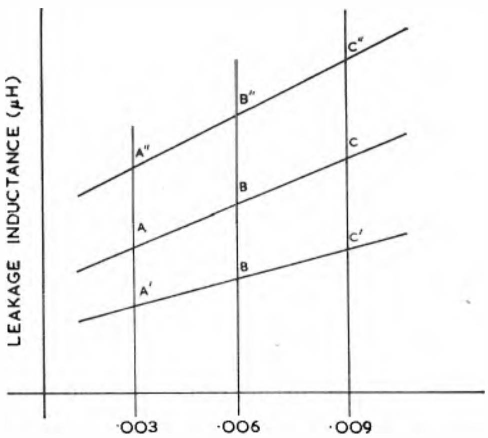


Fig. 3. Relationship between interwinding insulation and leakage inductance

When the transformer requirements, e.g. impedance or cut-off frequency, involve leakage inductances which are small it is necessary to take special care to reduce the loop inductance of the lead-out wire. This may be done by keeping the leads short and close together. A working figure for the loop inductance is 10 per cent or less of L_L. Using 24 s.w.g. wire, a lead 1in long has an inductance of about 0·02μH and referring to graph A it will be seen that this is only significant in the case of a 4-turn primary or less.

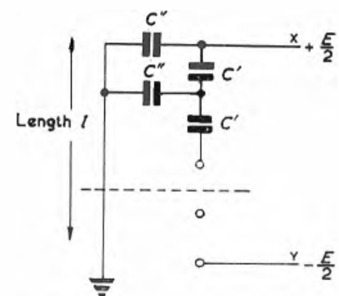


Fig. 4. Distribution of the capacitances of a single layer winding over a conducting former

Secondary Capacitance

When a single layer coil is wound on to an earthed conducting former, the capacitive effect exhibited by the coil is modified by the balance or unbalance of the voltage applied to it. Consider the cross-section of a winding shown in Fig. 4. The capacitance seen looking into the leads XY is a function of the capacitances C' between adjacent turns and the capacitances C'' of each individual turn to the former. The capacitances C' are small and add in series and hence may be neglected. The effective capacitance across XY is thus some combination of the capacitances C''.

Suppose the winding to be of length l and let the electrostatic capacitance between the layer and the former be C. Let a balanced voltage E be applied to the coil, i.e. X and Y are at potentials $+E/2$ and $-E/2$ respectively. The voltage distribution along the coil is thus as in Fig. 5.

Let the capacitance per unit length be p, i.e. $C = pl$, and consider an element δx of l at a distance x from the centre.

The current flowing through the elemental capacitance formed by x is given by

$$\delta I = j\omega p \delta x \, . \, E/2 \, . \, x/(l/2)$$

Hence the total current flowing from the positive side of the layer to the former is I, where

$$I = j\omega p \, . \, E/l \, . \int_0^{l/2} x \, . \, dx$$

$$\text{or } I = (j\omega C/8) \, . \, E$$

An equal current flows from the former to the negative side of the layer and thus, looking into XY, the winding appears to have a capacitive current I flowing into it.

But the current flowing through a capacitance C due to a voltage E is $i = j\omega C \, . \, E$. Hence the effect of the balanced voltage distribution across the winding is to make its capacitance appear as $\frac{1}{8}$ of its electrostatic value.

Using a similar argument the effective capacitance of the winding when an unbalanced voltage is applied to it is $\frac{1}{2}$ its electrostatic value.

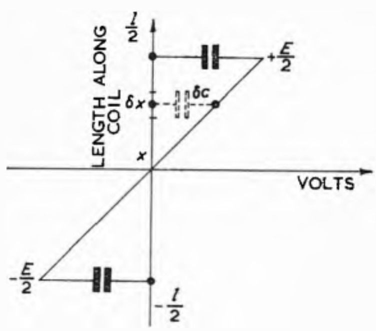


Fig. 5. Voltage distribution along the length of a balanced winding

The electrostatic capacitance between the layer and the screen may be regarded as that of a cylindrical capacitor so that

$$C_{ss} = \frac{0{\cdot}6128 \times l \times k}{\log (r_2/r_1)} \text{ pF}$$

where l is the winding length in inches, k the dielectric constant of the interleaving material and r_1, r_2 the radii of the screen and the layer winding respectively. An increment applied to both r_1 and r_2 has little effect on the result, thus small variations in the number of primary turns do not effect the capacitance.

Because of the circular cross-section of the wire, it is necessary to add a weighting factor to r_2, this will depend on the gauge of wire used, and is only effective when the interwinding insulation is less than ·003in in thickness. For design purposes it is convenient to plot a graph of C_{ss} against tape width and thickness of primary/secondary insulation. Graph C has been plotted with a weighting factor added to value of r_2 below ·003. This gives a reasonable average for wire gauges between 32 and 38 s.w.g.

Primary Capacitance

The capacitance of the layer winding is made up of the winding to core and the winding to screen capacitances C_{cp} and C_{sp} respectively. As in the case of the secondary, the inter-turn capacitance may be neglected and C_{cp} and C_{sp} can be considered as cylindrical capacitors. As before the static capacitance is modified by the voltage distribution with the result that for an unbalanced winding.

$$C_p = \frac{\frac{1}{2}(C_{cp} + C_{sp})}{2}$$

and for a balanced winding

$$C_p = \frac{(C_{cp} + C_{sp})}{8}$$

The capacitance of the unbalanced winding in the form of a concentric spiral is made up of the winding capacitance C_{wp} and the first-turn-to-core capacitance C_{cp}. The winding capacitance may be regarded as that of a series connexion of concentric cylinders and the expression of C_{wp} becomes

$$C_{wp} = \frac{(0{\cdot}6128)\, kw}{\log \left(\dfrac{r_0 + t_1 + (n-1)\, t_2}{r_0 + t_1}\right)} \text{ pF}$$

where k is the dielectric constant of the interleaving material

w is the tape width in inches

r_0 is the sum of the radius of the core + the thickness of the former

t_1 is the thickness of the conductor

n is the number of turns

t_2 is the thickness of the interleaving material.

C_{cp} may also be considered as the capacitance of a cylindrical capacitor formed between the core and the first turn of the primary winding.

$$\text{i.e. } C_{cp} = \frac{(0{\cdot}6128)\, kw}{\log (r_0/r_1)} \text{ pF}$$

where r_1 is the radius of the core and other symbols have the same meaning as before. Small variations in former thickness do not significantly effect the calculation of C_{wp} as any such small increment affects both the numerator and the denominator of the expression in the logarithm. When a balanced winding is under consideration, the capacitance between the last turn and screen C_{sp}, must be included, this is again considered as a cylindrical capacitor where r_1 is the radius of the last turn and r_0 that of the screen. To maintain balance C_{sp} must be made equal to C_{cp} and the primary capacitance which is again modified by voltage distribution becomes

$$C_p = \frac{\frac{1}{2}(C_{cp} + C_{sp}) + C_{wp}}{8}$$

for the balanced condition, and

$$C_p = \frac{C_{cp} - C_{wp}}{2} \text{ for the unbalanced condition.}$$

Practical Design

Frequency Limits and Choice of Winding Form

The lower and upper frequency limits f_1 and f_2 have conflicting requirements in that a low f_1 requires a large primary inductance which results in a large leakage inductance reducing the limit for f_2, and vice versa. Accordingly, in design either f_1 or f_2 may be taken as a starting point and the bandwidth arranged to accommodate the other limit. The relationship between f_2 and the cut-off frequency f_c of the equivalent filter is given by:

$$Z_2 = \frac{Z_0}{[1 - (f_2/f_c)^2]^{\frac{1}{2}}} \text{ where } Z_0 = (Z_1/r)$$

thus for $r = 1{\cdot}25$, $f_2 = 0{\cdot}6 f_c$. Applying the standard filter equations to the equivalent circuit referred to the primary

$$L_L = \frac{0{\cdot}6 \, . \, 10^6 \, Z_2}{\pi f_2} \, \mu\text{H and } C_p = n^2 C_s = \frac{0{\cdot}3 \, . \, 10^{12}}{f_2 \pi Z_0} \text{ pF}$$

From these equations it will be seen that f_2 depends on L_L, C_p and Z_0. For example a transformer having $f_2 = 200$Mc/s, $n^2 = 4$ and $Z_0 = 400\Omega$, would require $L_L = 0{\cdot}35\mu$H, and $C_p = n^2 C_s = 1{\cdot}2$pF. If Z_0 is reduced to 80Ω L_L becomes $0{\cdot}07\mu$H and $C_p = n^2 C_s = 6$pF.

It was remarked earlier that a single layer wire wound primary allows a low primary capacitance, but has a rather larger leakage inductance than that obtainable with a spiral tape winding. An indication of the limits on the performance which can be achieved using the types of winding described may be obtained from Fig. 6.

The factors limiting Z_o and f_2 are mainly the leakage inductance and the primary capacitance, all of which are inter-dependent. In practice the lowest values of the leakage inductance and primary capacitance which can be obtained simultaneously are approximately $0{\cdot}1\mu$H and 5pF respectively for a spiral tape primary winding. Hence using $C_p = (0{\cdot}3 \times 10^{12})/f_2\pi Z_o$, and letting $C_p = 5$pF, curve AL defines the limits on f_2 and Z_o imposed by C_p, all points

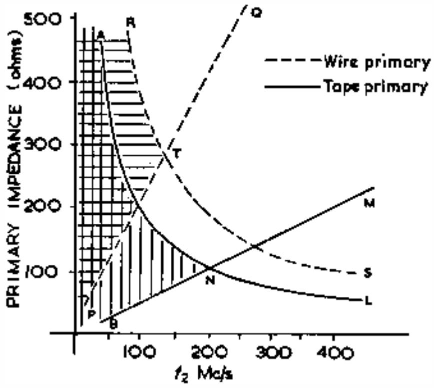


Fig. 6. Anticipated performance limits for tape and wire transformers using LA1 cores

(f_2Z_o) lying inside the concave portion of the curve being excluded. Similarly curve BM defines the limits imposed by L_L and all points (f_2Z_o) lying below BM are excluded. Thus the combinations of f_2 and Z_o which can be achieved in practice are those which lie within the region ANB of Fig. 6. The intersections of curves AL and BM correspond approximately to $f_2 = 200$Mc/s when $Z_o = 100\Omega$. Values of Z_o greater or less than this can be achieved only with a reduced upper frequency limit.

Again in the case of a single layer wire primary winding, the practical limits of L_L and C_p are approximately $0{\cdot}4\mu$H and 3·0pF, the corresponding curves for these values are shown in Fig. 6 by PQ and RS. The greater suitability of the single layer wire primary for higher impedances will immediately be seen, the permissible values being those which lie within the region RTP. In the cross-hatched region, both types of winding are equally suitable.

Materials and Graphical Aids

For design purposes it is convenient to plot graphs of the various functions. Graph A shows the relationship between the capacitances C_{sp}, C_{cp}, and the thickness of insulation, for various tape or winding widths. Graph B is a plot of the winding capacitance C_{wp}, against turns, tape width and interleaving thickness. The relationship between C_{ss} and screen to secondary insulation for a range of winding lengths is shown in Graph C; Graph D is plotted for the purposes of the examples which follow, from measurements of L_L as described previously.

Polystyrene tape was chosen for building up the various layers of insulation because of its low dielectric constant of 2·6 and because it does not stretch or flow when wound tightly. In the examples which follow a tape thickness of ·0015in was used as it was readily available. The copper tape used for spiral windings and screens was ·001in thick, the dimensions of the pot core limit the width to 0·375in maximum. For the wire windings enamelled copper wire was used. Table 1 gives the number of turns per 0·1in close wound for gauges between 20 and 40.

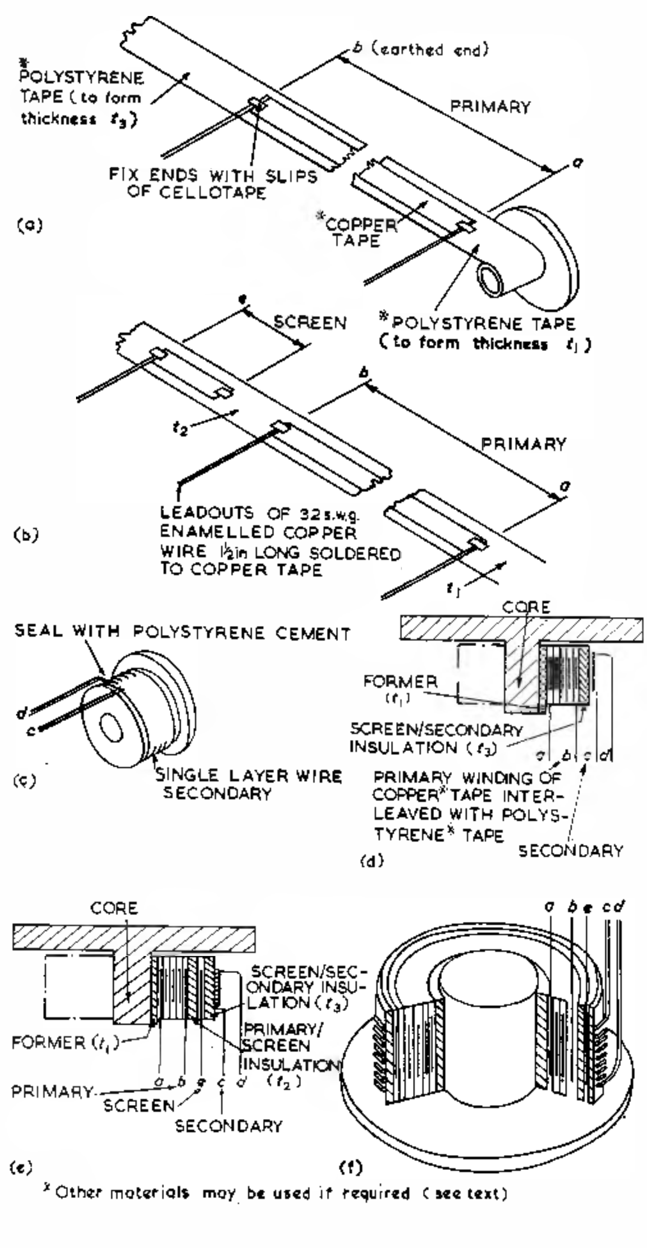
Illustrations of these constructions are given in Fig. 7

TABLE 1.

WIRE GAUGE	TURNS/0·1IN.	WIRE GAUGE	TURNS/0·1IN.
20	2·6	30	7·25
21	2·92	31	7·75
22	3·3	32	8·27
23	3·83	33	8·93
24	4·16	34	9·70
25	4·55	35	10·5
26	5·02	36	11·6
27	5·51	37	12·8
28	6·10	38	14·5
29	6·6	39	16·4
30	7·25	40	17·8

Fig 7. Constructional details of windings

(a) *Unbalanced tape primary*
(b) *Balanced tape primary and screen*
(c) *Secondary winding both types*
(d) *Section through balanced/unbalanced type*
(e) *Section through balanced/balanced type*
(f) *Isometric section of balanced type*



* Other materials may be used if required (see text)

Practical Examples

Design Procedure Based on f_1 Using Spiral Tape Primary

Let the requirement be for an 80Ω unbalanced to 640Ω balanced transformer having s.w.r. not greater than 1·25 at and between 1 and 40Mc/s. The primary inductance is found to be 56·6μH and therefore, using $n_p = (L_p)^{\frac{1}{2}}$, 8 turns are required for the primary winding.

Next, a thickness of interwinding insulation must be found which permits values of L_L and n^2C_s satisfying the standard filter equations:

$$L_L = (Z_o \cdot 10^6/\pi f_c)\mu\text{H and } n^2C_s = (10^{12}/2\pi f_c Z_o)\text{pF}$$

Eliminating f_c between these two equations gives

$$n^2C_s = (10^6 L_L/2Z_o^2)$$

For a balanced secondary:

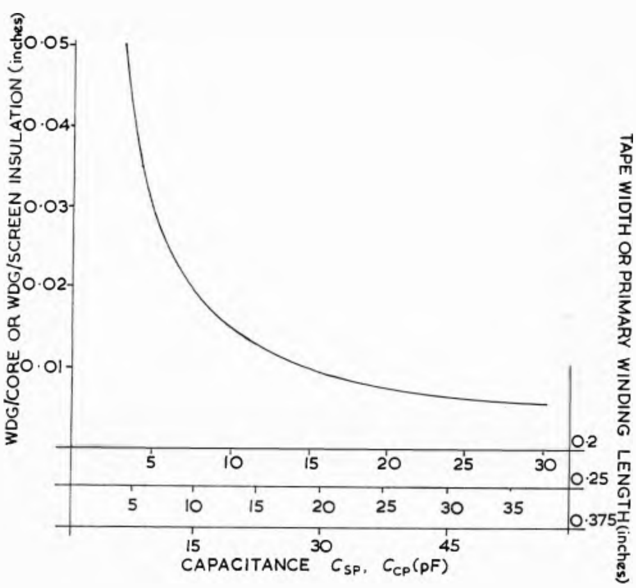
$$C_{ss} = (10^6 L_L/2Z_o^2) \cdot (8/n^2)\text{pF} \quad \ldots\ldots\ldots \quad (1)$$

Values of L_L for various interwinding insulation are now substituted in equation (1), and the values of C_{ss} obtained replotted on Graph C. The point of intersection of the two curves gives the required thickness. A different result will be obtained for each winding length scale, and any convenient set of conditions meeting the requirement of f_2 = 40Mc/s may be chosen. f_2 is determined by applying the following equation to each result.

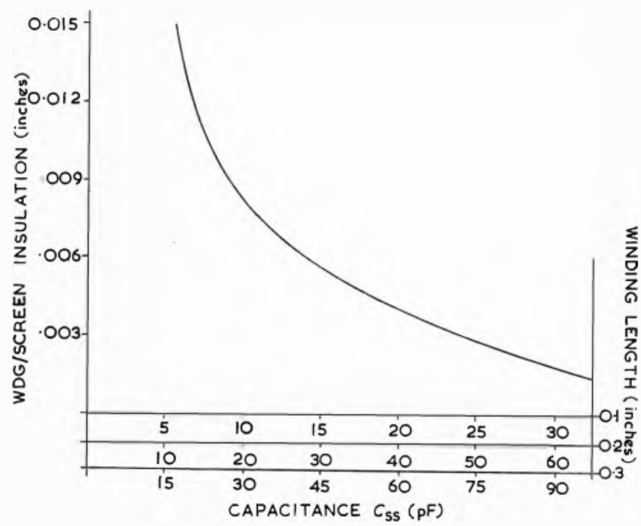
$$f_2 = (0{\cdot}6Z_o/\pi L_L)\text{Mc/s} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

These results are shown in Table 2

A winding length of 0·2in is chosen for convenience, and a primary tape width of 0·25in will be required to accommodate it. The winding capacitance C_{wp} of 8 turns of 0·25in tape interleaved with ·0015in insulation is shown by Graph B to be 18pF. In order to make $C_p = \frac{1}{2}(C_{cp} + C_{wp})$ equal to n^2C_s = 22pF, a former is required such that C_{cp} = 26pF. Using Graph A this is found to be ·007. The number of secondary turns is found by applying n_s =

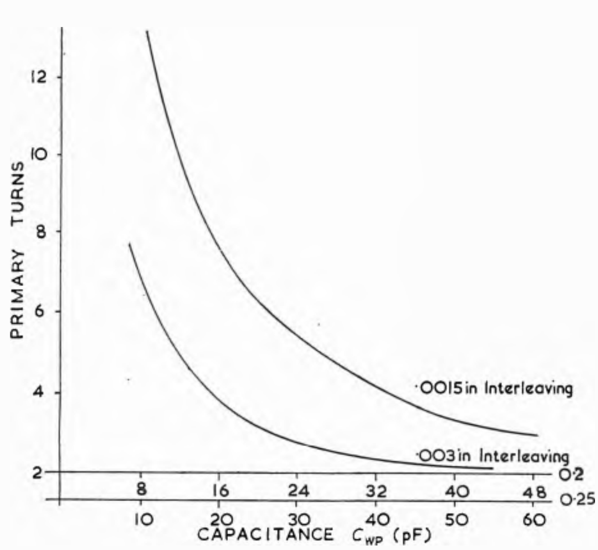


Graph A. Capacitance primary to core or screen (C_{CP}, C_{SP}) for various tape or winding widths and thicknesses of primary to core insulation



Graph C. Capacitance secondary to screen (C_{SS}) for various winding lengths and thickness of secondary to screen insulation

Graph B. Primary winding capacitance (C_{WP}) for various tape widths and primary turns spirally wound



Graph D. Relationship between leakage inductance and interwinding insulation for various primary turns

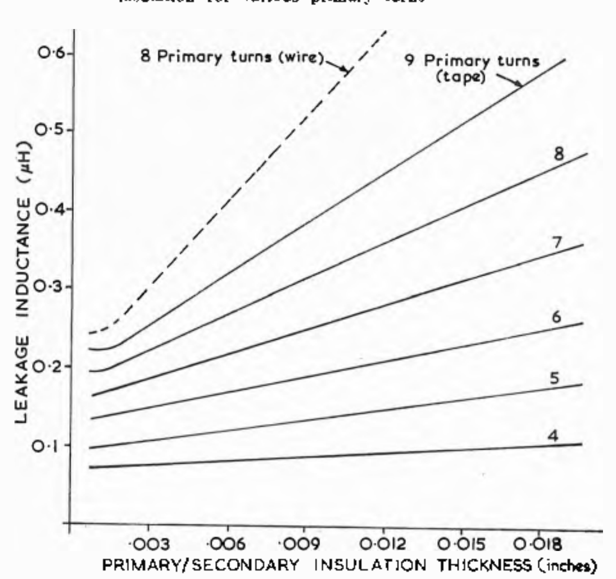


TABLE 2.

FROM GRAPH D		EQUATION (1)	FROM GRAPH C		FROM GRAPH D		EQUATION (2)
P/S INSULATION	L_L (μH)	C_{ss} (pF)	WINDING LENGTH	P/S INSULATION	C_{ss} (pF)	L_L (μH)	f_2 (Mc/s)
0·006	0·27	21	0·1	·0033 (2T)	21·5	0·22	70
0·018	0·45	35	0·2	·0072 (5T)	22·5	0·29	53
			0·3	·0117 (8T)	28·0	0·36	42.5

TABLE 3.

f (Mc/s)	INS. LOSS (dB)	BALANCE (dB)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.
1·0	0·5	50	0·96+j0·188	1·22	1·01 +j0·186	1·2
5·0	,,	,,	0·96+j0·042	1·06	1·025+j0·013	1·12
10·0	,,	,,	0·96+j0·021	1·05	1·03 −j0·015	1·12
20·0	,,	45	0·96+j0	1·04	1·05 −j0·05	1·05
40·0	,,	41	0·90+j0	1·1	1·00 −j0·04	1·09
60·0	1·0	38	0·77+j0	1·3	0·985−j0·015	1·16

$n \,.\, n_p$, to be 23 turns. In order to cover a winding length of 0·2in, 36 s.w.g. enamelled wire is required.

Thus the complete winding specification is:

Core:— Ferroxcube type LA1. (Ring piece ground so that $n_p = (L)^{\frac{1}{2}}$).

Former:— 5 turns ·0015in × 0·375in polystyrene tape.

Primary:— 8 turns of ·001in × 0·25in copper tape, interleaved with ·0015in × 0·375in polystyrene tape.

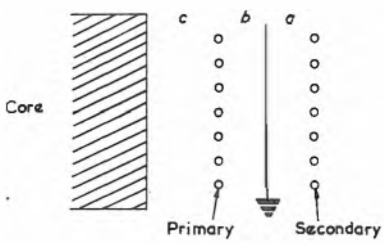


Fig. 8. Screen position using single layer wire primary

Interwinding insulation:— 5 turns of ·0015in × 0·375in polystyrene tape.

Secondary:— 23 turns of 36 gauge enamelled wire.

A transformer was constructed to this specification and measurements made of s.w.r. insertion loss and balance ratio, the results are shown in Table 3.

DESIGN PROCEDURE BASED ON f_1 USING SINGLE LAYER WIRE PRIMARY

Let the requirement be as before. Proceeding as above one finds that $n_p = 8$ and a secondary winding of 0·2in with three turns of insulation between primary and secondary results in $L_L = 0·36$ and $C_p = n^2C_s = 30$pF.

The three turns of insulation must be distributed over (a) and (b) in Fig. 8. As C_{sp} can be large without disadvantage, let (b) = one turn and (a) = two turns, then to keep C_{ss} to 30pF with the reduced screen to secondary separation, reduce the winding length proportionally to 0·15in. With (b) = one turn and a 0·25in winding length, C_{sp} = 60pF, to make $[(C_{cp}+C_{sp})/4] = 30$, C_{cp} must also be 60pF and therefore (c) must also be one turn. In order that the 8-turn primary covers 0·25in, 22 s.w.g. wire is required.

The complete winding specification is:

Core:— Ferroxcube type LA1 (ring piece ground so that $n_p = (L_p)^{\frac{1}{2}}$.

Former:— 1 turn ·0015in × 0·375in polystyrene tape.

Primary:— 8 turns 22 gauge enamelled wire.

TABLE 4.

f (Mc/s)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.
1	0·96 +j0·16	1·18	1·02 +j0·156	1·17
2	0·96 +j0·08	1·08	1·02 +j0·07	1·08
5	0·96 +j0·025	1·04	1·01 +j0·018	1·04
10	0·96 +j0·025	1·04	1·01 +j0·015	1·03
20	1·11 +j0	1·11	1·06 +j0·02	1·06
30	0·84 +j0·06	1·19	1·12 +j0·03	1·13
40	0·84 +j0·12	1·25	1·142−j0·08	1·17
50	0.77 +j0·204	1·42	1·11 −j0·013	1·12
60	0·705+j0·302	1·63	0·99 −j0·018	1·2

Primary/screen insulation:— 1 turn ·0015 × 0·375in polystyrene tape.

Screen:—1 unshorted turn ·001 × 0·375in copper tape.

Screen/secondary insulation:— 2 turns ·0015 × 0·375in polystyrene tape.

Secondary:— 23 turns of 40 s.w.g. enamelled wire.

A transformer was constructed to this specification and its s.w.r. measured, the results are shown in Table 4.

DESIGN PROCEDURE BASED ON f_2 USING SPIRAL TAPE PRIMARY

Let the requirement be for an 80Ω unbalanced to 640Ω balanced transformer, having an s.w.r. not greater than 1·25 at and between 10 and 100Mc/s.

Using standard filter equations:

$$L_L = \frac{Z_o \,.\, 0·6 \,.\, 10^6}{\pi f_2} = 0·15\mu\text{H and } n^2C_s = C_p = \frac{0·3 \times 10^{12}}{\pi f_2 Z_o} = 12\text{pF}$$

$\therefore$ as $n^2 = 8$, $C_{ss} = 12$pF.

The equation $C_p = \frac{1}{2}(C_{wp} + C_{cp}) = 12$pF is satisfied if C_{wp} and C_{cp} are each made 12pF, from Graph B, $C_{wp} = 12$pF with 5 turns of 0·2in tape interleaved with ·003in material, and from Graph A C_{cp} is 12pF with 0·2in tape if a 0·012in former is used. Now if $n_p = 5$ and L_L is required to be 0·15μH, reference to Graph D shows that a thickness of 0·012in primary/secondary insulation is required. To make C_{ss} 12pF with this primary/secondary insulation, the winding length required is 0·17in (Graph C), and 14 turns of 32 gauge wire are needed.

The bottom frequency limit f_1 may be checked by applying the equation:

$$f_1 = \frac{Z_o}{2\pi \times 0·225 \times L_p} \text{ Mc/s. In this case } f_1 = 6·3\text{Mc/s.}$$

TABLE 5.

f (Mc/s)	INS. LOSS (dB)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.	BAL (dB)
1	1·0	1·02+j0·34	1·39	0·985+j0·392	1·48	>50
2	0·75	1·02+j0·16	1·18	1·0 +j0·181	1·2	>50
5	0·5	1·04+j0·12	1·13	1·015+j·056	1·054	>50
10	0·5	1·04+j0·08	1·08	1·071+j.015	1·03	>50
20	0·5	1·02+j0·041	1·03	1·07 +j·02	1·07	48
30	0·5	0·99+j0·03	1·02	1·1 +j·03	1·1	42
40	0·5	0·96+j0·08	1·09	1·16 +j0	1·16	41
50	0·5	0·9 +j0·05	1·12	1·16 +j·025	1·17	40
60	0·75	0·9 +j0	1·1	1·18 +j·121	1·22	38
70	1·0	0·9 +j0	1·1	1·14 +j·141	1·2	36
80	0·75	0·83+j0·08	1·21	1·08 +j·161	1·14	32
90	1·0	0·83+j0	1·2	1·03 +j·136	1·142	27
100	1·5	0·87+j0·1	1·18	0·975+j·101	1·12	26

TABLE 6.

f (Mc/s)	INS. LOSS (dB)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.
0·5	0·5	1·04 −j0·402	1·52	0·92 −j0·382	1·36
1·0	0·25	1·06 −j0·127	1·15	0·92 −j0·183	1·23
10·0	0·25	1·06 +j0·06	1·09	1·0 +j0·05	1·054
20·0	0·25	1·05 +j0·12	1·14	1·1 +j0·06	1·12
30·0	0·5	1·03 +j0·17	1·18	1·25 −j0·07	1·15
40·0	0·75	1·03 +j0·32	1·37	1·24 −j0·32	1·42
50·0	1·0	0·994+j0·3	1·38	1·025−j0·49	1·61
60·0	1·5	0·96 +j0·42	1·52	0·77 −j0·49	1·83

The complete winding specification is:

Core:— Ferroxcube type LA1 (ring piece ground so that $n_p = (L_p)^{\frac{1}{2}}$.

Former:— 8 turns ·0015 × 0·375in polystyrene tape.

Primary:— 5 turns 0·2in × ·001in copper tape, interleaved with ·003in × 0·375in polystyrene tape.

Primary/secondary insulation:— 8 turns ·0015in × 0·375in polystyrene tape.

Secondary:— 14 turns 32 gauge enamelled wire.

Such a transformer was wound, and on measurement was found to have an s.w.r. of 1·42 at 100Mc/s. This was caused by too large a capacitive component on the secondary side and the transformer was rewound using 37 gauge wire with the results given in Table 5.

Encapsulation

A convenient finish for this type of transformer particular when required for outdoor use with aerial systems, is encapsulation. Typical examples are shown in Fig. 9 where Araldite casting resin has been used. Small quantity production of the transformer detailed in example (1) has been undertaken and consistent results obtained after encapsulation. The measurements shown in Table 6 were taken using one of these production transformers after encapsulation. The balance was not affected.

Production Cost

Complete transformers encapsulated as shown in Fig. 9, and suitable for use as wide band matching units for aerial systems, have been produced for under £3. The cost of the single plastic mould used was 5s., a multiple mould would be required for quantity production.

Conclusion

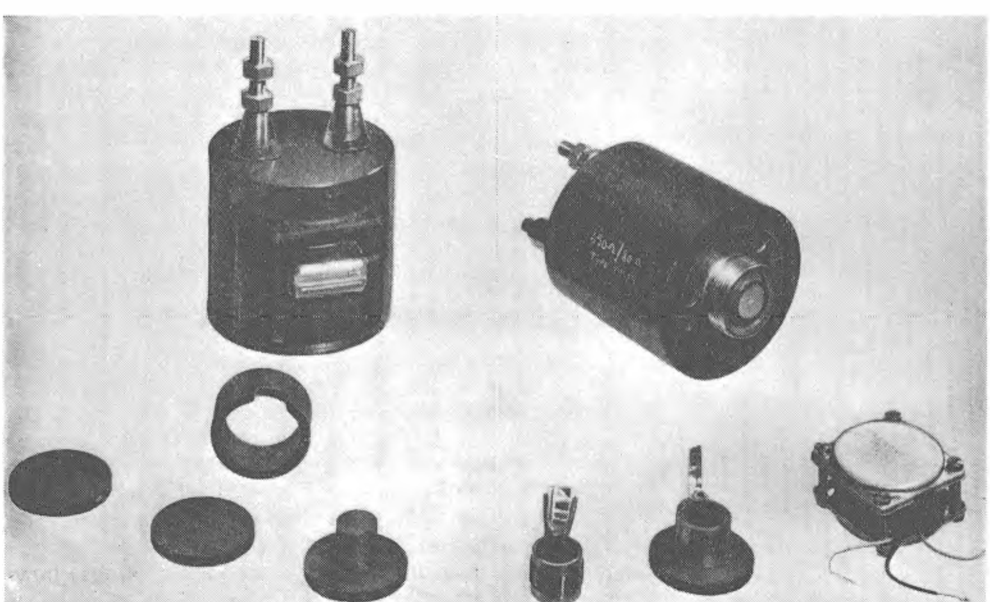
The article demonstrates that a wide band transformer which complies closely with a specified performance may be produced at low cost using this design method.

Some of the practical limitations imposed by frequency and impedance are indicated, but it should be remembered that these correspond to a required impedance match of s.w.r. = 1·25. If this restriction is relaxed f_2 will be increased by a factor determined by the equation given.

For aerial matching purposes the frequency range of a transformer is normally specified between 2dB points, using this criterion the frequency range of the transformer described in example (1) is 0·15 to 80Mc/s and that of example (3) 0·3 to 170Mc/s.

It is recognized that the performances obtained so far are by no means optimum. Further work will be devoted to extending the range of the transformers, one possible approach being to arrange the transformer elements to conform to an M of MM' configuration so that a greater portion of the pass-band of the equivalent filter may be utilized; another is to reduce all physical dimensions. By these means it seems likely that the upper frequency limit for small impedances and ratios may be extended to at least 500Mc/s.

Fig. 9. Component parts and completed transformer, encapsulated in Araldite, for use in aerial feeder systems



Acknowledgments

The authors wish to thank Mr. L. G. McAllister for his work in the early stages and are indebted to the Director, Government Communications Headquarters for permission to publish this article.

A Miniature Unit for Controlling the Duration of Radioactivity Measurements

By M. C. B. Russell*, B.Sc., A.M.I.E.E., and J. C. Boag*

A unit is described which may be associated with scaling circuits to control the duration of radioactivity measurements. The duration of the count is measured with a maximum error of 20msec in units of time which are displayed on a register. A count is automatically terminated by preset count or time limits, but its duration is always made equal to a whole number of time periods. The circuit is designed for inclusion in a scaler of conventional design[1].

DUE to the random nature of radioactive disintegration, any measurement of activity (expressed in disintegrations per minute) has a standard error equal to $\sqrt{n}$, where n is the total number of events counted. In any experiment, the choice of counting time is a compromise between what is short enough for the activity or other experimental factors to remain sufficiently constant, and long enough to count the number of nuclear events n required to make the percentage error, $100/\sqrt{n}$ sufficiently small. Normally a scaler is used to count the events and the operator, using a stop-watch, controls the count duration with a start and stop key.

The required counting time may range from a few seconds to several days. When the duration is short, timing errors arise due to stop-watch reading errors. When it exceeds a few minutes much time is spent by the experimenter waiting to terminate the count. For both reasons, various automatic methods have been used to control the counting and store the count and its duration, from which the activity may be calculated. (The count is defined as the number of nuclear events which are detected by a discriminator associated with the radiation detector, and is some fraction of the number of nuclear disintegrations which have occurred.)

Automatic Scalers

Various circuits and mechanisms have been developed for controlling the duration of a count and storing or recording the measured count rate[2].

In many applications some automatic control of count duration is required to ensure the accuracy of the result. If the half-life of a nuclide is comparable with the required duration of count it is important not to prolong the measurement. The required accuracy may be obtained if the count is terminated manually, but this is a waste of skilled time. The present instrument has been designed to meet the need for automatic control when a single activity measurement is required, and the user returns later to note the result and start another count whenever necessary.

In this unit, two alternative methods of automatic control may be used. These are respectively time control, in which the count is automatically terminated after a controlled length of time; or count control, in which the count continues until a predetermined number of events has been counted. When both controls are used together, counting continues until either the count limit or time limit is reached, whichever occurs first.

Time Source

A convenient way of measuring the duration of the counting period is to count time impulses of known periodicity. It is best to make the period of the time source long, so that only a few time impulses have to be counted. In general this results in circuit economy in any autoscaler, since it either reduces the wear on mechanical parts or reduces the required storage capacity of time scaling circuits. For the circuit which is described below, the period should not be made too long since the scaler does not commence counting until the first time-pulse occurs after the count key is depressed, and there is no visual indication of count rate until then. This waiting time would cause annoyance if it were more than half a minute, so means must be provided for counting at least the number of half-minutes in the longest duration of count required.

To simplify the design of the present unit it has been assumed that an installed time source is available with several alternative time periods, so that an adequate range of count duration can be obtained by counting between one and one hundred time impulses.

Various time sources are suitable. An inexpensive and reliable source is the G.P.O. electrically maintained pendulum clock which provides time impulses on 3 circuits at intervals of 1, 6 and 30secs. These impulse rates are slow enough to permit continuous drive of electromechanical relays for a long time without excessive wear and therefore find many applications in laboratories and plant at Harwell.

An alternative time source may be made, using a tuning fork oscillator of about 400c/s with a frequency dividing circuit. Cold-cathode scale-of-ten counting tubes or ferrite scaling circuits are particularly suitable for this purpose.

Time Control Register

The time impulses are repeated by a relay and are counted on a 4-digit electromechanical register which has a simple and reliable manual reset to zero. (Sodeco type TCeZ4Ec10c100 with 7 watt coil). This register has two pairs of digit-drum operated contacts, which in the present instance are arranged to close during the arrival of the tenth or hundredth impulse respectively. Either pair of contacts or those on the time-pulse relay may be used to obtain time limits of one, 10 or 100 time pulse periods, or the time limit may be switched off, and the register then reads the time at which the counting period ends.

Description of Circuit

The circuit, shown schematically in Fig. 1, has been designed to work in conjunction with a scaler similar to that described by Florida and Williamson[1] which displays the total count in decimal form up to 5 decades. A recent version of this scaler (AERE Type 1221C) has provision for extracting the pulse from the zero cathode of any of the 5 scaling tubes. The preset count limit is reached when the glow discharge arrives at the zero cathode of the chosen decade tube in the scaler; counting continues however until the arrival of the next time pulse.

* *A.E.R.E. Harwell.*

Relay TP repeats the time impulses to which the unit is connected. Relays A and B are arranged in a scale-of-two circuit of the type described by Barnes[3] and are used to detect the end of those time impulses at which the counting period begins and ends. Relay B is operated for the duration of the counting period and one of its contacts switches the scaler to the 'start' condition while another connects time impulses to the register SR. Valve V_1 is triggered when the count limit or time limit is reached.

is derived from a tens or hundreds contact on the time register it will trigger V_1 which operates CR before the release of the time impulse relay TP.

The first time impulse which is present after the change-over of CR_1 releases relay A by connexion of the reversed A winding via B_1, TP_1, B_2 and CR_1 to earth. One winding of relay B remains energized through A_1 and TP_1, but when the time impulse ends B releases, stops the scaler and disconnects the time register.

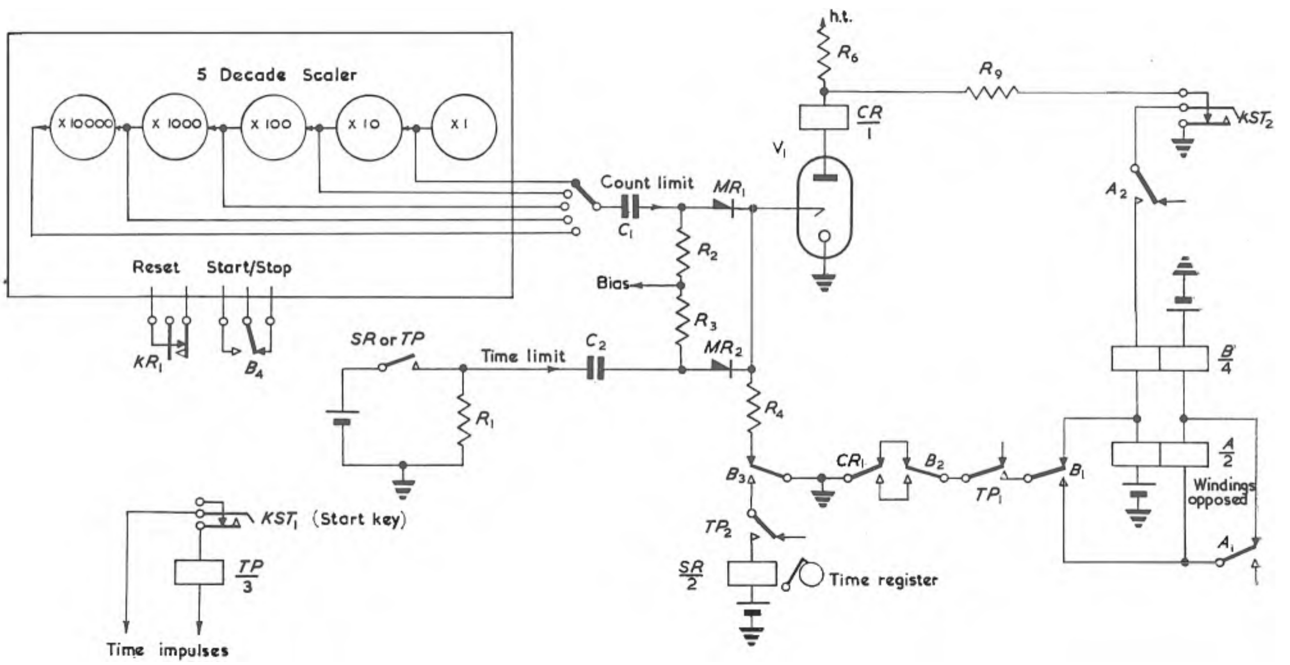


Fig. 1. Simplified diagram of the timing unit

Current through V_1 energizes relay CR which prepares for relays A and B to terminate the counting period at the end of the next time-pulse. To describe the operation of the circuit, a complete cycle will be described, commencing when the control key is thrown to reset and then to the start position.

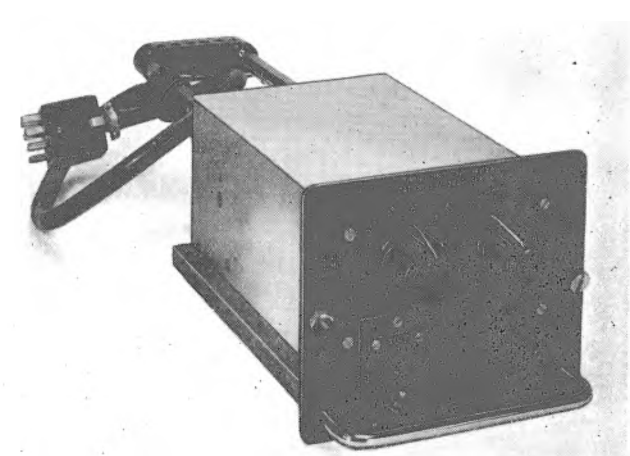
Two keys KR (reset) and KST (start) are combined into one control key. Upward movement operates KR which is non-locking and downward movement operates KST which is locking. Operation of KR resets the scaler through KR_1 and because KST is normal while KR is operated, relays A and B release. Contact B_4 ensures that the scaler will be in the 'stop' state after resetting. The anode of the cold-cathode valve V_1 is held below its maintaining potential by R_6 and R_9 through KST_2, so CR is released. The time register SR is now reset manually.

When KST is thrown to the start position, the arrival of the first time impulse in TP operates A through contacts CR_1, B_2, TP_1 and B_1. When this impulse ends, B operates through KST_2 and A_2, and contact B_4 switches on the scaler. The next, and subsequent time impulses are now registered on SR through B_3 and TP_2. This condition remains until arrival of a count or time limit signal triggers V_1. The count limit signal is a 40V positive pulse obtained from the output of any one of the scaler decades. The time limit signal is obtained from the time pulse relay TP or from contacts on the time register SR. These signals are applied to the trigger electrode of V_1 via capacitors C_1 and C_2, and through rectifiers MR_1 and MR_2 which serve to isolate the two channels and thus increase the input impedance of each. Either or both of the time and count limit signals may be connected according to the control function which is required. If the time limit

Limits of Time and Count Control

The count limit is derived from any one of the 5 scaling tubes and may therefore be 10, 10^2, 10^3, 10^4 or 10^5 of the scaler input pulses (the limits 10 or 10^2 are required when a fast scaler precedes the Dekatron scaler). The time limits are either one, ten or one hundred periods of the time source. If a G.P.O. pendulum is used with three time inputs of 1, 6 or 30sec period, the following limits may be selected; 1, 6, 10, 30 and (100) seconds and 1, 5, 10 and 50 minutes. In the unit shown in Fig. 2 the 100 second limit has been deleted to economize in the number of switch positions.

Fig. 2. The timing unit



Accuracy of Scaler Control

The scaler commences to count when the release of relay *TP* allows relay *B* to be energized via relay *A* and KST_2. The count ends when the release of relay *TP* de-energizes relay *B*. The time of count is therefore a whole number of the intervals between the trailing edges of the time impulses, plus an error due to the difference in the operate and release lags of relay *B*. The error, which is independent of the duration of the time impulses, measures less than 20msec when suitable G.P.O. type 3000 relays are used. The variation in the operate and release lags of relay *B* and in the release lag of *TP* has been measured for one particular timing unit with the apparatus shown in Fig. 3.

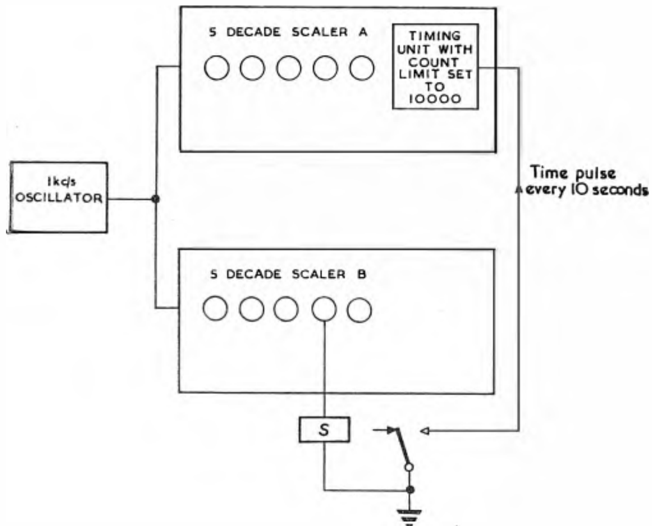


Fig. 3. Apparatus used to test the timing unit accuracy

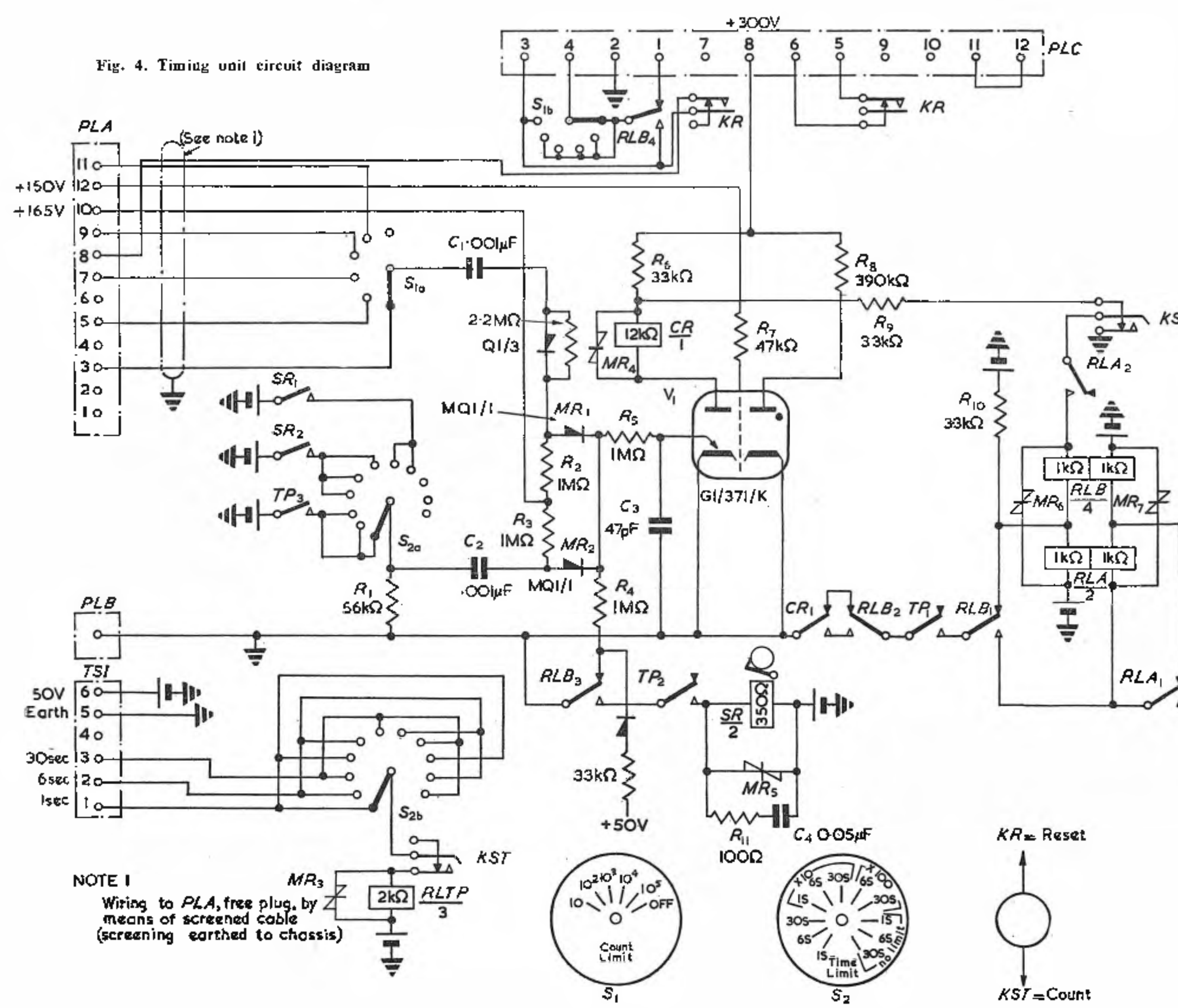
Scaler *A* is a standard scaler type 1221C which includes the timing unit. This scaler counts regular impulses at about 1 000c/s from an oscillator. The same impulses are counted on an identical scaler *B*, in the fourth decade of which a high-speed sensitive relay *S* (Carpenter type 3H47) operates on the receipt of every ten-thousandth impulse, and releases one thousand impulses later. A contact on the relay therefore closes every 10sec for 1sec, and this is used as a time source which is in synchronism with the pulses counted by scaler *A*.

The average of 50 readings of the number of oscillations counted during 20sec measures 19985·7 with a standard deviation of $\pm 1{\cdot}2$ counts. This result corresponds to a timing error of about -14msec with a standard deviation of 1·2msec. These results should not be regarded as more than an indication of the order of accuracy, since the error will change according to the adjustment of relay *B*.

Accuracy of Clocks

In practice variations may be expected to occur in the interval between time impulses if they are derived

Fig. 4. Timing unit circuit diagram



mechanically. The G.P.O. pendulum clock (Type 36) has been found to be surprisingly accurate in this respect. The pendulum has a two second period, and the one second circuit is obtained from two springsets in parallel, operated at either end of its swing. There is liable to be a difference of up to 50msec between the spacing of alternate pairs of 1sec impulses, depending on adjustment. For a typical clock the variation is found to be reduced to $\pm 1{\cdot}7$ or $\pm 2{\cdot}9$msec, if the duration of count is an even number of seconds, depending which springset is operative. When the 6sec and 30sec clock circuits were examined, their variation was found to be hardly more than that of the 2sec periods, being measured as $\pm 3{\cdot}1$ and $\pm 3{\cdot}9$msec respectively (in every case these figures include variation in oscillator frequency and in the release lag of the relays which are used to repeat the clock impulses to several circuits in the building).

Detailed Circuit

A detailed circuit of a timing unit, designed to use the power supplies and waveforms in scaler type 1221C, is shown in Fig. 4.

This circuit is based upon the schematic shown in Fig. 1 but has two extra facilities as follows:

(a) A 'no limit' position of switch S_2 in which the time of count is recorded however long it takes to reach the count limit. No corresponding position of the count limit switch is provided since it is considered that a count of 10^5 is sufficient, and an extra count register would be required to store more digits. It should be remembered that when a count limit of 10^5 is chosen on a scaler with only 5 decade presentation, all digits start again from zero. Since counting continues till the end of the next time impulse a small number is displayed to which 10^5 must be added.

(b) A position of the count limit switch, labelled 'off,' which converts the scaler to its normal function, exactly as it would be if the timing unit were disconnected. The start and reset keys of the scaler may then be operated in the normal way.

Construction

The unit has been constructed as a small plug-in sub-

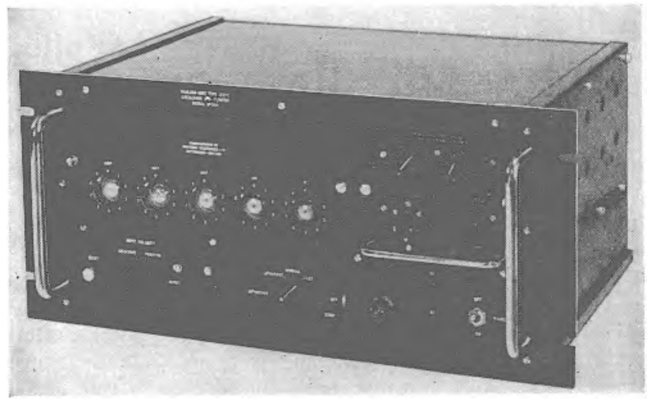


Fig. 5. The scaler complete with timing unit

chassis for scaler 1221C as shown in Fig. 2. The unit has a terminal strip which is accessible at the rear of the scaler, and also a 12-way lead terminating in a miniature plug which is put into a socket on the rear of the scaler. Connexions to three alternative time impulses and to a 50V d.c. supply are made via the terminal strip. The unit is shown fitted into the scaler in Fig. 5.

Conclusion

Clock pulses are available in many of the laboratories in which radioactivity measurements are made, and a timing unit has been designed to make use of these. The circuit is designed for accurate time and count control of a standard scaler, and the components are small enough to be included in the scaler as a sub-chassis.

Acknowledgments

The authors' thanks are due to R. C. M. Barnes for helpful criticism of this article, and to R. L. Jacob who designed the layout of the unit shown in Fig. 2.

REFERENCES

1. Florida, C. D., Williamson, R. A Cold-cathode Scaling Unit. *Electronic Engng. 26*, 186 (1954).
2. Cooke-Yarborough, E. H. The Counting of Random Pulses. *J. Brit. Inst. Radio Engrs. 11*, 367 (1951).
3. Barnes, R. C. M. Relay Scale-of-two Circuits. *Electronic Engng 26*, 432 (1954).

The Second Trans-Atlantic Telephone Cable

The manufacture of over 2 200 nautical miles of the submarine telephone cable required for the second trans-Atlantic project for laying between America and France by the American Telephone and Telegraph Co. Ltd., has been entrusted to Submarine Cables Limited, owned jointly by Siemens Edison Swan Ltd. and The Telegraph Construction and Maintenance Company Ltd. The sea route distance between Newfoundland and France is approximately 2 100 nautical miles and manufacture of the balance of the cable will be shared by the Simplex Wire and Cable Co. of America, Cables de Lyon of France and Felten and Guilleaume of Germany.

As in the case of the first project, the second will involve two cables of the same type for the main crossing, one for each direction of transmission. This part of the system will again provide 36 telephone circuits and will be equipped with flexible one way repeaters supplied by the Bell Telephone Laboratories of New York. The single cable for the Newfoundland-Novia Scotia link will use British Post Office rigid type repeaters designed for 60 two-way circuit operation. The internal units will be manufactured by Standard Telephone and Cables Ltd. while the steel casings will be supplied by Submarine Cables Ltd.

The design of the cable is substantially the same as for the first trans-Atlantic cable. The inner core consists of polythene extruded over a central copper conductor composed of a 0·131in wire surrounded by three tapes 0·0145in thick. Over this core, which is 0·62in diameter, is applied the outer conductor consisting of a layer of 6 copper tapes applied with a long lay and held in position by a copper binder tape and a protective fabric tape, both applied with a short lay.

The jute serving, galvanized steel armour wires and the outer protective serving depend on the depth and sea bed conditions in the locality of laying. For the deep sea cable, a layer of 24 high tensile steel armour wires, each 0·086in diameter is used, each wire being separately protected from corrosion by a coating of compound and an impregnated fabric tape before the wires are applied to the cable. The overall diameter of the deep sea portion of the cable is 1·2in with a weight of 3·1 tons/n.m. in air and 1·9 tons/n.m. in water. The depth of water over practically the whole of the Newfoundland-Novia Scotia section and for a considerable distance out from each of the terminal points of the main crossing is relatively shallow and this necessitates much heavier protection. Over 600 n.m. of cable heavily armoured with 12 steel wires each 0·30in diameter is to be provided by Submarine Cables Ltd. This type of cable has a diameter of 2·7in and weighs 26 tons/n.m. in air.

The cables will be laid in the summer of 1959 and the system is expected to be open for service by the end of that year. The work of laying the cables will be shared by H.M.T.S. *Monarch* (8 000 tons) belonging to the British Post Office and c.s. *Ocean Layer* (4 500 tons) owned by Submarine Cables Ltd.

Applications of a

New Type of Cold Cathode Trigger Tube

By G. O. Crowther*, B.Sc., A.C.G.I., and K. F. Gimson*, B.Sc.Eng., A.M.I.E.E.

(Part 2)

THE number of potential applications for cold-cathode tubes is large and it is almost certain that the trigger tube could give improved performance in a number of functions at present using either thermionic valves or electro-mechanical devices. It is the aim of the present article to indicate the type of application for which the cold-cathode tube is suited and to illustrate the design procedure in a number of specific examples.

It is probable that one of the factors which has delayed a more widespread adoption of the trigger tube has been the rapid development of the transistor. There are a number of general properties common to both devices, namely negligible standby losses, instantaneous operation and long life. But, apart from these general properties the two devices are basically different; the trigger tube is an on-off device while the transistor is essentially a linear amplifier, although like the thermionic valve it can also be used in switching type circuits. In the switching and industrial fields it is felt that both the transistor and the trigger tube will each have applications for which it is best suited. There will also, of course, be those where the choice is difficult. To try and clarify the position Table 1 which compares the primary parameters of the two devices has been prepared. In Table 2 three important properties of trigger tubes, such as the Z803U, have been given together with the type of application in which each is the important feature.

The trigger tube is often regarded as a sensitive relay and for a number of relay and high input impedance applications this is a useful analogy. The stable trigger tube can also be regarded as a voltage reference plus an amplifier; the trigger breakdown voltage representing the reference voltage. A complete voltage stabilizer using a single trigger tube has already been described and illustrates the principle. Although this circuit will not be discussed in the present article the idea of using the trigger tube as a reference voltage plus amplifier is used in protection circuits and a voltage regulator circuit to be described.

* *Mullard Ltd.*

Table 1

	TRANSISTORS	TRIGGER TUBES
Frequency range ..	1c/s—1Mc/s*	1 cycle in 10 min. —10kc/s*
Input impedance ..	10Ω—1kΩ	10^9—10^{11}Ω
Current gain ..	30—100	200—10^5 dependent on the circuit
Operating time ..	<1μsec	20μsec—1msec
Size	Smaller than submin.	Smallest is submin.
Max. output current	50mA—2A	5–40mA
Output impedance	—1.5MΩ	100—1000Ω
Ability to withstand electrical overloads	Poor	V. good
Mechanical strength	V. good	Good
Operating voltages	10—50V	100—400V
Change of characteristics with temperature	High	Small
Visual indication ..	No	Yes

* *These limits are partially set by the associated circuit.*

Table 2

Stable Ignition voltage	Timers Overvoltage alarm Voltage control Stable sawtooth generators Analogue to digital conversion Small signal "relay" operation
Negligible standby losses, instantaneous operation, long life	Protection equipment Battery operated standby equipment Large electronic equipment where heat dissipation is a problem
High input impedance	Touch control and detection Liquid level control etc. Detector of resistor change (e.g. level control) Photocell applications

Basic Circuits

TRIGGER CIRCUIT

In this section the basic methods of triggering and extinguishing a cold-cathode tube are discussed. Although this will be done with direct reference to the Z803U, the remarks will in general apply to all cold-cathode trigger tubes.

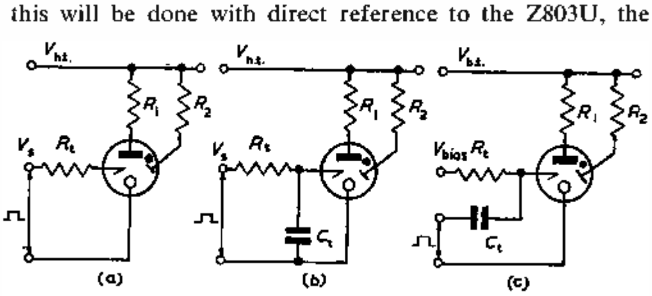


Fig. 15. Basic methods of triggering a tube

Fig. 15 shows the three basic methods of triggering the tube. Fig. 15(a) illustrates the current triggering method discussed earlier, the method is comparatively insensitive and unless care is taken unreliable operation may be obtained.

There are two conditions which must be met to ensure the tube is triggered:

$$V_s > V_{t\,(ign)} \quad \ldots \quad (1)$$

$$V_s > (R_t i_{tr}) + V_m \quad \ldots \quad (2)$$

where V_s is the input voltage.

i_{tr} is the transfer current for the anode voltage in use, taken from the transfer characteristic of Fig. 4.

R_t trigger circuit resistance

V_m is the tube maintaining voltage.

If $\quad R_t < \dfrac{V_{t\,(ign)} - V_m}{i_{tr}}$ .... (3)

(Approx. 300kΩ for Z803)

Then equation (2) is automatically satisfied when equation (1) is met. Thus the tube fires at a precise voltage, which is essential in applications requiring stability.

If equation (3) is not satisfied, the input firing voltage to cover tube spreads must be greater than $V_{t\,(ign)}$ by an amount

$$(R_t i_{tr}) - (V_{t\,(ign)} - V_m)$$

(See ref. 5)

It should be remembered however that while the above input voltage will fire all tubes over life a lower voltage, greater than $V_{t\,(ign)}$, may still fire some tubes which have low individual values of i_{tr}. In addition the transfer characteristic of an individual trigger tube may not be constant over life, only the maximum quoted value represents an absolute safe limit.

The circuit of Fig. 15(b) was also briefly described in the introduction. It is considerably more sensitive than the previous circuit, but there is a small delay while the capacitor C_t is charged from its initial value to the breakdown voltage.

When the capacitor C_t is discharged from $V_{t\,(ign)}$ to V_m the energy passed through the trigger gap is sufficient to ensure anode-cathode conduction and for this reason the trigger resistance can have a value many times greater than in the previous circuit. It should be mentioned that resistor R_t may be replaced by a photocell ionization chamber or other such devices.

The minimum values of capacitor C_t for given anode voltage ranges are quoted in published data. If a value less than the quoted one is used the amount of energy fed into the tube may not be sufficient to give reliable firing.

The magnitude of the resistor R_t will be limited by either the external leakages from the trigger to other points in the circuit or by the trigger prestrike currents mentioned previously. A plot of the maximum value of R_t (neglecting leakages) as a function of the trigger voltage was given in Fig. 11. If R_t is made greater than the quoted value the capacitor will only be charged to some value V, less than $V_{t\,(ign)}$, at which the current i through the resistor equals the pre-strike current i_p, and the tube will then never fire.

i.e.: $i = (V_s - V)/R_t = i_p$ or $V = V_s - R_t i_p$.

In a number of cases, for instance in timers, a capacitor which is much larger than the minimum value will be required. In such cases a limiting resistor in series with the capacitor is essential, to prevent the discharge causing damage to the tube.

The capacitor method of triggering a tube is preferable in applications requiring accuracy and stability, because, provided that R_t is not too large, the input voltage need only just exceed $V_{t\,(ign)}$ in order to trigger the tube. For the Z803U and an accuracy of about 1 per cent the maximum value of R_t is given by

$$R_t = \frac{V_{t\,(ign)}}{100 \times i_{p\,(max)}} \text{ (e.g. 10 to 20 M}\Omega\text{)}$$

where $i_{p\,(max)}$ is the maximum value of the pre-strike current.

It should be noted that for tubes with barium coated cathodes the pre-strike current will depend entirely on the ambient lighting and no fixed maximum value for R_t can be quoted for this class of tube.

Finally the circuit of Fig. 15(c) shows another capacitor firing circuit. It is essentially a pulse firing circuit and is not suitable for use in applications in which the input signal is a slowly changing voltage. It has the advantage that the breakdown is initiated immediately the input signal is applied. The factors in the choice of the capacitor C_t are similar to the previous case. The restrictions on the value of R_t are less stringent than in the two previous cases. The input bias to the trigger will normally be set several volts below the breakdown voltage and from Fig. 10 it will be seen that the pre-ignition currents are then less by a factor of 10^2, than the value when V_t is $V_{t\,(ign)}$. For this reason the resistor R_t can be considerably larger than in the circuit of Fig. 15(b). However immediately after extinction, the potential of the trigger will be below V_m and will be returned exponentially, to the bias voltage, with a time-constant R_tC_t. Thus if R_t is made too large there will be a considerable delay before the circuit regains complete sensitivity.

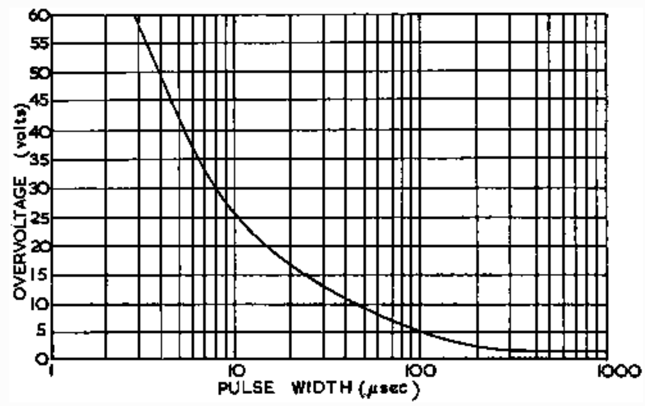


Fig. 16. Overvoltage on the trigger against pulse width (Z803U)

The firing pulse height required is dependent on the bias voltage, $V_{t\,(ign)}$ and the pulse width. For pulse's with a duration longer than 2msec the amplitude need only be $V_{(ign)} - V_{bias}$. If the pulse is shorter than 2msec a larger pulse is required and Fig. 16 shows a typical curve of the extra voltage required as a function of pulse width for the Z803U.

Because of the stability of the Z803U the bias voltage can be set just 2 or 3volts below $V_{t\,(ign)}$ and the pulse required is then small. However, if the anode current is large the bias voltage may have to be set to a lower value because of the hysteresis effect previously described.

The source impedance of the pulse should not exceed about 6kΩ since it is effectively in series with the capacitor and will limit the rate of discharge. If it is much greater a capacitor larger than the minimum value is then necessary and the circuit approaches that of Fig. 15(a).

In all three trigger circuits consideration has only been given to the maximum value of R_t, but not to the minimum value. The resistor R_t in all three circuits must be sufficient to limit the trigger current within the published limits.

$$\text{i.e. } R_t > \frac{V_m - V_{burn}}{i_{t\,(max)}}$$

A trigger tube can also be triggered by taking the cathode negative while holding the trigger at a constant potential. Cathode firing circuits in general are more complex since unless care is taken the signal source may have to supply the complete anode current. Two examples of cathode firing circuits will be described later.

Anode Circuit

The design of the anode circuit can normally be resolved into two parts, the selection of the anode load and the choice of the extinguishing circuit for both the anode-cathode and the trigger-cathode discharge.

The anode load is given by the expression

$$R_a = (V_{ht} - V_m)/i$$

where V_{HT} is the supply voltage and

i is the current which it is required to pass through the tube.

The value of R_a obtained is the total impedance in the loop and will of course include the internal impedance of the supply and any resistance between cathode and the negative line.

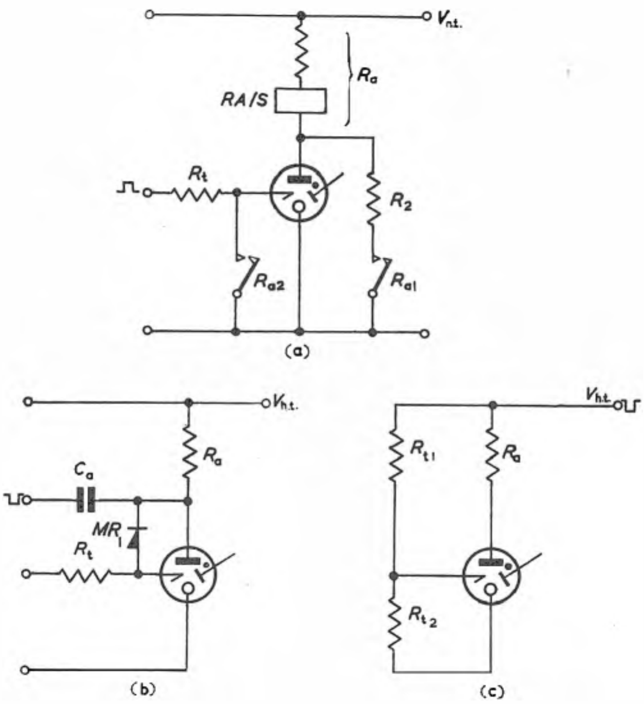


Fig. 17. Methods of extinguishing tubes

Three methods of extinguishing the anode discharge in conjunction with three methods of extinguishing the trigger discharge are shown in Fig. 17. It should be noted that other combinations of the anode and trigger extinguishing circuits shown can also be used.

Arrangement (a) is normally the method of extinction used in relay circuits. The tube only conducts momentarily, while the relay makes and closes contact R_{a1}, and for this reason the full peak current rating of the tube can be used.

R_2 is selected such that after the relay contact R_{a1} is closed the anode potential of the tube is at least 20 per cent below its normal maintaining value

$$\text{i.e. } R_2 = \frac{V_m\,(1\text{-}0{\cdot}2)}{V_{ht} - V_m\,(1\text{-}0{\cdot}2)}\,R_a$$

The 20 per cent margin below V_m allows for mains variations and component tolerances.

The current flowing when R_{a1} is closed can be quite large and is given by

$$i = \frac{V_{ht}}{R_a + R_2}$$

The trigger discharge is extinguished by shorting the trigger to cathode by the relay contact R_{a2}. If R_{a2} makes before R_{a1} the trigger will be strapped to the cathode while anode current is still flowing. Negative trigger current will thus flow momentarily and as already stated it is harmful to tubes such as the Z803U. Damage to the tube can be avoided by either ensuring that contact R_{a1} closes before contact R_{a2} or by inserting a limiting resistance in series with contact R_{a2}.

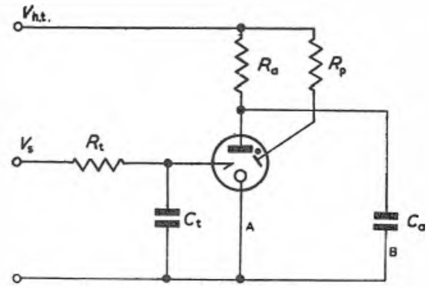
In Fig. 17(b) the tube is completely extinguished by a series of negative pulses fed capacitively to the anode. The negative pulse is also fed to the trigger via the diode MR_1 which is normally non-conducting.

The input pulse must be longer than the deionization time of the tube and the time-constant $C_a(R_a R_t)/(R_a + R_t)$ must be such that the pulse is not differentiated. Unless this is done, the positive spike which would occur at the termination of the pulse may cause anode cathode breakdown.

To avoid affecting the trigger circuit the reverse resistance of the diode must be much greater than the trigger circuit resistance R_t.

In Fig. 17(c) the h.t. supply voltage for both the anode and the trigger is lowered below V_m for a period longer

Fig. 18. Self extinguishing circuit



than the deionization time. It should be noted that if this method of extinguishing the trigger discharge is used with either of the other two methods of extinguishing the anode discharge, the resistor R_{t1} should be taken to the anode instead of the h.t. line.

Self Extinguishing Circuits

In all the previous circuits the tube has been extinguished by some external mechanism. In the circuits to be described now the tube is extinguished automatically. These circuits are of great importance in a wide range of application, perhaps the simplest of which is the relaxation oscillator.

A typical self extinguishing circuit is shown in Fig. 18. When the tube is triggered it will partially discharge capacitor C_a from the initial voltage V_{ht} to a potential below the normal glow discharge maintaining voltage. The amount the capacitor is discharged below V_m is dependent on the characteristics of the tube, the size of the capacitor, and the magnitude of any resistance in the capacitor discharge circuit. Provided R_a is sufficiently large and the time-constant C_aR_a is greater than the deionization time, the tube is extinguished and the capacitor recharged to h.t. potential via R_a.

The important feature in the operation of the circuit is the fact that the capacitor C_a is discharged below the maintaining voltage of the tube and not as might be expected to the maintaining voltage. To obtain an understanding of how the voltage can fall below the maintaining voltage the gas conduction process must be briefly analysed.

The period of conduction is very short in these capacitor discharge circuits and it is probable that there is not sufficient time for the conditions of a normal glow discharge to be completely established. Initially the current flow in the tube is mainly due to the movement of the highly mobile electrons, although a small percentage must be due to the slower positive ions. Consider the movement of one electron from the cathode; at the first collision with a gas molecule a further electron and a positive ion are liberated. There are now two electrons and at the next pair of collisions each will produce a further electron and a positive ion making four electrons and three positive ions. This process will lead to a large number of electrons arriving at the anode, leaving a cloud of positive ions near the anode, which drift slowly towards the cathode.

The process will continue until the capacitor C_a is discharged to about V_m when further ionization will decrease to zero. The drift of positive ions towards the cathode will however continue and will represent further current flow, i_p, through the tube. Provided that the recharge current of the capacitor through resistor R_a ($i_R = (V_{ht} - V_m)/R_a$) is less than the current i_p the capacitor C_a will be discharged further. For this reason there is a minimum value of R_a which for most trigger tubes is approximately 700kΩ to 1MΩ.

Fig. 19 shows the minimum value of R_a plotted against C_a

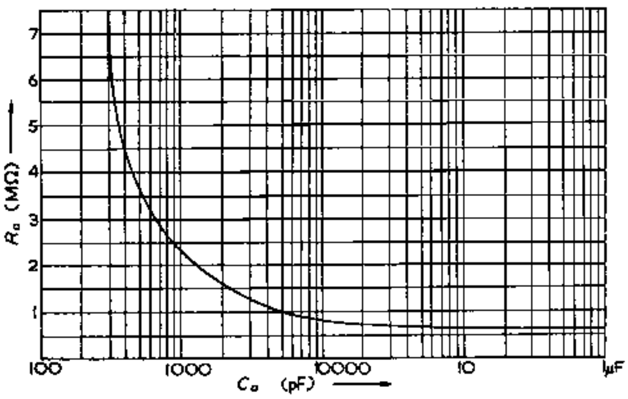


Fig. 19. Minimum value of anode resistance for self extinction as a function of the anode capacitor (Z803U)

for the Z803U. For small values of C_a the minimum value of R_a is set by the recovery time requirements whereas at values of C_a greater than about 0·01μF the minimum value of R_a is constant (680kΩ) and independent of C_a. Under these conditions the limiting factor is the recharge current of the capacitor C_a.

From what has been said it will be realized that the circuit relies for its operation on a very short discharge. It is, however, often desirable to insert a resistance in series with the capacitor discharge circuit either at a point A, to obtain a positive output pulse, or at point B, for a negative pulse. But if the resistance is made too large, the period of the discharge will be increased and affect the self-extinguishing properties of the circuit. There is, therefore, a maximum value to the limiting resistor.

It should be noted that when very large values of C_a are used a resistor is necessary to prevent damage to the tube.

So far only the use of self extinguishing components in the anode-cathode circuit have been discussed, but similar considerations arise if self extinction of the trigger is desired. The trigger circuit can be made self extinguishing whether or not the anode is self extinguishing.

It will be seen now that there is an additional advantage for the two capacitor firing circuits of Fig. 15(b) and (c); since, provided R_t exceeds a given minimum value (1MΩ for Z803U), the trigger circuit is self extinguishing. This is an important feature in simplifying the design of circuits.

Finally, to summarize, there are four main combinations of anode and trigger extinction systems.

(a) Trigger directly ignited, externally extinguished, anode externally extinguished.

(b) Trigger directly ignited externally extinguished, anode self-extinguished.

(c) Trigger capacitor ignited, self extinguished, anode externally extinguished.

(d) Trigger capacitor ignited, self extinguished, anode self extinguished.

With (d) it should be remembered that the anode must have regained its normal operating voltage before the trigger is re-ignited.

Timing Circuits

There are a number of instances where a simple timer is required such as in process control, welder control, valve heating delays, photographic work, etc. The RC timer using the cold-cathode trigger tube is simple, and can give an electrical output either via a pair of relay contacts or direct when a voltage is required for operating other electronic circuits. Very simple units may be designed which can operate direct from the normal mains voltages met in this country. The long term stability (one year) of a

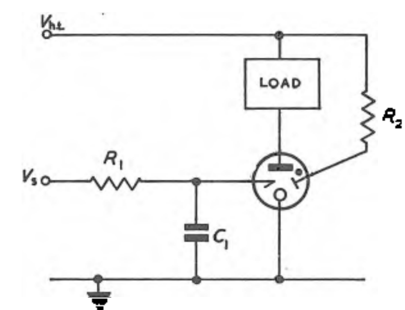


Fig. 20. The basic timer circuit

typical timer using Z803U is of the order of 2 or 3 per cent; the figure includes both tube and component variations. A short term stability of better than 0·1 per cent over an eight-hour period is obtainable.

The maximum timing period which can be obtained is limited by the tube to approximately three hours although for reasons of economy, the practical limit with RC circuits lies within the range one to ten minutes. For times greater than this it is probable that mechanical methods are more economical. The minimum timing period depends on the nature of the output circuit. If a relay output is used, then the minimum time is in general restricted by the pull-in time of the relay and may be in the region of 1 to 2sec. However, if a voltage output only is required the minimum time is of the order of 10msec and is set by the ionization time of the trigger-cathode gap.

A basic trigger tube timer is shown in Fig. 20. The capacitor C_1 is charged by way of the resistor R_1 and its potential rises exponentially from zero towards the input potential V_s until it is charged to the trigger breakdown voltage $V_{t(ign)}$.

The tube fires when the potential of C_1 reaches the value $V_{t(ign)}$. The load may consist of either a relay which pulls in when the tube fires, making or breaking contacts in other circuits, or a pure resistance across which a negative pulse is developed suitable for triggering other electronic equipment. A positive voltage can be obtained by inserting the resistance in the cathode.

The timing period achieved is given by the expression

$$t = C_1R_1 \log \frac{V_s}{V_s - V_{t(ign)}} \qquad (4)$$

where V_s = the trigger supply voltage.

When considering the accuracy and stability of timers it is clear from expression (4) that attention must be paid not only to the breakdown voltage $V_{t(ign)}$ but also to the trigger supply voltage V_s and the components C_1 and R_1.

It will be seen that variations in either R_1 or C_1 will cause a proportionate change in the time t, while changes in either V_s or $V_{t(ign)}$ have a greater effect. The following expressions relate the changes in either V_s or $V_{t(ign)}$ to the change in the time t. (See refs. 6 and 7.)

$$\Delta t/t = \beta\,(\Delta V_{t(ign)}/V_{t(ign)}) \quad \ldots\ldots\ldots\ (5)$$

$$\Delta t/t = -\beta\,(\Delta V_s/V_s) \quad \ldots\ldots\ldots\ (6)$$

where

$$\beta = \frac{\alpha}{(1-\alpha)\log[(1/(1-\alpha)]} \quad \ldots\ldots\ldots\ (7)$$

and $\alpha = (V_{t(ign)}/V_s)$

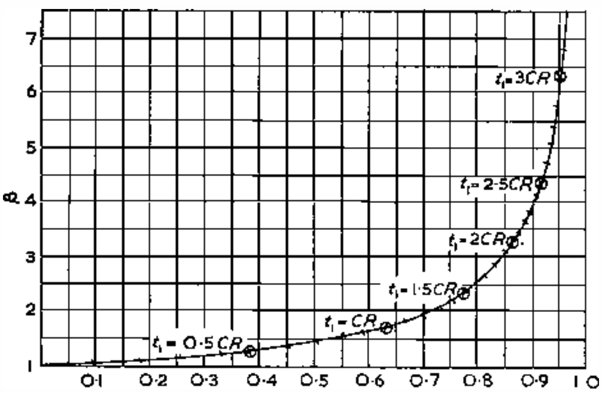


Fig. 21. Error multiplying factor and timing in exponential charging circuit

β is shown plotted as a function of $V_{t(ign)}/V_s$ in Fig. 21 and it will be seen that it is always greater than one, but approaches one as $V_{t(ign)}/V_s$ tends to zero.

In general a value of $V_{t(ign)}/V_s$ between 0·4 and 0·8 is used. Below the lower value V_s becomes impractically large for a very small improvement in performance. Above a ratio of 0·8 the value of β rises rapidly and small changes in either V_s or $V_{t(ign)}$ cause very large changes in timing. For convenience, the time t, as a multiple of the time-constant CR, has been marked off along the curve. Thus once the ratio $V_{t(ign)}/V_s$ has been selected the multiple of CR can be read directly from the curve, e.g. for $(V_{t(ign)}/V_s) = 0{\cdot}4$ $t = 0{\cdot}5 \times R_1C_1$.

In some applications it may be desirable to charge the capacitor C_1 from a fixed voltage V_o instead of from zero. The expressions (5), (6) and (7) still apply except that α becomes $(V_{t(ign)} - V_o)/(V_s - V_o)$. The curve for β is still the same provided the new value for α is used.

So far, consideration has only been given to the stability of a timer during life and not to the initial accuracy. For this the tube to tube tolerances of $V_{t(ign)}$ must be taken into account together with the initial tolerance on V_s which is normally derived from one or more stabilizer tubes. The simplest method of correcting for tube to tube tolerances is to adjust V_s by means of a potentiometer across the stabilizer so that the ratio $V_{t(ign)}/V_s$ is constant. Once this has been done the value of the product R_1C_1 is fixed. These components must therefore be selected to the required initial accuracy and for their stability. In addition, attention should be paid to the effect of temperature changes both permanent and reversible on these components. Further the insulation resistance of the capacitor C_1 must be much greater than R_1. It is normal to use either good quality paper or plastic film capacitors for C_1.

Maximum and minimum times for a timer of this type can now be considered. In relay operating systems the minimum time, as has been stated, is limited by the pull-in time of the relay and for accurate timing the period should be made between 50 and 100 times greater than the maximum variation in pull-in time of the relay.

In applications where a voltage output only is required the minimum time is set by variations in the ionization time. Fig. 22 shows fluctuations of the ionization time as a function of the timing period. For 1 per cent stability, other

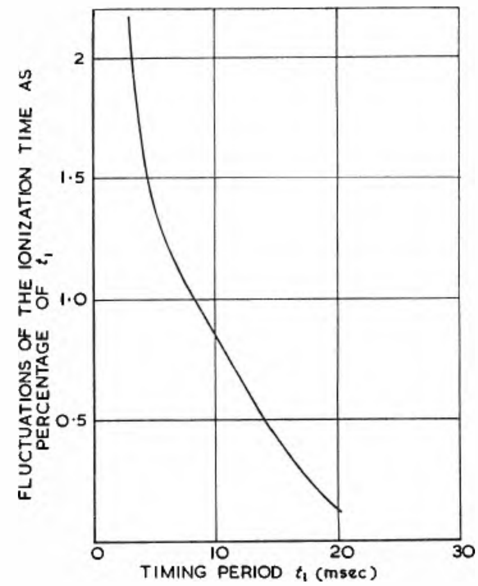


Fig. 22. Fluctuations of the ionization time as a percentage of the timing period t_1 against t_1

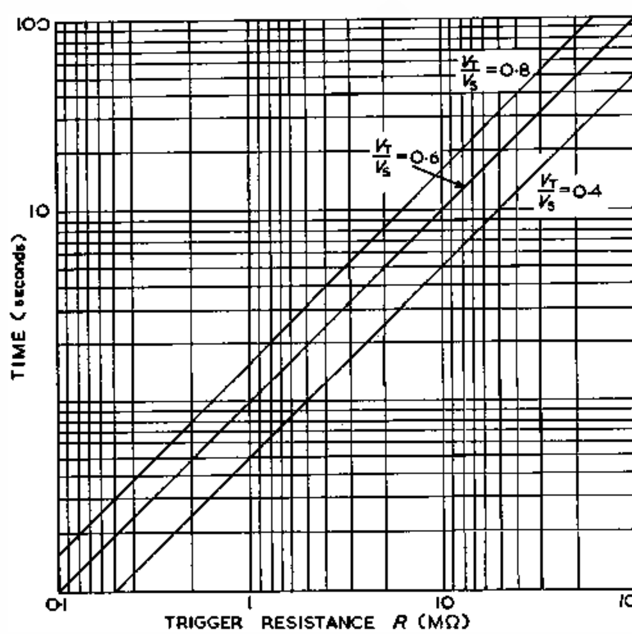


Fig. 23. Timing period t as function of the trigger resistance For V_T/V_S = 0·8, 0·6, 0·4

factors being ignored, the minimum timing period is 10msec.

The considerations for maximum timing interval are more complex as costs, insulation and tube effects must be taken into account. The timing period can be made as large as is desired provided the trigger capacitor C_1 is made sufficiently large. However, a maximum value for C_1 is usually set between 1 and 10μF by consideration of cost and physical dimensions.

Once the value of C_1 is settled the maximum timing period, neglecting for the present the question of economics, is set by the trigger pre-strike current. The maximum timing period is obtained for a given value of capacitor when it is charged with a constant current just in excess of the pre-strike current. For Z803U using a 1μF capacitor it is three hours. The errors in timing due to the pre-strike current are very small and are less than 0·5 per cent for the maximum timing conditions.

It is doubtful if a trigger tube circuit for these long times would be an economic proposition due to the cost of high value resistors, the high insulation for the trigger circuit and the stable trigger supply voltages.

Fig. 23 shows the timing period as a function of the charging resistor R_1, for a number of supply voltages, when the charging capacitor C_1 is 1μF.

Examples of complete timer circuits are shown in Figs. 24 and 25. The circuit shown in Fig. 24 meets the requirement for a general purpose timer giving a typical long term stability of about 2 per cent. A simpler design, shown in Fig. 25 is suitable for these applications where a stability of between 5 and 10 per cent is required, as in a number of industrial process controls, motor starting sequence timers, and timers to give heating up times for large electronic tubes.

The timer of Fig. 24 has been designed to have an initial accuracy of better than $\pm$5 per cent and to give the following ranges:

Resistance selector switch position	Timing period (in seconds) when charging capacitance is: (a) 0·1μF $\pm$2%	(b) 1·0μF $\pm$2%
1	1·0	10
2	1·5	15
3	2·1	21
4	3·1	31
5	4·8	48
6	7·0	70

The timing is initiated by momentarily breaking the Switch S_1 and thus de-energizing the relay A/N. The contact A_2 will then be opened and the timing capacitor (C_1 or C_2) can be charged. When the relay A/N is made, at the end of the timing period, the tube only conducts momentarily since the relay is self-locking on contact A_1. Contact A_2, which should close after A_1, resets the trigger capacitor to zero potential. Other contacts of the relay are employed in the external circuits.

The trigger supply is obtained from two stabilizer tubes (type 90C1) running at approximately 4mA. The potentiometer R_5 is used to compensate for tube to tube tolerances of both the stabilizers and the trigger tube. The circuit has been designed for a ratio of $V_{t(ign)}/V_s$ of 0·83, where V_s is measured from point A to earth.

The arrangement of the timing resistors (R_7 to R_{12}) and the means of range switching employed are slightly more economical than other methods, because half of the resistors in this arrangement may have wider tolerances and yet give approximately the same accuracy.

For a single range timer it may be desired to compensate for tolerances in the timing resistor and capacitor with R_5 as well as for tube to tube tolerances. However, for the same performance a value of V_s greater than can be obtained from two 90C1's would be required.

The timer operates directly from the mains and covers the complete mains voltage range in two steps, 200 to 225V and 230 to 250V. The shorting link across R_1 is removed for the high voltage range.

The timer shown in Fig. 25 is basically the same as that of Fig. 24 except that the trigger supply voltage is derived

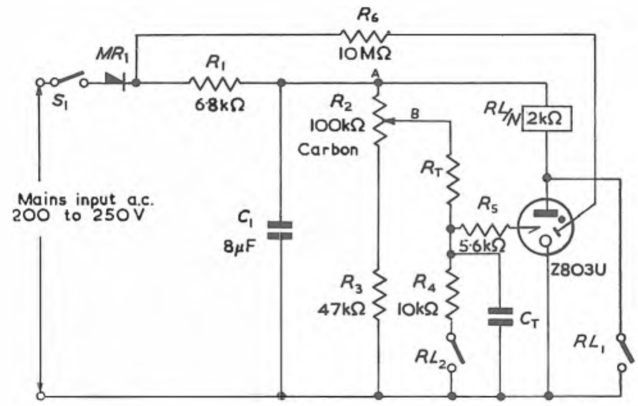


Fig. 25. Complete simple timer

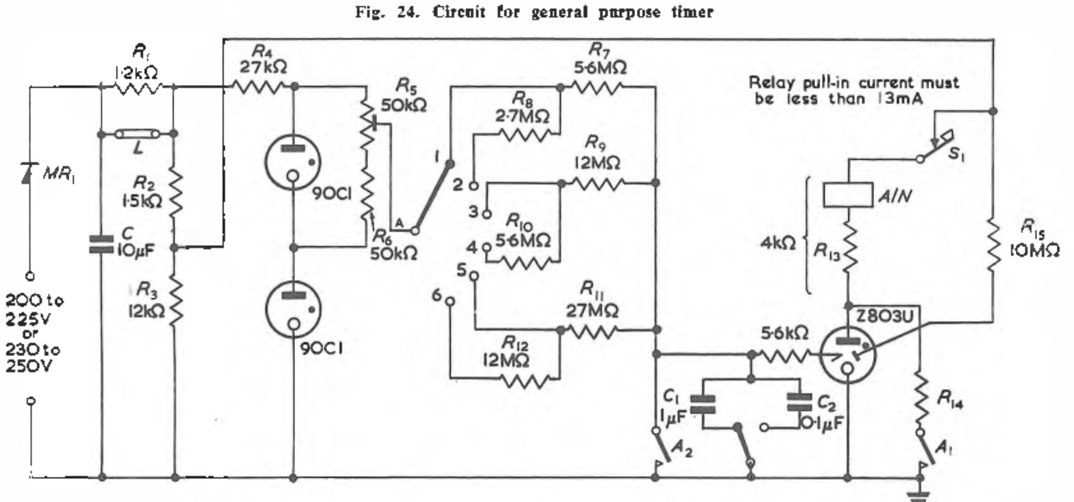
direct from the mains rather than from the two stabilizers. The performance of the timer will therefore only be inferior to the previous circuit by an amount proportional to the mains fluctuations.

The circuit has been designed to operate from the complete voltage range 200 to 250V. The timer is started by closing the mains switch S_1. The d.c. voltage at point A then rises in about 100msec to between 185 and 280V dependent on the local mains voltage.

A ratio of $V_{t(ign)}/V_s = 0{\cdot}8$ is used (where V_s is the voltage at B) giving a timing period $T = 1{\cdot}6R_TC_T$. The voltage at B is preset after installation to allow for the value of the local mains and also the tube to tube spread in $V_{t(ign)}$.

The priming discharge for the trigger tube is obtained via the limiting resistor R_6. Until the voltage at A has reached a fairly steady level the minimum d.c. voltage needed to initiate and maintain a priming

Fig. 24. Circuit for general purpose timer



discharge may not be present. This coupled with the normal delay in establishing such a discharge sets the minimum timing period to about 5sec.

Protection and Overload Circuits

Overvoltage and overload protective systems are fields in which the cold-cathode trigger tube has not been used extensively up to the present time, owing to the poor stability of the tubes previously available. However, very simple protective circuits using cold-cathode tubes can be made to perform the functions at present performed by trip relays and other more complex electromechanical devices. With the high stability of the newer tubes in both running and standby conditions the trigger tube circuits should have a much more stable trip point than their electromechanical equivalents. In addition they will require much less maintenance and be relatively impervious to atmospheric corrosion and temperature effects.

Because of the very small amount of power necessary to operate the cold-cathode tube, similar circuits can be designed which will cover a wide range of protection duties from high voltage power lines, medium sized electric motors and transformers down to sensitive meter movements. For the same reason they will have negligible effect on any current or voltage transformers to which they are connected.

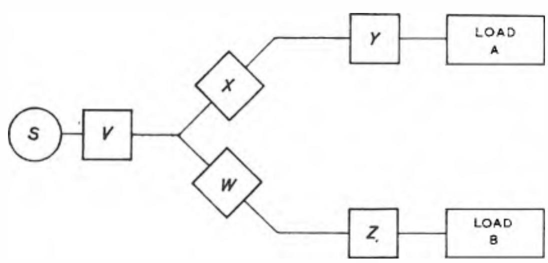


Fig. 26. Radial feeder system

Basically any protection device(s) must work in such a way that the complete system is protected against overload or overvoltage. However in the event of an overload only the faulty portion of the system must be isolated from the remainder.

Some of the operating characteristics required in general protection systems are demonstrated in the radial feeder shown schematically in Fig. 26. The power source *S* supplies loads A and B via the appropriate feeder lines.

The function of trips at *Y* and *Z* is to protect the system against excessive loads at A and B respectively. For this purpose a simple overload device is sometimes required which trips immediately the load exceeds a predetermined value; the trip would then be reset manually once the fault had been cleared.

Trips at *X* and *W* protect the supply against faults along the *XY* and *WZ* feeders respectively. Similarly the trip at *V* guards against a fault between it and the feeder junction.

By designing trip *X* in such a way that it will operate after an overload has been sustained for a short but definite period it is possible to set *X* to trip at the same overload level as *Y*. This gives full protection but ensures that a fault at load A only operates trip *Y*.

However, in complex systems a preferable approach is to give trip *X* the inverse time lag characteristic shown in Fige. 27 (full line); such a trip relay will ignore small transient overloads and normally introduce a delay in tripping which is inversely proportional to the overload. A similar characteristic is required at *W*.

If trip *V* has the dotted characteristic of Fig. 27 the remainder of the feeder is completely protected. The feeder from *V* has normally to carry the total load A + B, this factor determines the displacement between the characteristics of trips *V* and *X*.

In many cases it will be desirable to use trips with inverse time characteristics at both ends of a feeder, e.g. at *Y* as well as at *X*, thus making the system impervious to transient overloads. Under such circumstances it is clearly desirable to have the time lag between the operation of the two trips for any given overload, as small as possible. The minimum time is set by the possible variation in performance of any trip over long periods. In this respect the trigger tube should show to advantage over electromechanical relays.

Another form of protection which may be required at A and B is an overvoltage trip to protect the loads from

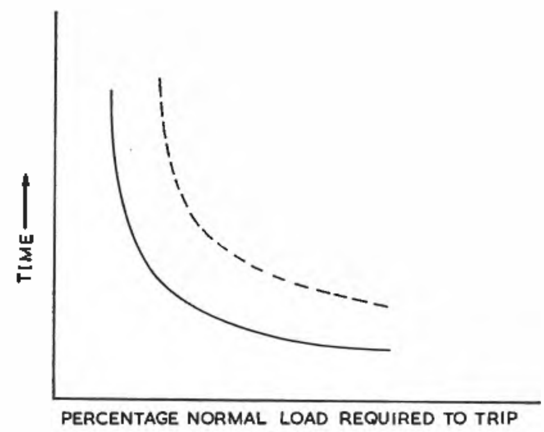


Fig. 27. Inverse time lag characteristic

excessive line voltage. In this form of protection it is useful to have the circuit self-resetting once the line voltage has returned to normal. It should be noticed that it is not feasible to use a self-resetting circuit for the overload circuits discussed previously, since the trip itself removes the fault by breaking the circuit. The result would be for the trip to chatter and no doubt cause considerable damage to the system.

In most overvoltage protection systems it will be undesirable for the system to be disconnected for every short surge and to overcome this disadvantage a circuit with a time delay is again required.

To sum up, for most electrical protection requirements the following basic circuits are needed:

(1) A simple instantaneous trip circuit with manual resetting.

(2) A delayed trip circuit with manual resetting.

(3) A delay trip with inverse time-overload characteristics.

(4) A self resetting instantaneous trip circuit.

(5) A self resetting delayed trip circuit.

Manual Resetting Overload Circuits

Fig. 28 shows a simple instantaneous trip circuit with manual resetting. A voltage proportional to the load on the supply is fed to the trigger of the tube and adjusted by R_1 to be below $V_{t(ign)}$ under normal conditions; the tube is therefore quiescent.

However if the load increases to the predetermined overload value the potential of the trigger is raised to $V_{t(ign)}$ and the tube ignites. Anode-cathode current pulls in the relay RA/N in the anode circuit which in turn holds itself closed on contact A_1 and extinguishes the tube. Other contacts on RA/N may be employed to operate circuit-breakers on other devices. The circuit is reset by momentarily disconnecting the h.t. supply.

The load on the current transformer can be made extremely small and thus should permit the use of an

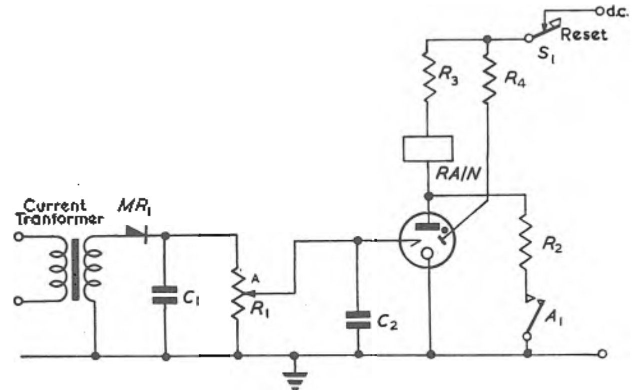


Fig. 28. Manual resetting circuit

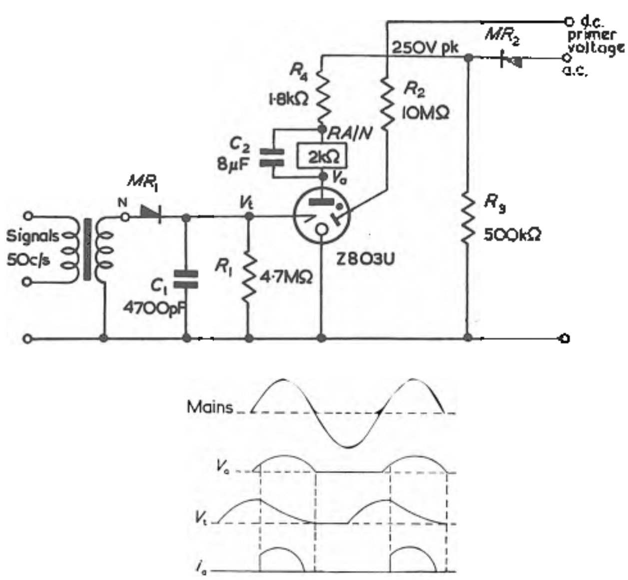


Fig. 29. Self resetting circuit with a.c. input

economical and simple design. For other applications the transformer could be replaced by another voltage source.

The capacitors C_1 and C_2 will introduce a delay in the operation of the circuit but by suitable choice of components this can be made negligible. To obtain a simple overload circuit with delay, C_2 is made large and a resistance R is inserted between A and B to give the required delay time. The delay is approximately given by

$$t = RC_2 \log \frac{V_A - V_0}{V_A - V_{t(ign)}}$$

where V_A is the voltage at point A during the overload.
V_0 is the voltage across C_2 prior to the overload.

The performance of the circuit is otherwise identical to that already described.

It will be noted that the delay time is proportional to both the overload and the previous load conditions, the delay time is reduced if either are large. This may be a desirable feature as it will ensure more rapid breaking of the circuit under the most severe fault conditions. However if a more consistent delay time is required it can be obtained by using one of the self-resetting circuits, described below, in conjunction with an accurate timer circuit.

Self-resetting Circuits

Since the cold-cathode tube, like the thyratron, cannot be extinguished by voltages applied to the control electrode the circuits discussed up till present require an external resetting mechanism. If however, the tube is extinguished at regular intervals by lowering the anode voltage, (e.g. with a rectified but unsmoothed a.c. anode supply), it is possible to switch the tube on and off by voltages applied to the trigger. If the trigger potential remains steady either above or below $V_{t(ign)}$ then the tube continues to conduct or not conduct as the case may be after each extinction period. If the trigger potential falls below $V_{t(ign)}$ between one of these extinction periods the tube remains extin-

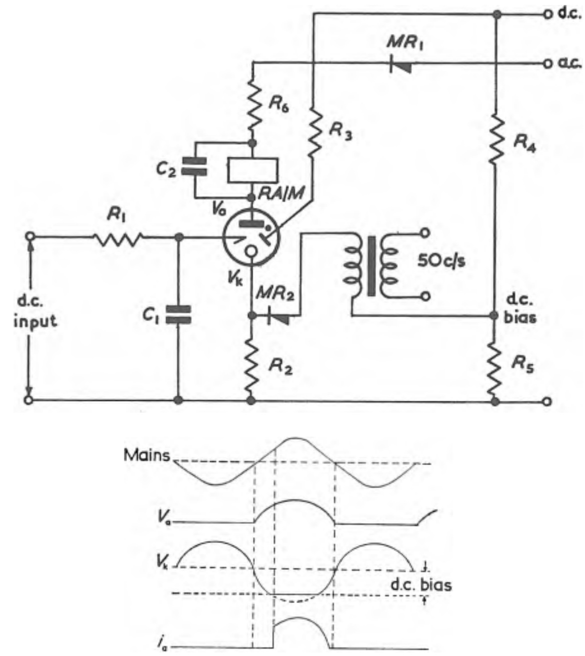


Fig. 30. Self resetting circuit with d.c. input

guished after the next extinction period. It will be seen therefore that the tube may be turned on or off by raising the trigger potential above and below $V_{t(ign)}$.

Figs. 29 and 30 show two basic self-resetting circuits and appropriate waveforms. The circuits shown in Fig. 29 has an a.c. signal voltage applied to the trigger and that in Fig. 30 has a d.c. signal voltage. In both circuits an unsmoothed half-wave rectified voltage is applied to the anode, and thus the anode potential is lowered below the burning voltage once every cycle.

In the case of the circuit shown in Fig. 29, the a.c. trigger signal voltage must be phase advanced with respect to the anode voltage V_a such that the peak trigger voltage occurs when V_a is in excess of the minimum anode working voltage. If the peak voltage of the signal exceeds $V_{t(ign)}$ the tube will conduct and the breakdowns are synchronized by the trigger signal.

In the circuit of Fig. 30 the trigger cathode ignition is synchronized to the mains by the chopped 50c/s voltage at the cathode. The cathode voltage during the chopped portion is zero volts and the tube ignites during this period as the cathode-trigger volts are then a maximum.

The trigger hysteresis described earlier has a detrimental effect on the performance of these two circuits. It introduces a backlash into the circuit and thus to extinguish the tube the signal must be lowered below the level at which

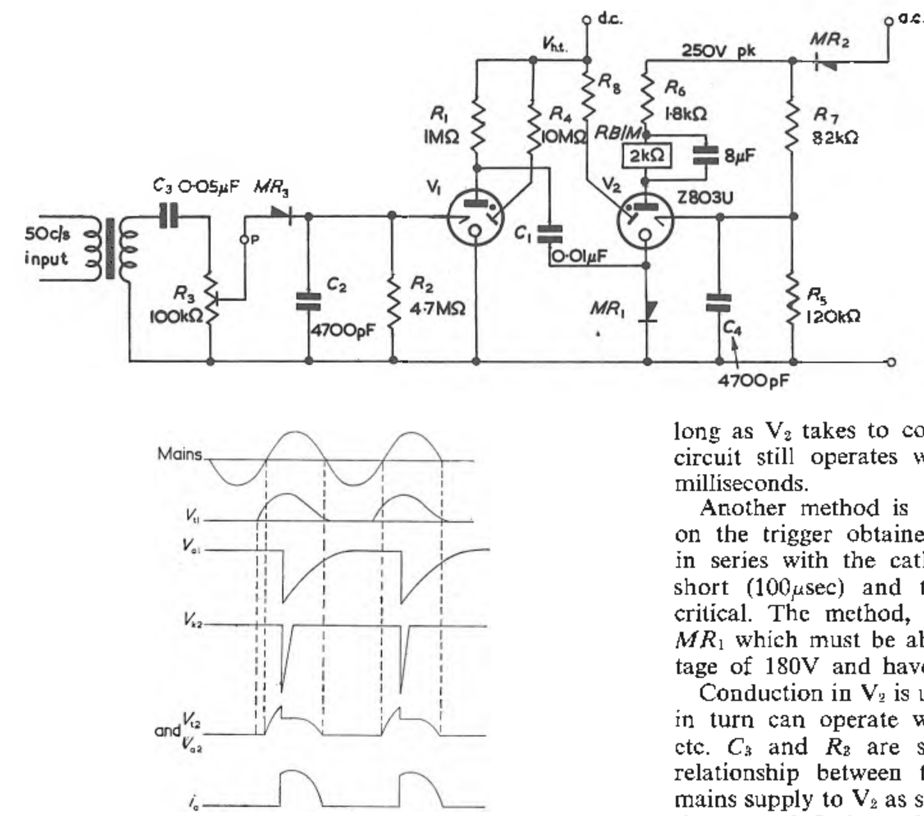


Fig. 31. Self resetting circuit with small backlash

conduction commenced. The magnitude of the backlash will depend on the anode-cathode current and will vary from 1 to 5V depending on the circuit and tube.

It was stated earlier that the hysteresis effect was dependent on both the duration and magnitude of the current and by reducing both to very small quantities the hysteresis could be made negligible.

Fig. 31 shows a circuit in which the duration of the discharge has been made short by the use of a self extinguishing circuit. This form of circuit has the required property of complete control by the trigger; but the average current which can be obtained from such a circuit is restricted by the conditions for self extinguishing to less than 100μA, which is insufficient to pull in a relay.

In the circuit of Fig. 31 V_1 is connected in a self extinguishing circuit as the sensing and amplifier valve. The second valve V_2 is the power output valve and has half-wave rectified a.c. applied to its anode and trigger. V_2 is quiescent unless it receives a negative pulse on its cathode from V_1. V_1 is triggered in synchronism with the mains using the methods already described.

The coupling circuit between V_1 and V_2 is unconventional and requires more description. Consider the circuit just prior to breakdown in V_1; the anode potential of V_1 is at full h.t., while the cathode of V_2 is substantially at earth potential. Diode MR_1 is conducting in the forward direction due to the small priming discharge current of V_2. When V_1 is triggered the anode potential of V_1 falls immediately to V_m and this fall of $V_{ht} - V_m$ is transmitted via the capacitor C_1 to the cathode of V_2. The diode is cut off and therefore momentarily C_1 cannot discharge. V_2 will thus have a large overvoltage between trigger and cathode during this period and after the ionization time will commence to conduct. The capacitor C_1 is then rapidly discharged via V_1 and V_2 in series. V_1 then behaves as if it were in a conventional self extinguishing circuit and is extinguished. The cathode of V_2 is returned to earth potential and the tube continues to conduct for the remainder of the positive half cycle in series with MR_1. The capacitor C_1 is recharged through R_1 and the forward resistance of MR_1 in series.

It should be noted that during the ionization time of V_2 V_1 passes a small steady current (approximately 130μA) determined by R_1.

The main advantage of the method described for triggering V_2 is that the pulse persists for as long as V_2 takes to commence conduction. With care the circuit still operates with delay time as long as several milliseconds.

Another method is to trigger V_2 with a positive pulse on the trigger obtained from a small resistor connected in series with the cathode of V_1. The pulse however is short (100μsec) and the tolerances are therefore more critical. The method, however, does eliminate the diode MR_1 which must be able to withstand a peak inverse voltage of 180V and have a small forward impedance.

Conduction in V_2 is used to operate a relay RB/M which in turn can operate warning circuits or circuit-breakers, etc. C_3 and R_3 are selected to give the correct phase relationship between the trigger voltage of V_1 and the mains supply to V_2 as shown by the waveforms. The potentiometer of R_3 is used to correct for tube to tube spread in $V_{t(ign)}$.

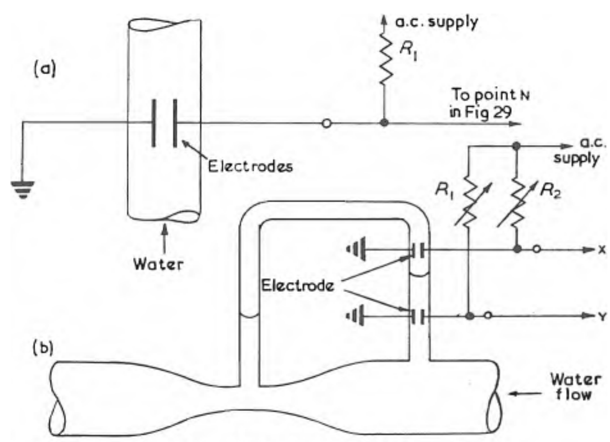


Fig. 32. Water flow detector

In this circuit the hysteresis or backlash effect should always be less than 0·5V, as compared with the 1 to 5V of the previous two circuits (Figs. 29 and 30). The circuit, however, is more complex and can only be justified in applications requiring sharp control.

So far the use of these circuits in electrical protection systems has been discussed but there are numerous applications in other fields where a fault or the completion of a process can be indicated by a change in the conductivity between two probes or electrode. The electrodes can form one arm of a potentiometer in the trigger circuit, a change in conductivity changing the trigger voltage from the ignition to non-ignition level or vice-versa. For example, Fig. 32 shows two systems for monitoring the flow of water in a cooling system where equipment can be damaged if the water supply fails. Fig. 32(a) shows an arrangement where such equipment is switched off if the pipe empties but is immediately switched on again once the pipe is filled with water. It employs either the circuit of Fig. 30 or Fig. 31 dependent on the sensitivity required.

Fig. 32(b) shows a system which checks that the flow of water either exceeds, or is below a certain rate, by the use of two probes and two sensing circuits derived from Fig. 31. The outputs X and Y feed into point P of Fig. 31 of their respective circuit.

Fig. 33 shows a meter overload protection circuit. The circuit is set up by adjusting the cathode-earth potential to be just less than $V_{t(ign)}$. In this application it is doubtful whether the cost of the supplies would justify its use with portable sensitive meters. However, when meters are incorporated in large equipments where suitable supplies are already available, the circuit offers a useful solution.

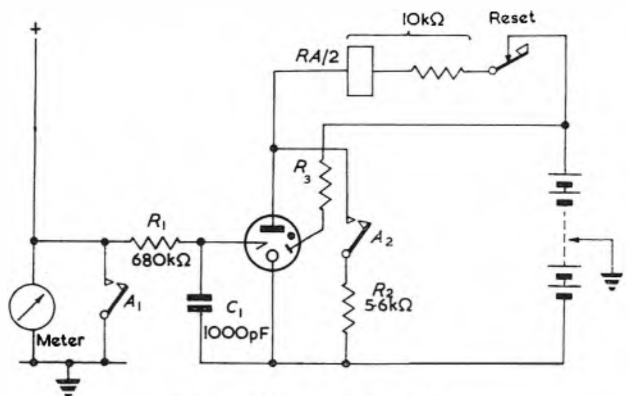


Fig. 33. Meter protection circuit

Simple Servo Systems and Voltage Regulators

In the previous section it has been shown that circuits can be designed which turn on or off as the trigger potential is raised above or lowered below the trigger breakdown voltage. If two such circuits are employed, set to operate at slightly different levels and the output employed to control the quantity being measured, a simple servo system is obtained. The system will maintain the quantity within the two preset limits to the stability of the tube and important components. For example in the water flow system described previously the output of the circuits could be employed to operate a mechanical valve. Thus the flow of water could be maintained between two limits.

Fig. 34 shows a complete servo type a.c. voltage regulator employing the above principles. It uses two circuits of the type shown in Fig. 31. A similar circuit could be designed for a d.c. voltage regulator.

In the circuit of Fig. 34, when the output voltage is correct the trigger voltage of V_1 is set such that V_1 and V_2 are conducting and that of V_4 such that V_3 and V_4 are non-conducting. In this condition contact RA_1 and RB_1 are broken and the servo motor is stationary. If the output voltage falls below the preset level the valves V_1 and V_2 will cease to conduct. Relay contact RA_1 will then be closed and the servo motor will operate in such a manner as to boost the voltage. Likewise if the voltage rises above the predetermined level V_3 and V_4 commence to conduct and contact RB_1 is closed. The servo can either operate a variac or change taps on a transformer.

It should be noted that the method of connexion of the trigger of V_2 and V_3 differs from that shown in Fig. 31. If a sudden overvoltage occurs the angle of firing of V_1 will advance and it is possible that the output pulse from the anode of V_1 may occur before the anode potential on V_2 is sufficiently large to permit triggering of V_2. This circuit will then fail to correct the output voltage. This has been overcome by obtaining the trigger voltage for V_2 from the output and adjusting the bleeder chain R_1 and R_2 so that V_2 does not conduct unless V_1 is conducting. However under excessive overload conditions the trigger potential on V_2 will exceed $V_{t(ign)}$ of the tube and V_2 will fire independently of V_1. The same principle applies to the operation of V_3 and V_4.

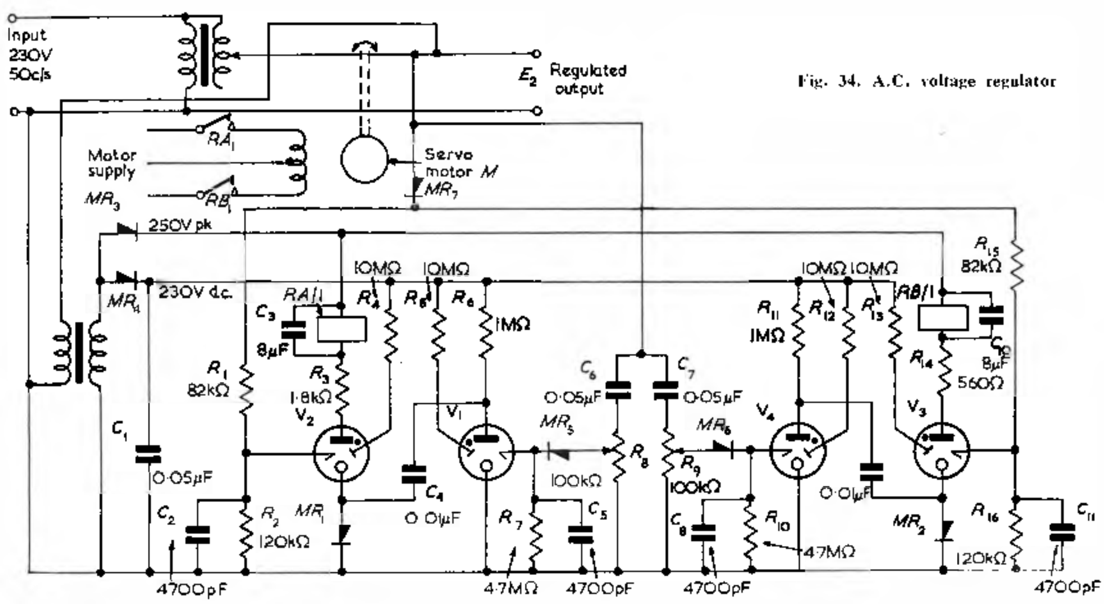


Fig. 34. A.C. voltage regulator

The stability of the two limits is determined by (a) the stability of $V_{t(ign)}$ of the Z803U, which is approximately 1 per cent and (b), the stability of the components, which should be better than 1 per cent. The overall stability of the limits should be better than ± 2 per cent.

The determination of the range between the limits is not simple. However, to prevent overlap of the two limits under the most unfavourable conditions, the range cannot be less than 4 per cent of the output. An additional tolerance must also be allowed to prevent hunting due to the inertia of the motor and relays. The exact amount will vary from system to system and must be adjusted accordingly.

(To be continued)

Difference Counters

By A. F. Fischmann*, M.E.E., A.M.Brit., I.R.E.

The difference counter is a device counting the difference in number between two sequences of pulses inserted at two different input terminals. Some applications are discussed, and two practical examples are given each being designed for a different speed. Finally, the design of a decimal difference counter is outlined.

A COUNTER may be classified as a difference counter if it is capable of counting the difference in number between two sequences of pulses inserted at two different input terminals. In Fig. 1, these terminals are designed as d_a (for addition), and d_s (for subtraction). The two corresponding outputs C_a and C_s transmit a carry if the stage has run through a full cycle in a certain direction.

Applications

The use of difference counters in a computer may simplify certain arithmetic operations. The author feels that in the past their use has not been sufficiently exploited, and it is hoped, that this article might contribute to a wider application of difference counters in single purpose computers and associated fields.

Various applications have been reported in the literature[1,2,3,4,5,8] and some novel arrangements will be described in this article. In certain cases, it may be necessary to record the number of randomly occuring pulses. As an example, one might think of a radioactive source whose disintegration should be faithfully recorded on magnetic tape. It would be extremely wasteful to design the tape recording system for a bandwidth which is capable of handling the maximum resolution it is intended to achieve. A better engineering solution will be obtained if the bandwidth is designed to handle the highest average pulse repetition frequency expected.

The block diagram of a suitable system is shown in Fig. 2. The zero position of the counter is defined in such a way that with a single input at d_a an 'add' carry appears at its output C_a (in this case, the zero position of the difference counter corresponds to the conventional 'zero minus one' position of a counter). The carry C_a opens a gate, and successive pulses appearing at d_a will be simply stored in the difference counter without affecting the output directly. As long as the gate remains open, evenly spaced clock pulses from the master clock will appear at its output, and are at the same time fed back into the 'subtract' input terminal d_s. Owing to the feedback, each pulse appearing at the output will reduce the number contained in the difference counter by one. Therefore, at a time when the gate has transmitted a number of clock pulses equal to the number of random pulses received by the counter, the latter returns to its previously defined zero, and a 'subtract' carry is transmitted which closes the gate until the next random pulse is received at the input. The pulse repetition frequency of the clock pulse generator should be slightly higher than the mean pulse repetition frequency of the random pulse source; the difference counter should be designed for the necessary resolution and be capable of holding the maximum number of pulses which might be expected to accumulate owing to the limited rate at which the counter will be cleared.

* *Scientific Department, Israeli Ministry of Defence.*

This arrangement may be used as a memory device if a switch is inserted between the master clock and the gate. As long as the switch is open, the incoming number will be stored in the difference counter. By closing the switch the memory will be cleared, and the number of clock pulses appearing at the output will be equal to the number previously stored in the difference counter.

Another application of a difference counter for obtaining multiplication directly will be given. For this purpose, a scaling down element is added to the arrangement of

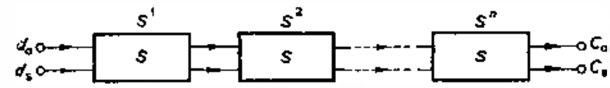


Fig. 1. Basic difference counter

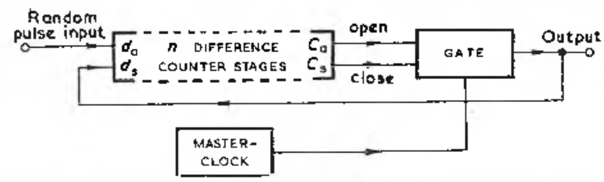


Fig. 2. System for counting random pulses

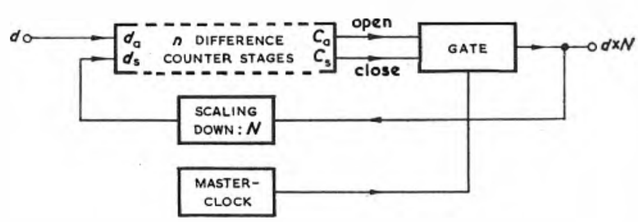


Fig. 3. Addition of scaling element for multiplication

Fig. 2, as shown in Fig. 3. The number d to be multiplied is fed into the terminal d_a. As before, the first pulse at d_a will cause an 'add' carry to appear at C_a, and thereby open the gate. The clock pulses appearing now at the output will also be fed into the scaling down element at whose output there will appear a single pulse for every N pulses fed into its input. As the output of the scaler is connected to the d_s terminal of the difference counter, one count will be subtracted from the number contained in the difference counter with every N^{th} clock pulse appearing at the output of the gate. Therefore, the latter will remain open until $d \times N$ clock pulses have passed through the gate. The last of these pulses will return the difference counter to its zero position as defined for Fig. 2, close the gate, and thereby terminate the action. Multiplication by any desired number N may be performed, provided the scaling down element may be set to this number.

With the aid of difference counters, an accounting machine for the direct addition and subtraction of pounds, shillings, and pence may be built by cascading one duodecimal, one vicenary (radix 20), and the required number of decimal stages. The design of a decimal difference counter is outlined later in this article and the other radices may be obtained by combining ternary, quaternary, and quinary stages.

The Design of Various Difference Counters

Two kinds of difference counters may be distinguished. One of them is based on the property of a multi-stage binary counter that it will reverse its counting direction if the carry input to a following stage is switched from one output lead of the preceding stage to the other, in all stages [1,2,6,7,8]. For this kind, the name 'reversible counters' would be most appropriate. The counter itself has only a single input into which both the add and subtract pulses are fed. For each pulse arriving at its input the counter will be instructed by a suitable sensing circuit to count in the proper direction.

The second kind of difference counters might be called 'straight difference counters' for the following reason: two input terminals are provided, one for addition, and one for subtraction. The direction of counting depends on which input terminal was selected. No separate action is required to reverse the counting direction, as in the case of the 'reversible counter'. Such a counter has been described[9,10], using ten pentodes in a ring-counter, and ten twin triodes for the transfer of the count. Some multi-electrode cold-cathode tubes[11,12] and certain high vacuum beam switching tubes[13], are also suitable for bi-directional switching, and may be used where speed is not of primary importance[14,15].

Low Speed Difference Counter

Two different versions of a 'straight difference counter' will now be described, each being designed for a different speed. Although their design was not carried out beyond the experimental stage, it is believed that the promising practical results obtained justify their inclusion in this article. The slower one is capable of operating at a maximum counting rate of 2×10^5p/s. It consists of a ring-counter of unspecified order, a diode matrix connecting each of the two inputs d_a and d_s to the anodes of the valves in the ring-counter, and two carries. The prototype of a suitable ring-counter has been discussed elsewhere[16]. There it has been shown that rings of a base up to ten are feasible, and a ternary stage of this kind is shown for completeness in Fig. 4.

This ring-counter was modified for bi-directional triggering by adding the coupling capacitors C' thereby making it completely symmetrical. Its action is as follows. It will be assumed that initially V_2 is conducting, and that a trigger of negative polarity is applied to the anode of V_0. This trigger will be transferred through the coupling capacitor C' and resistor R_1 to the grid of the conducting valve V_2, causing the anode voltage of the latter to increase. As the anode of V_2 is in turn connected through a coupling capacitor C and resistor R_1 to the grid of V_0, a regenerative action starts between V_2 and V_0 at the end of which V_2 will be 'off', and V_0 will be 'on'.

For complete analysis of this circuit it is also necessary to consider the voltage waveform appearing at the grid of V_1, during the regenerative action between V_2 and V_0. In order to prevent erratic operation this grid should, during that period, be sufficiently negative in order to keep V_1 cut off. Being connected through the capacitors C and C' and through the resistors R_1 to the anode of V_0 and V_2, its waveform is essentially determined by the transient voltage appearing at these two anodes. Under certain simplifying assumptions it may be shown mathematically that V_1 will remain cut off if $R_g \gg R_L$, and $R_g \gg r_a$, where r_a is the internal anode resistance of the valves. This behaviour is essentially explained by observing that the negative step

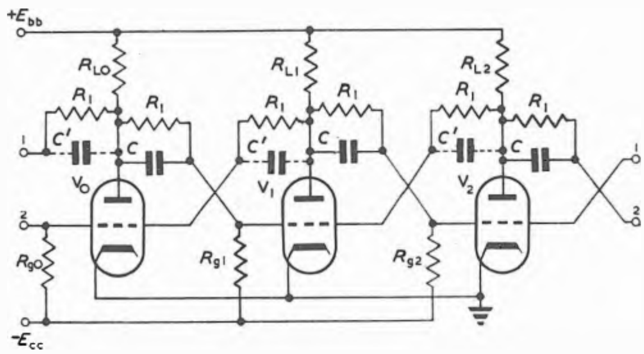


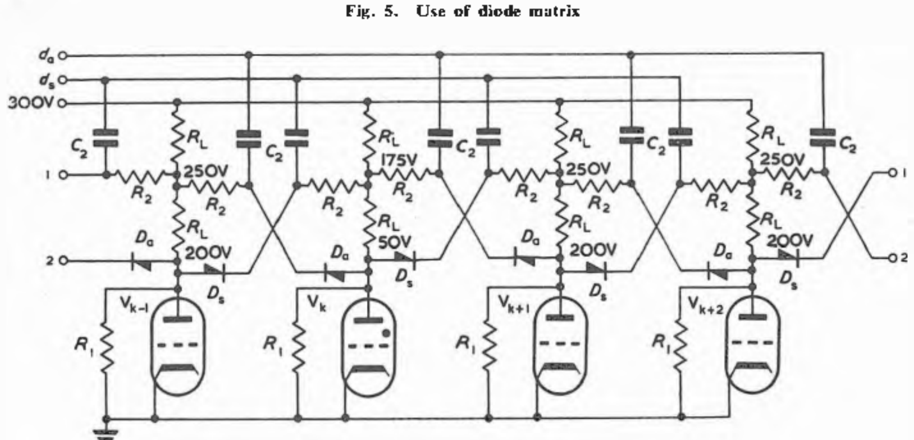
Fig. 4. Ternary stage of low speed counter (Terminals 1 and 2 at both margins of the diagram are connected together)

at the anode of V_0 (which is made to conduct) has a shorter rise-time than the positive step appearing at the anode of V_2 (which is being cut off), owing to the different time-constants associated with the anode circuit of the two valves. Therefore, as both anode waveforms of V_0 and V_2 are combined at the grid of V_1, the tendency of that grid to go negative prevails, and as V_1 was cut off at the beginning, it will remain cut off for the whole duration of the transient. These considerations are substantiated by practical results with the counter which show that the regenerative action is reasonably stable as long as the trigger is applied to the anode of the proper valve.

The same analysis may be extended to stages of higher than ternary order. However, it should be borne in mind that each grid is connected through the parallel combination of R_1 with C or C' to every anode except its own, i.e. in a counter stage of base n there will be connected to each grid $n - 1$ such combinations. Therefore, the time-constants associated with that circuit increase with the order of the ring-counter, and the speed of response decreases correspondingly.

The diode matrix is shown in Fig. 5. It is connected to four valves belonging to a ring-counter of unspecified order.

Fig. 5. Use of diode matrix



The anodes of the 'off' valves are prevented from reaching the full anode supply voltage by the voltage divider consisting of R_1 and the anode load resistor R_L of the valves. Valve V_k is conducting. The approximate voltage at the various points is indicated in the diagram. If now a trigger of an amplitude equal or less than −50V is applied to the terminal d_a, it will be transferred through the conducting one among the 'add' diodes D_a to the anode of V_{k+1} only, provided the internal capacitance of the other diodes is negligible. Similarly, a negative trigger at d_s will appear at the anode of V_{k-1} only, under the same restrictions as before.

Fig. 6 shows the diagram of the carry stage, and its waveforms. The valves V_o and V_n belong to a ring-counter of order $= n + 1$. An inspection of the V_o and V_n anode voltage waveforms for the 'add' counting direction in Fig. 6(a) indicates that an 'add' carry should be produced at the time t_o when the anode of V_o goes negative, and the anode of V_n goes positive simultaneously. This is accomplished by the valve V_{ca} ('add' carry) which will produce a negative pulse at the terminal C_a at the time t_o. The same is accomplished by V_{cs} for the 'subtract' counting direction. No pulse will be produced by V_{cs} in case of an 'add', nor by V_{ca} in case of a 'subtract' counting direction, provided the maximum permissible counting rate of 200kc/s is not exceeded. The speed of this circuit is mainly limited by the time-constant $R_{gc}C_3$.

The complete diagram of the single stage ternary straight difference counter is given in Fig. 7. The highest rate of response for stable operation is 2×10^5p/s. The values for the trigger shape are: maximum rise-time 1·5μsec; minimum width 1μsec; (in this case the rise-time was 0·1μsec) minimum amplitude 100V. From an inspection of the voltage values in the diode matrix, Fig. 5, it is seen that because of its minimum amplitude of 100V, actually each trigger will appear at all anodes, although with different amplitude. It is this difference in amplitude rather than the complete absence of the trigger itself at the anodes of the various valves which will cause the counter to operate in the wanted direction.

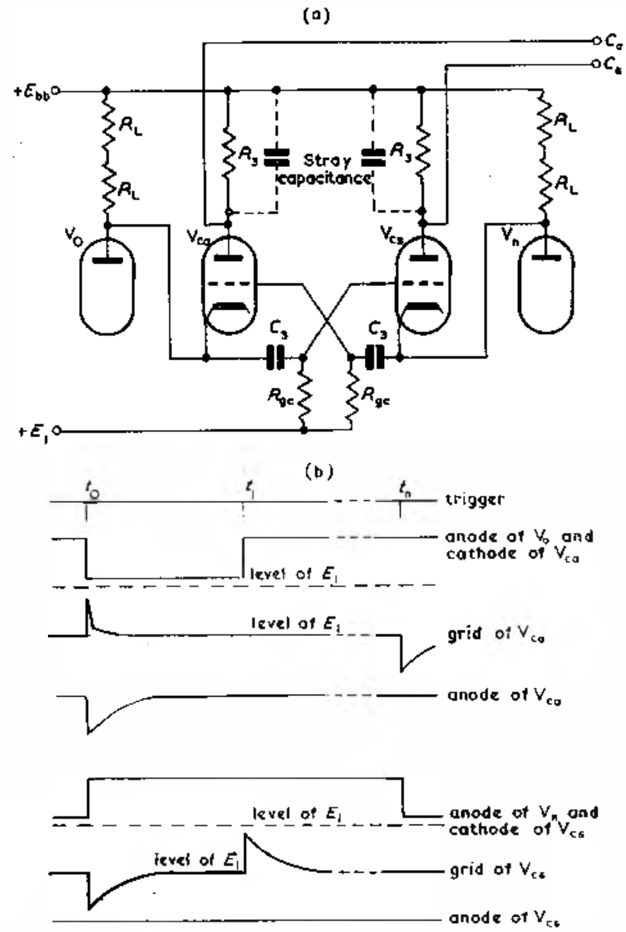
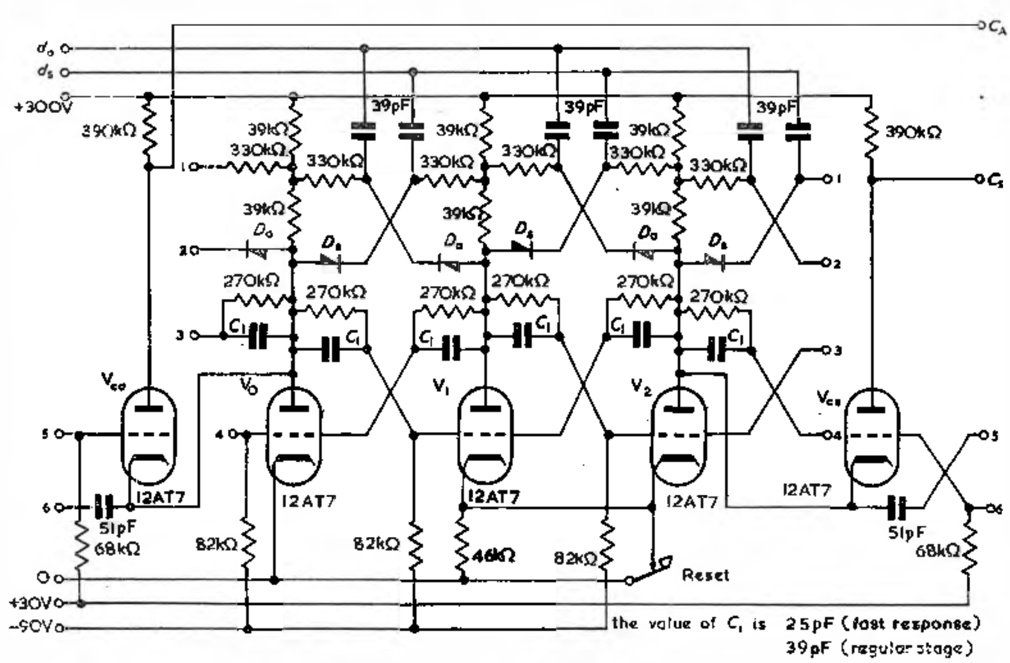


Fig. 6. Carry stage

Such stages may be directly cascaded as the carry output is capable of driving the next stage. An anti-coincidence input circuit is required to prevent the simultaneous arrival of pulses at the add and subtract input of the first stage. Waveforms obtained with this circuit are shown in Fig. 8.

It is known that cascaded counting units with feedback between them may be used to produce different scale-factors. By cascading one ternary and one quarternary stage one would obtain a count of twelve. By feeding back into the input two counts from the output for every ten counts received at the input, decimal counting may be obtained.

Fig. 7. Single stage ternary difference counter (Terminals one through six at both margins of the diagram are connected together)



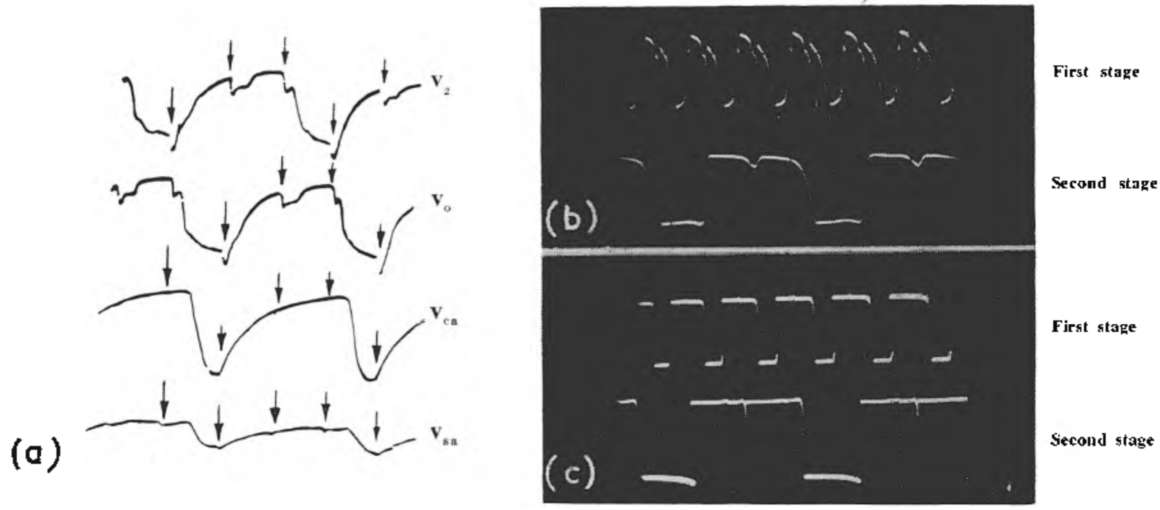
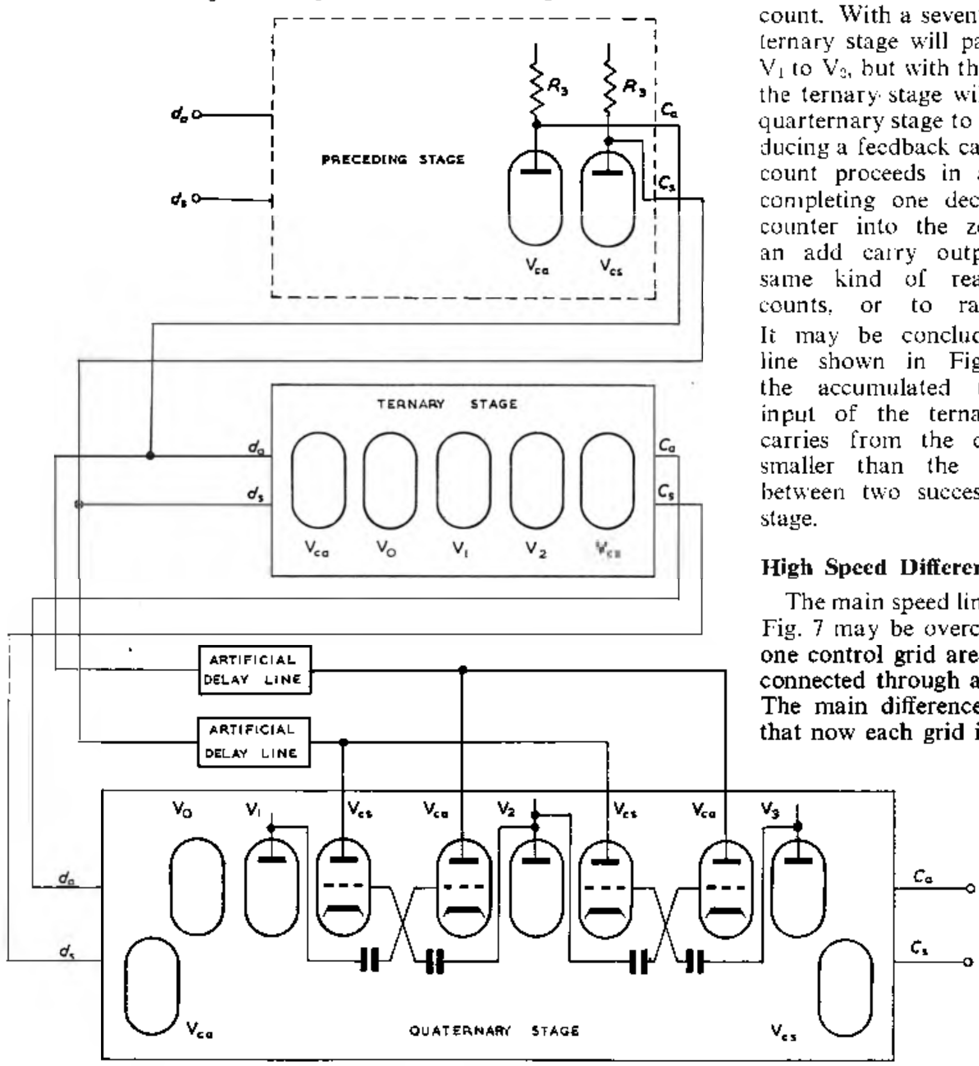


Fig. 8(a). Actual waveform with 2×10^3p/s. input at d_a ("add" counting direction). (b) and (c). Photographs of two successive stages on a double beam oscillograph. The mirror image is shown (time-base from right to left). (b) Input 2×10^3p/s. (c) Input 5×10^3p/s

The author proposes the logical arrangement of Fig. 9 whose operation is explained as follows. In the zero position, valves V_0 in the ternary and in the quaternary stages are conducting. The first five counts proceed in a regular fashion. However, with the sixth count the 'on' state in the quaternary stage will pass from V_1 to V_2, and a feedback carry will be produced by V_{ca} which is connected between these two valves.

Fig. 9. Arrangement for decimal counting

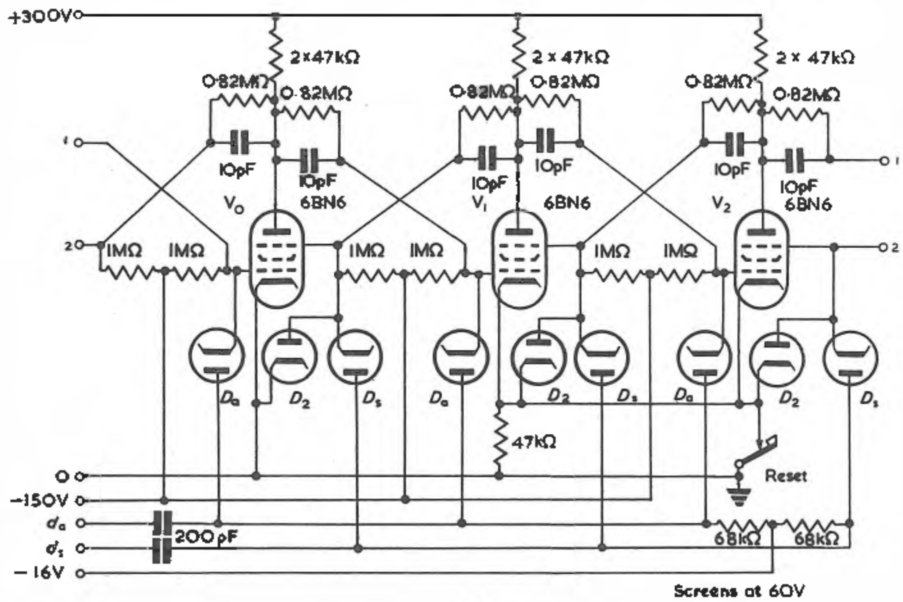


This carry is fed into d_a of the ternary stage which will therefore advance by an additional count. With a seventh count, the 'on' state in the ternary stage will pass in a regular fashion from V_1 to V_2, but with the eighth count, the carry from the ternary stage will cause the 'on' state in the quarternary stage to pass from V_3 to V_3, again producing a feedback carry pulse. The ninth and tenth count proceeds in a regular fashion, the latter completing one decimal cycle by returning the counter into the zero position, and producing an add carry output to the next stage. The same kind of reasoning applies to subtract counts, or to randomly intermingled ones. It may be concluded that the artificial delay line shown in Fig. 9 would be required if the accumulated trigger delay between the input of the ternary stage and the feedback carries from the quarternary stage should be smaller than the minimum permissible delay between two successive inputs into the ternary stage.

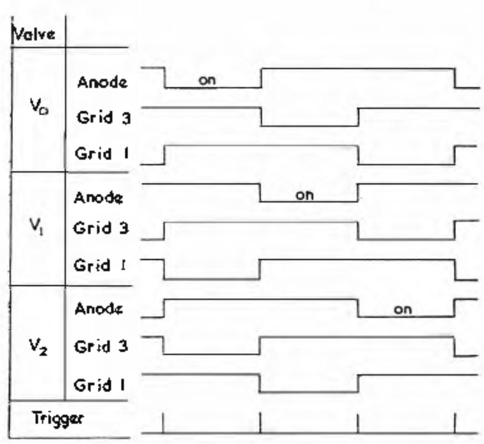
High Speed Difference Counter

The main speed limitation of the circuit shown in Fig. 7 may be overcome if valves with more than one control grid are used. As before, each grid is connected through a voltage divider to an anode. The main difference between the two circuits is that now each grid is connected to a single anode only, whereas before it was connected to $n - 1$ anodes, in a counter of base n. As in the previous case only a single valve may be 'on', namely the one whose control grids are all connected through voltage dividers to the anodes of all the 'off' valves and therefore are all above the cut off level. All the other valves are 'off' because one of their control grids is connected through a voltage divider to the anode of the single 'on' valve, and is therefore biased below its cut off voltage.

The base of counting obtainable with such stages is equal to the number of control grids available in each valve, plus one. Fig. 10 shows a ternary stage which uses dual control valves. Speeding up capacitors are connected between the anodes and the grids. This counter may be directionally triggered through the diodes D_a and D_s, as shown in Fig. 10. The operation of this circuit is explained as follows. The connexions from each anode to the $\left\{\begin{matrix}\text{first}\\ \text{third}\end{matrix}\right\}$ grid of the adjacent valve are made in the $\left\{\begin{matrix}\text{'add'}\\ \text{'subtract'}\end{matrix}\right\}$ direction $\left\{\begin{matrix}V_0 \rightarrow V_1 \rightarrow V_2 \rightarrow V_0\\ V_3 \rightarrow V_1 \rightarrow V_0 \rightarrow V_2\end{matrix}\right\}$. Therefore, assuming V_1 to be conducting, V_2 will be cut off by its first grid, and V_0 by its third one. An inspection of Fig. 9(a) shows that if a positive trigger is applied to all the $\left\{\begin{matrix}\text{first}\\ \text{third}\end{matrix}\right\}$ grids, it will cause $\left\{\begin{matrix}V_2\\ V_0\end{matrix}\right\}$ to conduct, thereby causing the state of conduction to advance in the $\left\{\begin{matrix}\text{'add'}\\ \text{'subtract'}\end{matrix}\right\}$ direction. In order to keep the driving point impedance at d_a and d_s reasonably high, these terminals are connected to the grids through diodes D_a and D_s which are biased with a voltage slightly greater than the cut off voltage of the valves. This arrangement will effectively connect only that grid to the trigger input which is at cut off potential, i.e., that particular grid which one intends to trigger. For the circuit shown, the pulse length required for reliable operation is rather critical, and should be about $0{\cdot}05\mu$sec. If the trigger pulses were of longer duration they would interfere with the negative going grid of the valve which is just being cut off, causing the counter to operate erratically. The pulse amplitude is not critical, about 10V minimum are required. The diodes D_2 prevent the suppressor grid from reaching a positive potential with respect to the cathode in the case when the valve is cut off by the first grid. This circuit is capable of operating at a speed of 10^7c/s.



(a)



(b)

Fig. 10. Ternary stage for high speed counter. (Terminals 1 and 2 at both margins of the diagram are connected together)

Acknowledgments

This article is based on a portion of the Master's Thesis by the author in the Department of Electrical Engineering, Polytechnic Institute of Brooklyn. The work was done in the Department of Physics and Biophysics of Sloan Kettering Cancer Research Institute, New York.

The help of Mr. Robert L. Schoenfeld as thesis adviser is gratefully acknowledged. In addition, thanks are due to Mr. Mones Berman, head of the electronics Laboratory, Department of Physics and Biophysics, Sloan Kettering Institute, for suggesting the problem and providing the facilities for doing the experimental work.

REFERENCES

1. Peck, E. R., Obetz Wendell, S. Reversible Fringe Counting. *J. Opt. Soc. Amer. 43*, 505 (1953).
2. Branin, F. H., Jr. Bidirectional Electronic Counter. *J. Opt. Soc. Amer. 43*, 839 (1953).
3. Winter, A. Magnetic Shaft Position Digitizer. *Electronics 26*, 214 (Aug. 1953)
4. Meyer, M. A., Gordon, B. M., Nicola, R. N. An Operational Digital Feedback Divider. *Trans. Inst. Radio Engrs. Prof. Group on Electronic Comput. EC-3* 17 (March 1954).
5. Meyer, M. A. Digital Techniques in Analog Systems. *Trans. Inst. Radio Engrs. Prof. Group on Electronic Comput. EC-3*, 23 (June 1954).
6. Barney, K. H. A Binary Quantizer. *Elect. Engng. N.Y. 68*, 962 (1949).
7. Trent, R. L. A Transistor Reversible Binary Counter. *Proc. Inst. Radio Engrs. 40*, 1562 (1952).
8. Fenmore, V. H. Reversible Binary Counter. *Electronic Engng. 27*, 204 (1955).
9. Regener, V. H. Reversible Decade Counting Circuit. *Rev. Sci. Instrum. 17*, 375 (1946).
10. Martens, F. H. Differential Counting with Reversible Decade Counting Circuits. *Rev. Sci. Instrum. 20*, 424 (1949).
11. Bacon, R. C., Pollard, J. R. The Dekatron. *Electronic Engng. 22*, 173 (1950).
12. Branson, L. C. Reversible Dekatron Counters. *Electronic Engng. 27*, 266 (1955).
13. Kuchinsky, S. Multiple Output Beam Switching Tubes for Computers and General Purpose Use. *Convention Record of the I.R.E. Pt. 6*, 43 (1953).
14. Churchill, J. L. New Trigger Circuits for Use with Cold Cathode Counting Tubes. *J. Brit. Instn. Radio Engrs. 2*, 497 (1952).
15. Kandiah, K. Multi Electrode Counting Tubes. *J. Brit. Instn. Radio Engrs. 15*, 221 (1955).
16. Beckwith, H. Flip Flop Counter Has Expanded Range. *Electronics 28*, 149 (Jan. 1955).

Simplified Pulse Transformer Design

By J. H. Smith*

A brief outline is given, for the non-specialist, of the design technique of pulse transformers. The discussion includes design of low-power pulse transformers and materials used, high power types and the relevant parameters, the Multiar transformer with a circuit description. Finally, a parameter for core selection in video or audio power transformers is derived.

A CONSIDERABLE number of books have discussed pulse transformer design[1,2,3], but these are written in such great detail that a considerable amount of time is used in referring to them. For the circuit designer, who is occasionally called upon to design pulse transformers, a simplified approach is desirable, so that designs may be predicted on paper, using only a minimum of mathematics. A logical approach is to consider the transformer equivalent circuit and endeavour to simplify it by using materials that enable losses to be neglected. The techniques described have been used by the author for a considerable number of designs and have been found satisfactory for experimental and engineering purposes.

The m.k.s. system of units has been used throughout the text, but for convenience a practical formula is also given in relevant cases.

Small Pulse Transformers

The availability of ferrite materials enables the design of small power transformers to be simplified. The high permeability and low losses of ferrites, coupled with cheapness, reliability and ease of assembly, make them an excellent material for use in pulse transformers. The methods to be described will apply to any type of core, except that core losses, which are to be neglected, may be significant. The equivalent circuit of a pulse transformer can be taken as that shown in Fig. 1, where:—

C_p = Primary stray capacitance

L'_L = Leakage inductance referred to primary

C'_p = Referred secondary capacitance

R'_L = Referred load

L_p = Primary inductance.

The effect of the above transformer on a perfect pulse is shown in Fig. 2.

Although each application will have its own particular requirements, it is generally required to preserve the pulse shape to an optimum. Designing the pulse transformer in rational steps, the first consideration is droop, which is due to the shunting (differentiating) effect of the primary inductance L_p. L_p is shunting the generator impedance and R'_L in parallel. If this parallel resistor is R' there is a droop of slope VR'/L_p. (See Fig. 3.)

The pulse length is T. As normal pulse droop is the order of 5 per cent,

$$L_p = 20TR' \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(1)}$$

If L_p is very large, (greater than 30mH), a larger droop may have to be tolerated. In practice, most pulse generators have a low impedance output. The actual value is best determined by shunting the output with various resistors, until the output voltage drops by 50 per cent, the resistor then equals the generator impedance. Note that this load is not necessarily the one used as the transformer load R_L. The primary turns N are now calculated from

$$L_p = \frac{N^2 \cdot A \cdot \mu'\mu_0}{l} \quad \ldots\ldots\ldots\ldots\ldots \text{(2)}$$

or,

$$L_p = \frac{4 \cdot \pi \cdot N^2 \cdot \mu \cdot A \cdot 10^{-5}}{10l} \text{ mH}$$

The overswing is due to the pulse droop and cannot be reduced except by reducing the droop; however, it can be given a long time-constant, if desired, by using a diode shunt. From the maximum power transfer theorem, $R'_L = R_G$ (generator impedance). This condition will give a maximum transfer of total pulse power, but it is seldom required for small pulse transformers: therefore R_L should be determined by the circuit designer.

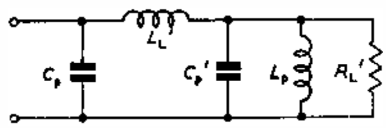


Fig. 1. Equivalent circuit of pulse transformer

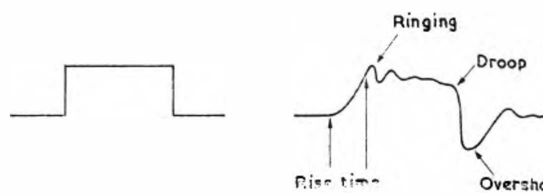


Fig. 2. The effect of a transformer on a perfect pulse

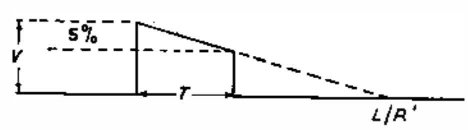


Fig. 3. 'Droop' due to primary inductance

The rise-time t of the pulse must now be considered. A maximum power transfer of the transient (leading edge) is required, with little or no ringing at the top of the pulse waveform. To satisfy these conditions, the ratio of energy stored (in the transformer), to energy dissipated in the load, must be a minimum.

This ratio is

$$E_S/E_T = \frac{KV^2 + L_L i^2}{2 \cdot i^2 \cdot R_L' \cdot t \cdot L_L} \text{ where } C'_p \cdot L_L = K$$

i.e. the winding configuration is constant, variations in turns are allowed for by changing the wire gauge.

Differentiating this equation and equating to zero, gives the result

$$R'_L = R_o = \sqrt{(L_L/C'_p)} \quad \ldots\ldots\ldots\ldots \text{(3)}$$

This makes the assumption that $iC'_p \ll iR'_L$. The rise-time (90 per cent value) is given as

$$t = \pi \cdot \sqrt{L_L \cdot C'_p} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(4)}$$

* Solartron Research Laboratory.

The effect of variations in R'_L is shown as in Fig. 4. R_o is known as the transient impedance of the transformer. The parameters L_p, L_L, C'_p can be calculated from the physical dimensions[1]. L_p is determined from equation (2). A formula for L_L is derived by assuming that the core has infinite permeability, hence all the leakage flux is in the space between the primary and secondary windings[1,2]. The

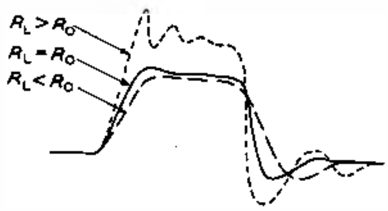


Fig. 4. Effect of variations in R_L

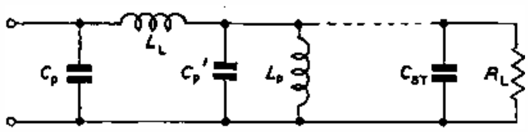


Fig. 5. Stray capacitance shunting R_L

energy of this 'waste' flux is equated to $1/2\ L_L \cdot i^2$, giving the result:

$$L_L = \frac{N^2 \cdot \delta \cdot U \cdot \mu_o}{l} \qquad \ldots\ldots\ldots\ldots\ldots (5)$$

or

$$L_L = \frac{4 \cdot \pi \cdot N^2 \cdot \delta \cdot U \cdot 10^{-2}}{10 l}\ \mu\text{H}$$

δ = Insulation (cm)

l = Winding length (cm)

N = Primary turns

ϵ = Dielectric constant of inter-layer insulation

U = Circumference (mean) of windings (cm)

μ = Permeability of core.

The windings must be the same length, and the wire thickness must be small compared with the insulation.

The capacitance of the winding is given as:

$$C_p = \frac{0{\cdot}0885 \cdot \epsilon \cdot U \cdot l}{\delta}\ \text{pF} \qquad \ldots\ldots\ldots (6)$$

$\therefore$ Where there is no capacitive load

$$R_o = \frac{377 \cdot N \cdot \delta}{l\sqrt{\epsilon}}\ \Omega \qquad \ldots\ldots\ldots\ldots\ldots (7)$$

This method can only be used for a few cases, such as a step-down transformer to an oscilloscope monitor point, where the stray capacitance is negligible.

In most applications there is a large stray capacitance shunting R_L, this gives an equivalent circuit, as in Fig. 5. The primary capacitance C_p, and the referred secondary capacitance C'_p are very small compared with the stray capacitance C'_{ST} referred to primary. If this stray capacitance can be estimated, the following procedure may be adopted:

As $R'_L \simeq \sqrt{(L_L/C'_{ST})}$, then, from equation (3) $L_L = R'^2_L\ C'_{ST}$. As L_p and N are already known from equations (1) and (2), the only unknown will be δ.

Hence from equations (5) and (3),

$$\delta = \frac{R'^2_L \cdot C'_{ST} \cdot l}{N^2 \cdot U \cdot \mu_o}\ \text{cm} \qquad \ldots\ldots\ldots\ldots (8)$$

or

$$\delta = \frac{R_L^2 \cdot C'_{ST} \cdot 10 \cdot l \cdot 10^3}{4 \cdot \pi \cdot U \cdot 2{\cdot}54}\ \text{inches}$$

δ is subject to the insulation being adequate for the voltage level. The rise-time can be calculated from

$$t = \pi\sqrt{(L_L \cdot C'_{ST})}$$

If this is too great, L_L or C'_{ST} must be reduced, or R_L can be smaller. The rise-time may be increased, in certain applications, by allowing the ripple to be present on the leading edge.

Lastly, the flux density must be calculated.

The basic transformer equation is

$$V_p = \frac{N \cdot d\,\Phi}{dt}$$

Φ = flux = $B.A$. (A is the cross-sectional area in square centimetres.

$$V_p = NA \cdot \frac{d(B_{(max)})}{dt}$$

$$\int_0^T \frac{V_p \cdot dt}{NA} = \int d \cdot (B_{(max)}). \qquad T \text{ is pulse length.}$$

$$B_{(max)} = \frac{V_p \cdot T}{NA}\ \text{e.m.u.}$$

or, in practical units,

$$B_{(max)} = \frac{V_p \cdot T \cdot 10^8}{NA}\ \text{gauss} \qquad \ldots\ldots\ldots\ldots (9)$$

This flux must not saturate the core material. For the grade of ferrite material commonly used, A4 or A2, the maximum flux is 2 500 gauss approximately. Numerous tests on pulse transformers have shown that it is desirable to test them under working conditions. Bridge measurements are often inaccurate, due to the fact that the magnetizing force is not known, and incorrect for the transformer.

When designing pulse transformers, the different merits of each case must be considered to determine which of the formulæ quoted should be used. In general, for coupling transformers with a turns ratio of 1 : 1 approximately, the formulæ (1), (2), (3), (4), (8) and (9) should be used.

For step-down transformers to monitor points, equations (1), (2), (3), (4), and (7) may be found more satisfactory. In cases where rise-time is an important criterion, a method using equations (1), (2), (3), (4), (5) and (6) should be devised.

In all cases, equation (9) should be checked.

Great care must be exercised in the application of these formulæ, as many desirable transformers are a practical impossibility.

It may be worth remembering that the majority of pulse transformers work into impedances less than 500Ω.

High Power Pulse Transformers

Although high power pulse transformers can be designed using the methods outlined, consideration of power and cost justify a different approach. As the cost of grain-oriented silicon iron (which is the normal material used) is very high, a method of core selection is required. A formula in terms of power, pulse length and core material can be found in the following manner. From equation (9)

$$V_p = \frac{N \cdot A \cdot B_{(max)} \cdot 10^{-8}}{T} \qquad \ldots\ldots\ldots\ldots (10)$$

where T is pulse length.

Then, from equations (2) and (10)

$$N^2 = \frac{V_p^2 \cdot T^2 \cdot 10^{16}}{B_{(max)}^2 \cdot A^2}$$

Substituting in equation (2) gives

$$L_p = \frac{V^2 \cdot T^2 \cdot 10^{16} \cdot \mu' \cdot A \cdot \mu_0}{B_{(max)}^2 \cdot A^2 \cdot l}$$

As the volume of the core is Al,

$$\text{Volume} = \frac{V_p^2 \cdot T^2 \cdot \mu' \cdot 10^{16} \cdot \mu_0}{B_{(max)}^2 \cdot L_p} \quad \ldots\ldots\ldots\ldots (11)$$

The third consideration used to estimate the core volume is pulse droop, which is normally 5 per cent (see Fig. 3).

$L/R = 20T.$

$L_p = R20T.$

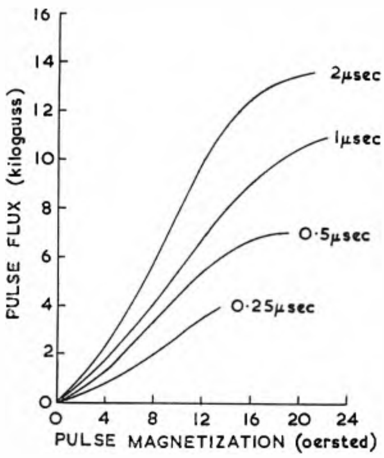


Fig. 6. Variation of magnetizing curve with pulse length

This is the worst case, as the shunt impedance of the generator, has been neglected.

Substituting for L_p in equation (11)

$$\text{Volume} = \frac{V_p^2 \cdot T^2 \cdot \mu' \cdot 10^{10} \cdot \mu_0}{B^2_{(max)} \cdot R \cdot 20T} \text{ cubic metres}$$

or,

$$\text{Volume} = \frac{4 \cdot \pi \cdot V_p^2 \cdot T^2 \cdot \mu \cdot 10^3}{B^2_{(max)} \cdot 10 \cdot R \cdot 20T} \text{ cm}^3$$

By using the linear portion only of the BH loop (Fig. 6) $\mu \simeq B/H$, and V_p^2/R = pulse power P.

$$\text{Volume} = \frac{2 \cdot \pi \cdot P \cdot T \cdot 10^6}{B_{(max)} \cdot H_{(max)}} \text{ cm}^3 \quad \ldots\ldots\ldots\ldots (12)$$

$B_{(max)}$ is progressively smaller for shorter pulse lengths, and should be found from manufacturers published data. Equation (12) will give the minimum volume required. The heating effects of this core must be considered separately.

Fig. 6 shows how the magnetizing curve varies for various pulse lengths T_1, T_2, T_3 . $T_3 < T_1$.

The losses in pulse transformers are very important as they are required to determine the heating effect and the efficiency of the transformer. Published curves show losses in terms of ergs/cycle, i.e. ergs/pulse/cm³ of core. The loss per pulse will be

$$\text{Volume} \times \text{ergs/cycle} \times 10^{-7} \text{ joules.}$$

Pulse energy is

$$\int_0^T Pdt = PT \text{ joules} \quad \ldots\ldots\ldots\ldots (13)$$

where P is in watts, T is in seconds.

Therefore the efficiency of the transformer is

$$\frac{\text{Pulse energy} - \text{loss per cycle}}{\text{pulse energy}} \times 100 \text{ per cent}$$

Copper losses are neglected as they are negligibly small. The mean power in the core will be − losses in joules × p.r.f. watts per second. From this figure the temperature gradients of the core, oil and can, are calculated. An alternative method of estimating temperature rise (preferred by the author) is to refer to tables quoting the power for a given transformer assembly, hence the losses for that assembly. Curves for pulsed core losses can be obtained from the English Electric Co. Ltd. Finally, leakage inductance must be considered. This is very important in high power pulse transformers; in particular, those that are driving magnetron loads.

Voltage insulation considerations make large variations of leakage inductance impossible. The latter may be varied by changing the material for insulation, or changing the winding configuration. Alternatively a selection of either a double or single loop core can be chosen. The leakage inductance required can be found from equation (3), modified for the turns ratio used[1]. In the case of a magnetron load, the stray capacitance of the load will swamp all the transformer capacitances, hence we have:

$$R_0 = \sqrt{(L_L/C_m')}$$

C_m' = Magnetron or load capacitance referred to primary

$$C_m' = C_m \cdot N^2$$

where N is the turns ratio

R_0 is the effective primary load resistance and transient impedance of the transformer.

∴ if load is R_L and C_m' then

$$L_L = \frac{R_L^2 \cdot C_m}{N^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (15)$$

This is the value required to give a transient match. An estimate of the leakage inductance can be made by referring to tables[1].

Materials Used for Pulse Transformers

Insulation

Several grades of insulation have been satisfactory, the most useful being varnished nylon, varnished terylene, varnished glass-cloth and polythene tape. These materials all have a working voltage of between 500 to 900V/mil. With the exception of polythene, these materials cover the R.C.S. 40/100 specification, i.e. 100°C.

Suitable Cores

(1) Mullard Ferroxcube 2E core type FX 1238. This uses a standard transformer bobbin, and is the standard size known as 450 or 531 in the R.C.S. specification. This can also be mounted in a standard oil filled can, hence covering all the humidity specifications.

(2) A range of pulse transformers for various applications were designed and manufactured by D. E. Lane, using 2E cores, type FX 1105/A4 and E and I cores, FX 1105/A4 and FX 1106/A4 respectively. These transformers were potted in Stypol 207E resin, hence covering all R.C.S. requirements. The only disadvantage of these cores is that special bobbins, clamps and moulds have to be manufactured. For small power transformers these cores are the most suitable manufactured at the present time.

(3) For high power application, 500kW up to several megawatts, grain oriented silicon iron is used. At the time

of writing, ·002in thickness was the best available, but thinner laminations are desirable to reduce losses. This material can be supplied by English Electric Co. Ltd., with relevant data. For medium power applications the Mullard Ferroxcube A2 grade is suitable.

Wire

The insulation between turns often causes trouble, due to the windings being 700V per turn approximately. The wire insulation to be expected is 800V/mil. for Lewmex and equivalent grades. This wire also covers the R.C.S. 40/100 temperature category. If the transformer is run at these temperatures, suitable solder and oil must be used. It is worthwhile noting that adhesives should not be used in the construction of pulse transformers; with the exception of polystyrene and other cements, as manufactured for use with coils.

Conclusion

A method has been described of designing low and high power pulse transformers. The methods shown are all derived from standard works on the subject, and simplified to a form suitable for the non-specialist. All the major design requirements are satisfied, and providing the correct materials are used, such as hermetically sealed cans, suitable solder joints, oil and wire, adequate insulation and terminals to stand voltage rating, there is no reason why the transformers should not be satisfactory for the majority of engineered units.

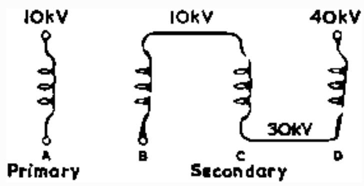


Fig. 7. Simple winding of large transformer

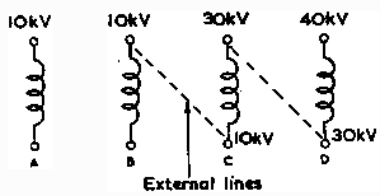


Fig. 8. Reduction of leakage inductance

Appendix 1

Consider a large transformer, designed for 10kV input and 40kV output. The primary turns may be 10. The insulation material could be 1 000V/mil., and 4 mil. may be available. The simplest winding is as shown in Fig. 7.

The windings are continuously wound. As both the primary and winding B have a gradient of 0-10kV, they may be bifilar wound. The second winding C is 20 turns and the third 10 turns, giving a 4:1 step up. The insulation between AB and C will be for 30kV, therefore eight layers of insulation will be used, using 32 mil. The insulation between C and D will be for 30kV, therefore this also uses 32 mil. This winding is perfectly satisfactory, but if a lower leakage inductance is required, it may be wound as shown in Fig. 8.

In this winding, different layers are linked externally, shown as dotted lines. If the same insulation is used, AB is bifilar wound, B to C is 20kV max. and uses five layers of insulation (20 mil.). As C and D are also 20kV max. then insulation is also 20 mil. It is readily seen that this winding is an improvement on Fig. 7, and should be used whenever it is required to reduce inductance to a minimum.

Appendix 2

The Multiar Circuit

Although the transformer used in this circuit is not a pulse transformer as such, it will come under the general heading, and uses the same design technique.

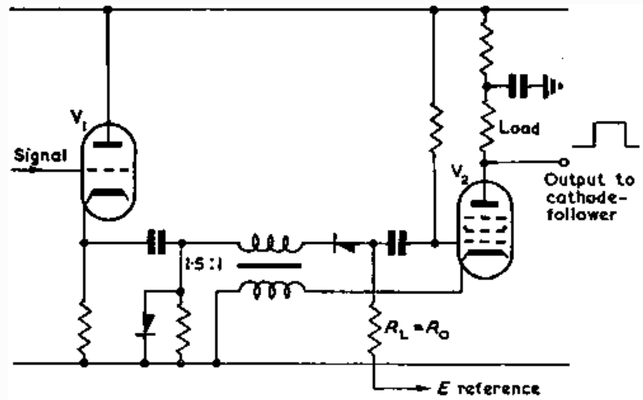


Fig. 9. The Multiar circuit

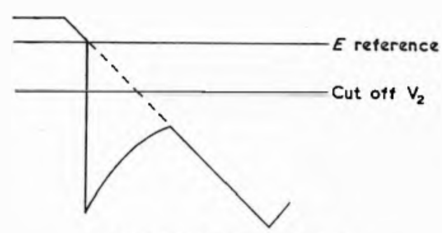


Fig. 10. Differential pulse

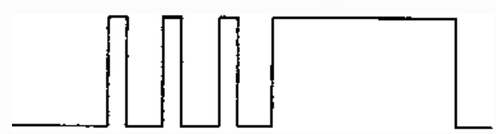


Fig. 11. Double triggering due to too rapid decay of differentiated cathode waveform

This circuit is most useful for amplitude comparison, pulse reforming and in pulse generators. The basic circuit giving optimum performance is as shown in Fig. 9.

A brief description of the circuit operation is as follows: V_1 is initially conducting, the grid is clamped to earth potential by R_g and i_g. If an input is applied to V_1, it falls below ground potential until it reaches the reference voltage; a voltage (δV) is then applied to the grid, tending to cut the valve off. However, as soon as V_g moves, regenerative feedback occurs, via the cathode, and the valve is rapidly cut off. It remains in this state until the grid potential falls, causing this valve to conduct and the action is again regenerative.

For the purpose of the transformer design, it must be appreciated that the transformer supplies only the initial cut-off potential, hence only the differentiated pulse is required to be fed-back, see Fig. 10.

It can be seen from Fig. 10, that the decay of the differentiated waveform must not be too fast, or the circuit will give double triggering, see Fig. 11.

The transformer must be designed primarily to give a fast rise-time, therefore equations (3), (4), (6) and (7) should be used. For these equations N (hence L_p) is found, and the rate of decay can then be calculated.

This circuit is capable of giving 0·05μsec rise-times, for any length of pulse input, with an output voltage of the order 30V. The figure of merit for any valve is not exceeded. The switch-over potential is determined by the diode.

APPENDIX 3

An expression for the core volume of audio and video power transformers can be found in a similar manner to that of the high power pulse transformer. As this is a useful starting point to the design of this type of transformer, it is mentioned here. The two main considerations of audio transformers are: the high frequency performance, determined by leakage inductance and stray capacitance, and also depending on the winding configuration and type of amplifier used, e.g. ultra linear; the low frequency response is determined by the primary inductance. The effects of core losses have been neglected. The estimate of core volume is given in terms of power and frequency response.

The three equations governing the low frequency response are:

$$V_p = 4{\cdot}44 \, . \, B_{(max)} \, . \, N \, . \, F \, . \, A \, . \, 10^{-8} \quad \text{(1C)}$$

$$L_p = \frac{N^2 \, . \, \mu' \, . \, A \, . \, \mu_o}{l} \quad \text{(2C)}$$

$$Z_L = 2\pi F \, . \, L_p \quad \text{(3C)}$$

$B_{(max)}$ = Maximum flux density for linearity required

N = Primary turns

F = Lowest frequency required (3dB point)

A = Core area (cm²)

l = Magnetic path (cm)

μ = Permeability

Z_L = Anode load of valve

The formulæ derived are for push-pull stages only, where there is no d.c. flux in the core. Single ended stages are designed using l.f. choke techniques.

From equations (2C) and (3C),

$$\frac{N^2 \, \mu' \, . \, A \, . \, \mu_o}{l} = L_p = (Z_L/2\pi F)$$

$$N^2 = \frac{l \, . \, Z_L}{2 \, . \, \pi \, . \, \mu' \, . \, A \, . \, F \, . \, \mu_o} \quad \text{(A)}$$

From equation (1C),

$$N = \frac{V_p \, . \, 10^8}{4{\cdot}44 \, . \, B_{(max)} \, . \, F \, . \, A}$$

$$\therefore N^2 = \frac{V_p^2 \, . \, 10^{16}}{(4{\cdot}44)^2 \, . \, B^2_{(max)} \, . \, F^2 \, . \, A^2} \quad \text{(B)}$$

Equating (A) and (B),

$$F = \frac{V_p^2 \, . \, 10 \, . \, 2 \, . \, \pi \, . \, \mu' \, . \, \mu_o}{(4{\cdot}44)^2 \, . \, B^2_{(max)} \, . \, A \, . \, l \, . \, Z_L}$$

As $W = V_p^2/Z_L$ (maximum output in watts)

$$F = \frac{W \, . \, \mu \, . \, 10^{14} \, . \, 2 \, . \, \pi \, . \, \mu_o}{(B^2_{(max)} \, A \, . \, l \, . \, (4{\cdot}44)^2} \simeq \frac{W \, . \, \mu \, . \, 4 \, . \, 10^7}{B^2_{(max)} \, . \, A \, . \, l}$$

in practical units.

The low frequency point (3dB) for a given core volume (cm) and power output is:

$$F\,(3\text{dB}) = \frac{W \, . \, \mu \, . \, 4 \, . \, 10^7}{B^2_{(max)} \, . \, V_p} \text{ c/s} \quad \text{(16)}$$

or

$$F\,(3\text{dB}) = \frac{W \, . \, 4 \, . \, 10^7}{B_{(max)} \, . \, H_{(max)} \, . \, V_p} \text{ c/s} \quad \text{(17)}$$

For a given frequency response, output power, and core material, the volume of a core required is:

$$V = \frac{W \, . \, \mu \, . \, 4 \, . \, 10^7}{B^2_{(max)} \, . \, F} \text{ cm}^3 \quad \text{(18)}$$

or

$$V = \frac{W \, . \, 4 \, . \, 10^7}{B_{(max)} \, . \, H_{(max)} \, . \, F} \text{ cm}^3 \quad \text{(19)}$$

From the above equations the following points may be noted:

(a) A material of high permeability and high saturation flux, will give a smaller core, hence lower leakage inductance.

(b) A smaller output power will give a better frequency response.

(c) A valve of low output impedance is generally better, due to smaller number of turns, giving lower copper losses.

This starting point in the design of output transformers neglects distortion due to the non-linearity of the *BH* loop. The frequency response and distortion of any given transformer may be improved by having it wired in the negative feedback loop of the amplifier.

Acknowledgments

Most of the work for this article was carried out at Decca Radar Ltd, and is published by kind permission of Mr. S. R. Tanner, Research Director. The author would also like to thank Mr. C. E. G. Bailey, of Solartron, for his comments and criticisms.

The Independent Television Service in Scotland

The new Scottish television transmitter at Blackhill can transmit programmes from the Glasgow studios of Scottish Television Ltd. and from all the main I.T.A. studios in England. This extensive service is made possible by a network of coaxial cable links, and super-high frequency radio systems, all of which have been provided by the British Post Office.

The connexions between Manchester and Glasgow and the new transmitter form the longest single vision link in the whole of the independent television network. This bothway vision link is routed on coaxial cables, and these follow the general direction of the west-coast railway route.

The vision circuits are carried on two coaxial pairs; and, within the same cables, there are other coaxial pairs and audio pairs carrying telephone circuits. From Carlisle, there are four coaxial pairs in the northward direction and six in the southward. Thus, throughout the whole route, the line plant and buildings are shared between the television and telephone services.

Between Telephone House, Manchester and the Post Office Repeater Station at Kirk-o'Shotts the vision signals are transmitted in the frequency band 0·5 to 4·0Mc/s, using the vestigial sideband mode, with a high-level carrier located at 1·056Mc/s. At the sending end of each section of the main link the video waveform is applied to a modulator which, by means of a double frequency-changing process, translates the signal into the frequency band to be transmitted to line. An inverse frequency-changing process is applied at the receiving end. The translating equipment was designed and installed by the General Electric Co. Ltd. In this way the total link is built-up from separate sections, each of zero gain; and, together, these serve to connect the main Post Office network switching centres at Manchester, Glasgow, and Kirk-o'Shotts. At each of these three places the vision signals appear in the standard video waveform, fully corrected at the standard switching level of 1V peak-to-peak across an impedance of 75Ω unbalanced to earth. Vision signals from studio circuits appear in similar condition, thereby enabling full flexibility of routing.

The main line amplifier system, which conveys the vision signals in the frequency band 0·5 to 4·0Mc/s, is the most recently developed broadband line system to be used by the British Post Office and was designed and installed by Standard Telephones & Cables Ltd.

The PN Junction on a Variable Reactance Device for F.M. Production

By D. C. Brown*, B.Sc., Ph.D., and F. Henderson†, D.C.Ae.

A completely transistorized modulator has been developed for the production of frequency modulation of a 10Mc/s carrier. A frequency deviation of 100kc/s is obtained by a change in the modulating signal of 0·15V and very good frequency modulation results with very little amplitude modulation.

THE small signal equivalent circuit of a pn junction is shown in Fig. 1, R_1 being the resistance across the junction, R_2 the resistance of the bulk p- and n-type material and C_J the capacitance of the junction. When the reverse voltage applied to such a junction is increased, the width of the space-charge region increases, making C_J decrease. The change in capacitance of a germanium pn junction (in which the density of donors in the base region

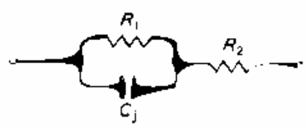


Fig. 1. The small signal equivalent circuit of a pn junction

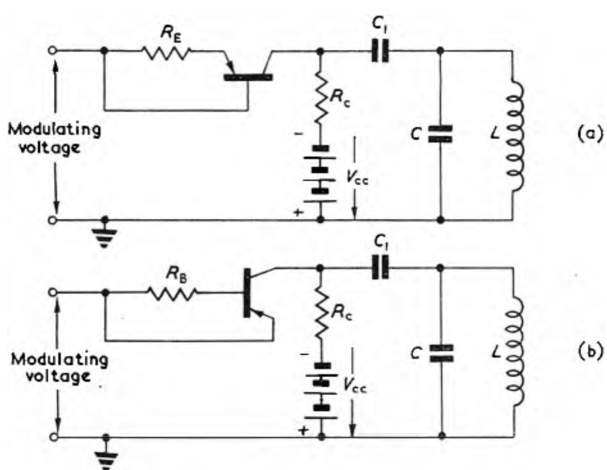


Fig. 2. (a) Grounded base and (b) grounded emitter configuration of the variable reactance transistor

In each case $C_1 \gg C$

is N_I) with applied voltage V, is found to agree very closely with the theoretical relationship[1]

$$C_J = 3{\cdot}4 \times 10^{-4}\,(N_I/V)^{\frac{1}{2}}\ \text{pF/cm}^2 \quad \ldots\ldots\ldots \quad (1)$$

If such a variable capacitance is placed in parallel with an inductor L and a capacitor C, then if C_J is very much greater than C, the resonant frequency of the tuned circuit changes in a linear fashion with change in the voltage applied to the pn junction.

A number of references occur in the literature to the use of this property for the production of frequency modulation and for automatic frequency control[2,3]. The change in collector impedance of a transistor as its emitter current is varied has also been used to produce frequency modulation[4]. However, the upper carrier frequency at which such a device can be used is dependent on the maximum frequency at which the transistor can operate. When the work described in this article was being done none of the available junction transistors were capable of operating at 10Mc/s; so the latter method could not be used. Likewise none of the available junction diodes had a sufficiently large capacitance to make them suitable for use at 10Mc/s Hence it was decided to use the collector capacitance of an audio transistor as the variable reactive element.

D. A. Thomas[4] deliberately used the damping effect on the oscillator tuned circuit of the variable output resistance of the transistor to obtain frequency modulation. The frequency deviation produced by this method is related in a very non-linear fashion to the modulating signal. In order to make the frequency deviation of the oscillator depend purely on the change in capacitance of the pn junction an

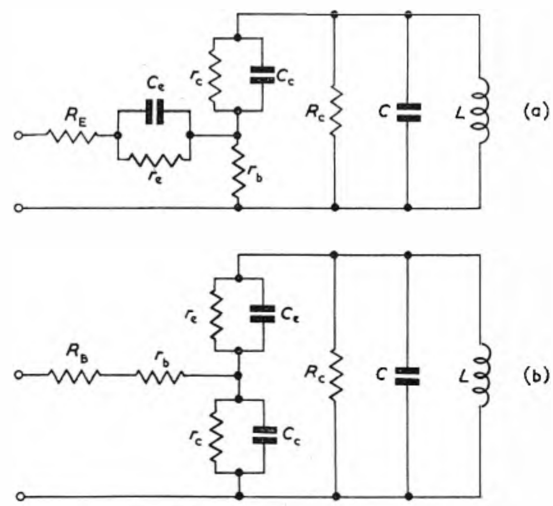


Fig. 3. The equivalent circuits of (a) the grounded base and (b) the grounded emitter configuration of the variable reactance transistor

As $C_1 \gg C$ it is neglected and r_b, r_c, r_e, C_c and C_e have their usual meaning

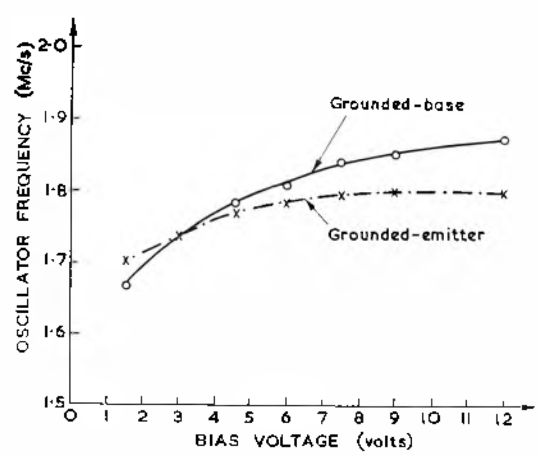


Fig. 4. Variation in frequency with bias voltage for the grounded emitter using an OC70

Bias voltage = voltage applied between collector and base

oscillator is chosen (a transitron) so that the change in shunt resistance has little effect on the oscillator frequency. The operating conditions of the variable reactance transistor are so chosen that the output resistance does not change appreciably as the applied voltage is altered.

Two possible means of using a transistor in this fashion are shown in Fig. 2 and the equivalent circuits in Fig. 3. When a voltage is applied to the grounded base configuration the percentage change in capacitance is greater than that of the grounded emitter configuration. This is illustrated in Fig. 4 which gives the change in frequency of an oscillator on applying a signal to a transistor in both the grounded base and grounded emitter configuration, the transistor being in parallel with the oscillator tuned circuit. The

* *The College of Aeronautics.* † *Gloster Aircraft Co. Ltd.*
This is part of the work submitted by F. Henderson for the Diploma of the College of Aeronautics.

temperature stability of the grounded base configuration is also inherently better than that of the grounded emitter configuration. For these reasons the grounded base configuration is chosen for use in the modulator.

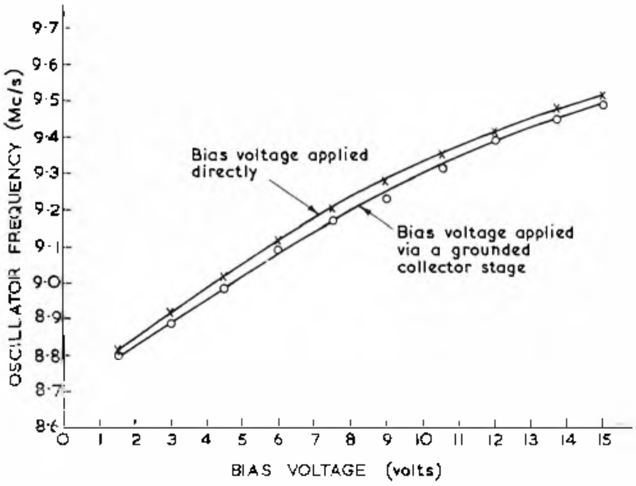


Fig. 5. Variation in frequency variation of bias voltage for a V10/50 (a) via a grounded collector stage and (b) directly

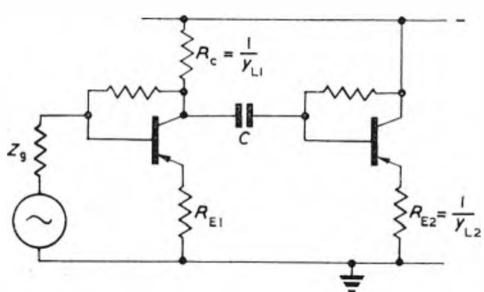
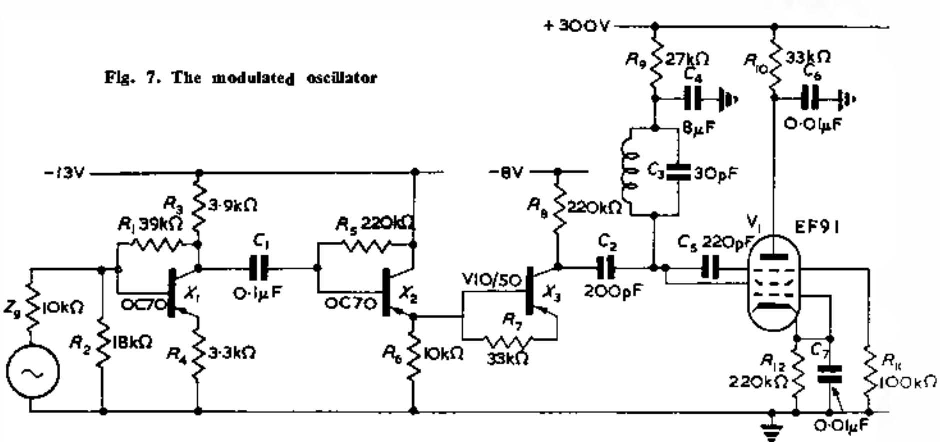


Fig. 6. The modulator amplifier

Fig. 7. The modulated oscillator



The Modulator

Fig. 5 shows the variation of the frequency of a transitron oscillator as the applied voltage is varied. The departure from linearity is quite small for applied voltages (or bias), between three and nine volts. The variable reactance transistor is therefore operated at a quiescent bias of six volts. In order to increase the low frequency response of the modulator amplifier d.c. coupling is used between the variable reactance transistor and the previous stage, and as a high input impedance is required to this stage, a grounded collector configuration is chosen. To obtain a quiescent operating bias of six volts (i.e. a base-collector voltage of six volts), as the voltage drop across the emitter load of the grounded collector stage is two volts, eight volts is chosen for the collector voltage supply on the variable reactance transistor.

Using such a method of injecting a bias voltage into the variable reactance transistor proves to be very efficient. This is shown in Fig. 5 which compares the frequency shift of an oscillator as the bias is applied directly from a battery and from a battery via the grounded collector stage.

As a frequency deviation of 100kc/s is desired when the signal input has an amplitude of 0·15V, an amplifying stage is required before the grounded collector stage having a voltage gain of about ten times. The modulating signal source has an impedance of 10kΩ; therefore the input impedance of the amplifier stage must be larger than 10kΩ. A grounded emitter configuration was chosen, with degeneration introduced by means of an emitter resistor, to both increase the input impedance and improve the stability factor*. The modulator amplifier circuit is shown in Fig. 6.

The input impedance of the grounded collector stage, Z_{in}, using the hybrid matrix parameters is given by

$$Z_{in} = \frac{1 + h_{11}y_{L2}}{h_{22} + y_{L2}(1 + h_{21})} \quad \ldots\ldots\ldots\ldots (2)$$

Hence for a transistor where y_{L2} is 10^{-4}mho, h_{11} is 70Ω, h_{22} is 7×10^{-7}mho, h_{21} is $-0{\cdot}97$ and h_{12} is 7×10^{-4} the input impedance Z_{in} is approximately 330kΩ. A 0·1μF coupling capacitor is therefore quite large enough to obtain a good low frequency response from the modulator amplifier. The input impedance of the grounded collector stage is so high that it can be ignored in the analysis of the grounded emitter stage.

The grounded emitter stage has an input impedance Z_{ie}' given by

$$Z_{ie}' = \frac{h_{11}h_{22} - h_{12}h_{21} + h_{11}y_{L1}}{h_{22} + y_{L1}(1 + h_{21})} \quad \ldots\ldots\ldots\ldots (3)$$

In this special case where degeneration is introduced by means of an emitter resistor R_E the input impedance becomes

$$Z_{in}' = \frac{(h_{11} + R_E)(h_{22} + y_{L1}) - h_{12}h_{21}}{h_{22} + y_{L1}(1 + h_{21})} \quad \ldots\ldots (4)$$

when R_E is 3·3kΩ and y_{L1} is $2{\cdot}5 \times 10^{-4}$mho, Z_{ie}' is approximately equal to 60kΩ, which is large compared to the source impedance. An additional stabilizing network is employed in the grounded emitter stage which results in a stability factor of 3·2.

The circuit diagram of the modulated oscillator is shown in Fig. 7, it is found to be very reliable, producing good quality frequency modulation with hardly any amplitude modulation. A frequency of 100Mc/s is by no means the upper limit at which this technique could be used. and employing some of the new r.f. transistors now available it should be possible to produce frequency modulation at hundreds of megacycles per second.

REFERENCES

1. SHOCKLEY, W. The Theory of p-n Junction Semi conductors. *Bell Syst. Tech. J. 28*, 435 (1949).
2. GIACOLETTO, L. J., O'CONNELL, J. A Variable Capacitance Germanium Junction Diode for U.H.F. *RCA Rev. 17*, 68 (1956).
3. PAN, W. Y., RAMANIS, O. Automatic Frequency Control of Television Receiver Using Junction Diodes. Transistors 1, p. 598 (R.C.A. Labs. Princeton N.J.).
4. THOMAS, D. E. Single-transistor f.m. Transmitter. *Electronics 27*, 130 (Feb. 1954).

* *Stability factor is defined as $(\partial I_C/\partial I_{C0})$, I_C being the collector current and I_{C0} the collector current when the emitter current is zero.*

LETTERS TO THE EDITOR

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

L-Type Iterated Networks

DEAR SIR,—In the correspondence columns of your issue for September, Mr. Forejt has given what is to me a most welcome result—a neat tabular method for calculating the elements of the n^{th} power of a certain 2×2 matrix. This matrix is $[A]$ in the equation:

$$[V_1/I_1] = [A]\ [V_2/I_2]$$

where V_1, I_1 and V_2, I_2 are the input and output voltages and currents for a network of iterated L-sections.

I should like to give an example of the use of Mr. Forejt's tabular scheme. Consider a network made up of three sections, each consisting of a series resistance R, and shunt admittance jY. The open-circuit voltage transfer ratio is given by $C_{11}^{(3)}$, and from the table:

$$C_{11}^{(3)} \equiv 1 + bjRY - 5R^2Y^2 - jR^3Y^3$$

If now we want this ratio to be wholly real, we have

$$6RY = R^3\ Y^3$$
$$\text{or}\quad b = R^2\ Y^2$$

Substituting this value in the expression for $C_{11}^{(3)}$:

$$C_{11}^{(3)} = 1 - 6 \times 5$$
$$= -29$$

—a well known result: it means that a three-section phase-shift network gives an attenuation of 29, and a phase-shift of π.

Now by use of Mr. Forejt's table, the amount of attenuation occurring in phase-shift networks of any number of sections can easily be calculated, using only the simplest of arithmetical operations.

Yours faithfully,
J. E. ROBSON,

Waymouth Gauges & Instruments Ltd.
Godalming.

The Characteristics of Varistors

DEAR SIR,—To Holbrook and Dulmage's recent article on the characteristics of metallic varistors[1] the writer would like to add the following observations.

Varistors with symmetrical characteristics, such as silicon carbide resistors, usually have each half of their characteristic resembling more nearly the reverse (blocking), rather than the forward (conducting), part of the characteristic of a rectifier. The current I and voltage V of these varistors are often stated to follow a power law, so that $|I|=a|V|^n$, a being a constant and n typically ranging from 2 to 4. Actually, this law is often not a very good fit over any very large range, although the reverse current of certain diodes with small junctions seems to follow a square law for large currents[2,3]. Also, the reverse current of certain silicon junction rectifiers, in which carrier generation in the depletion region is the dominant effect, follows quite well the law $|I|=a|V|^{1/n}$, n ranging here from 2 to 3, and being theoretically either 2 or 3 in certain simple cases[4,5].

Simpson and Armstrong[6] showed that many point contact rectifiers which they studied had reverse characteristics following the law

$$I = a\ e^{b|V|^{1/2}}\ (a \text{ and } b \text{ being constants}) \quad\ldots\ldots\ldots\ldots (1)$$

quite well over most of their working range. This was ascribed to the Schottky effect, the lowering of a barrier by the applied voltage. The writer has found that equation (1) represents the characteristic of many silicon carbide resistors rather better than does a power law, for $|V|$ greater than a volt or two. The modification of equation (1) into

$$I = a \sinh (b|V|^{1/2}) \quad\ldots\ldots\ldots\ldots (2)$$

gives a characteristic which goes through the origin, and represents the facts quite well for these varistors. It is noteworthy that equations (1) and (2) give in inflexion point; thus they can represent a characteristic which is first concave towards the V axis, and then convex, better than can a power law, which has no inflexion point. In fact, equation (1) or (2), modified by having $|V|$ raised to some other power, can represent the characteristics of a great variety of devices.

The reverse current in rectifiers may also be influenced by surface conduction[7], channels[8], and such phenomena. However, these are things so variable, and so dependent on circumstances, that it seems unwise to try to make them the basis of any theory of varistor action. The effect of self-heating may also need to be considered[9]; it is the cause, e.g., of the non-linearity in the characteristics of thermistors with sufficiently slowly varying signals.

As for the forward characteristics of rectifiers (which, of course, may be made into a symmetrical characteristic by connecting two rectifiers in parallel with opposite polarities), the form $\exp(qV/kT)$ for the current (V is positive), usually fails completely above two or three-tenths of a volt, and even the form $\exp(qV/nkT)$, n often being two, is seldom very good above one-half volt. It is rather surprising that, despite this, these expressions for the characteristic are so often stated without qualification in books and elsewhere. Herlet has given a rather different treatment[10], which leads to a square law for the forward characteristic at high levels.

In fact for higher applied voltages, around or more than a volt, the characteristic always approaches a straight line, with an intercept on the V axis of a few tenths of a volt[11]. Armstrong *et al.*[12] have given the expression for the characteristic

$$V = I\ \frac{R_R I + V_O}{I + I_N} \quad\ldots\ldots\ldots\ldots (3)$$

R_R is the 'residual resistance' the resistance represented by the slope of the straight line to which the characteristic is asymptotic. V_O is the intercept of the straight line on the V axis, and I_N is a constant. Equation (3) arose in the analysis from a rather drastic simplification of a theoretical expression; however, it could also be considered as just a convenient interpolation formula. If the characteristic is given, the straight asymptote may be drawn easily and R_R and V_O found; then I_N can be calculated, say by solving equation (3) for $V=V_O$.

An interesting feature of equation (3) is that an ordinary resistor is connected in series with the rectifier, the voltage V across the combination is given as a function of the current I by an expression like equation (3), but with different values in place of R_R and V_O. This is easily seen by writing the expression for the voltage across the combination, and bringing it to a common denominator. Thus this expression is convenient in investigating the voltage division of a rectifier and resistor combination as a function of the current. On the other hand, equation (3) may be solved for I as a function of V. The resulting expression, since it contains a surd, is not so convenient in such applications as calculating the harmonics produced when a sinusoidal voltage is applied. However, the required integrations and other operations can always be carried out graphically or numerically, if necessary.

Yours faithfully,
H. L. ARMSTRONG,
Pacific Semiconductors Ltd,
Culver City, California.

REFERENCES

1. HOLBROOK, G. W., DULMAGE, A. L. Some Characteristics of Metal Varistors. *Electronic Engng. 29*, 386 (1957).
2. GUNN, J. B. *Proc. Phys. Soc. B65*, 908 (1952).
3. PAIGE, E. G. S. The Isothermal Reverse Voltage-Current Characteristics of Small Area Alloy Contacts on Germanium. *J. Electronics 2*, 378 (1957).
4. PELL, E. M., ROE, G. M. *J. Appl. Phys. 27*, 768 (1956).
5. VELORIC, H. S., PRINCE, M. B. High-Voltage Conductivity-Modulated Silicon Rectifier. *Bell Syst. Tech.J. 36*, 975 (1957).
6. SIMPSON, J. H., ARMSTRONG, H. L. Reverse Characteristics of High Inverse Voltage Point Contact Germanium Rectifiers. *J. Appl. Phys. 24*, 25 (Jan. 1953).
7. LAW, J. T. A Mechanism for Water-induced Excess Reverse Dark Current on Grown Germanium n-p Junctions. *Proc. Inst. Radio Engrs. 42*, No. 9, 1367 (Sept. 1954).
8. CHRISTENSEN, H. Surface Conduction Channel Phenomena in Germanium. *Proc. Inst. Radio Engrs. 42*, No. 9, 1371 (Sept. 1954).
9. BURGESS, R. E. *Proc. Phys. Soc. B68*, 908 (1955).
10. HERLET, A. *Z. Naturf. 11a*, 498 (1956).
11. SHULMAN, R. G., MCMAHON, M. E. *J. Appl. Phys. 24*, 1267 (1953).
12. ARMSTRONG, H. L., METZ, E. D., WEIMAN, I. Design Theory and Experiments for Abrupt Hemispherical P-N Junction Diodes. *I.R.E. Trans. Electron Devices, ED-3*, No. 1, 1 (Jan. 1956).

The author's reply:

DEAR SIR,—In reply to Mr. Armstrong's letter, it should be appreciated from the outset that the laws suggested in the foregoing article are based entirely on observations made on a large number of varistors and are not based on any consideration of theoretical physics. While it is evident that certain theoretical formulae will express the characteristics of varistors over specific ranges of voltages, few of these formulae describe any one varistor over a wide range of voltages both positive and negative in sense.

It is agreed that the reverse resistance of metallic rectifiers is an extremely variable quantity within any given batch of samples. Although this may be of little significance in a normal rectifying role, it is often necessary in other situations to obtain standardization by using a shunt 'trimming' resistor which is considerably smaller than the nominal reverse resistance.

With regard to the exponential form of the forward resistance characteristic, it is agreed that this becomes of little value for substantially large values of forward voltage unless a constant series resistance is added to the equivalent circuit. The inclusion of such a resistor results in a linear relation between voltage and current for the higher positive voltages.

Equation (3) in Mr. Armstrong's letter is a very practical approach for all positive voltages and is, moreover, derived from a theoretical consideration of the solid state physics involved. It cannot, of course, be applied to negative voltages but, with this limitation, it provides a very convenient expression for the resistance of a rectifier in terms of current.

Although a discussion of silicon carbide varistors was not included in the original paper, I am interested in Mr. Armstrong's comments on their resistance laws. The work done by Ashworth *et al.*[1], based on an empirical relation of the form

$$V = kI^{\beta}$$

permits the use of some elegant mathematical methods for analysing the behaviour of this type of varistor. It is interesting to note that this form of equation which allows a linear display of voltage-current characteristics on a log-log basis can, and has been, applied to metallic and thermionic rectifiers over limited ranges.

It is certainly evident that a generalized and practical equivalent circuit for varistors which will describe their behaviour over all values of rated voltages and currents has not yet been developed from fundamental considerations. At the moment the user must choose either an empirical relation or employ a theoretical equation over a limited range.

Yours faithfully,

G. W. HOLBROOK,
Head of Department of Electrical Engineering,
Royal Miltary College,
Kingston,
Ontario, Canada.

REFERENCE

1. ASHWORTH, —, NEEDHAM, —, SELLARS,—. Silicon Carbine Non-Ohmic Resistors. *Proc I.E.E.*, *93*, Part I, 385 (1946).

Publications Received

HI-FI YEAR BOOK, second edition, has recently been published. It summarizes the progress made during the past twelve months. In the directory sections it lists the ranges of current equipment, together with abridged specifications, prices and manufacturers addresses. Miles Henslow Publications Ltd, 99 Mortimer Street, London, W.1. Price 10s. 6d.

RADIOISOTOPES, A NEW TOOL FOR INDUSTRY by Sidney Jefferson. The aim of this book is to give the businessman and the layman an insight into the many and varied ways in which industry has benefited, and can benefit in the future, from the use of radioactive materials. George Newnes Ltd, Tower House, Southampton Street, London, W.C.2. Price 17s. 6d.

THE 1957 INSTRUMENT DIRECTORY AND BUYERS' GUIDE, an auxiliary to "Instrument Practice", has recently been published by the United Trade Press Ltd, 9 Gough Square, Fleet Street, London, E.C.4. Price 10s.

SYNCHRONOUS CONVERTER is the subject of a leaflet (Publication 994) issued by George Kent Ltd, Luton, Bedfordshire.

TALBOT STEAD TUBE CO. LTD have issued a small booklet of useful engineering data for the benefit of users of their steel tubes and bars. The aim of the booklet is to present in compact form the data most commonly required and to enable speedy reference to be made to particular tables. Talbot Stead Tube Co. Ltd, Green Lane, Walsall.

THE INDEX OF TECHNICAL ARTICLES is a new monthly index of articles published in British Technical periodicals. Published by IOTA Services Ltd, 38 Farringdon Street, London, E.C.4.

VALVE AND TELEVISION TUBE EQUIVALENTS by B. B. Babani is the title of a manual recently published by Bernards (Publishers) Ltd. The Grampians, Western Gate, London, W.6. Price 5s. The author mentions that it includes some 3 000 more valves than other known publications on the subject.

M.A.C. A MEDIUM SIZED COMPUTER by A. F. Watson is a book published by The Morgan Crucible Co Ltd. Battersea Church Road, London, S.W.11. It gives an account of how this computer can be used, and is based upon a series of notes which were prepared to enable the author to train several members of the company in various aspects of the use of a Hollerith Type H.E.C. 4 computer. Although it deals specifically with one type of computer, it does detail the experiences associated with its application to a commercial organization.

RESONANT CIRCUITS, REPAIRING TELEVISION RECEIVERS, LC OSCILLATORS, and HOW TO INSTALL AND SERVICE INTERCOMMUNICATION SYSTEMS are four recent additions to the publications of John F. Rider, Inc, applicable to American practice. John F. Rider Publisher, Inc, 116 West 14th Street, New York 11, N.Y.

NATIONAL RESEARCH DEVELOPMENT CORPORATION have issued a report and statement of accounts for the year 1 July 1955 to 30 June 1956. This covers both development projects referred to in previous reports and new projects. Her Majesty's Stationery Office, Kingsway, London, W.C.2. Price 1s.

TELEVISION PROGRAMMING AND PRODUCTION by Richard Hubbell, 3rd edition. The following essentials are dealt with in this book. Camera techniques, montage, picture composition, shooting scripts, video and sound effects, use of music, lighting, acoustic perspective, how a programme is produced and directed. Chapman & Hall Ltd, 37 Essex Street, Strand, London, W.C.2. Price 32s.

TRANSISTOR TECHNIQUES is the subject of a recent addition to the Gernsback Library, Inc, 154 West 14th Street, New York 11, N.Y. Price $1·50.

UNDERSTANDING HI-FI CIRCUITS by Norman H. Crowhurst is a book which covers many of the technical points of this subject, special output stages, feedback, damping, inverter and driver stages, input stages, speaker distribution, volume controls, etc. Gernsback Library, Inc., 154 West 14th Street, New York 11, U.S.A. Price $2·90 paper cover, $5·00 cloth cover.

TRANSMISSION CIRCUITS by E. M. Williams, Professor and Head, Department of Electrical Engineering, Carnegie Institute of Technology, and J. B. Woodford, Assistant Professor of Electrical Engineering, Carnegie Institute of Technology. The text of this book comprises material used for some years in a course for senior students in electrical engineering at Carnegie Institute of Technology. It is assembled primarily for use in teaching rather than as a reference work. The Macmillan Co., New York. London Branch, 10 South Audley Street, London, W.1. Price 30s.

SEMICONDUCTOR ABSTRACTS, Volume III, contains abstracts of literature on semiconducting and luminescent materials and their applications compiled by the Battelle Memorial Institute. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, 37 Essex Street, London, W.C.2. Price 80s.

AN INTRODUCTION TO THE CATHODE-RAY OSCILLOSCOPE by Harley Carter. The author of this book has refrained from any attempt at mathematical treatment, and has endeavoured to render the explanations sufficiently simple for those having only slight knowledge of electronic circuits. Those who seek a simple explanation of the principles, construction and application of the cathode-ray oscilloscope will find this book of interest. Philips Technical Library, Eindhoven, Holland. Cleaver-Hume Press Ltd, 31 Wrights Lane, London, W.8. Price 12s. 6d.

ELECTROTECHNIQUE GENERALE by M. Denis Papin. The 5th edition of this aide-memoire has recently been published. In this edition greater emphasis has been given to electromagnetism and to the work of Sommerfeld, Ampere and Giorgi. Dunod Editeur, 92 rue Bonaparte, Paris vie. Price 480F.

AMPLIFIERS: DESIGN AND CONSTRUCTION edited by F. J. Camm deals with amplifier design and construction, from basic practical and design considerations to constructional details of a wide variety of amplifiers for tape recorders, gramophones, etc. George Newnes Ltd, Tower House, Southampton Street, London, W.C.2. Price 17s. 6d.

PICTORIAL MICROWAVE DICTIONARY by V. J. Young and M. W. Jones is a recent publication of John F. Rider, Publisher, Inc, 116 West 14 Street, New York 11, N.Y., U.S.A. Price $2·95. Derivation, explanation, definition and illustration are combined in such a way as to give the reader a coverage of microwave activity. It will be of interest to laboratory technicians, engineering students and others.

THE BBC RIVERSIDE TELEVISION STUDIOS: THE ARCHITECTURAL ASPECTS, by E. A. Fowler, is the title of No. 13 in the BBC Engineering Monograph series. BBC Publications, 35 Marylebone High Street, London, W.1. Price 5s. post free. Annual subscription 25s, post free.

RADIO RESEARCH 1956 includes the report of the Radio Research Board and the report of the Director of Radio Research. Radio wave propagation over a very wide frequency band has been one of the principal subjects of research during 1956. The research contributes information required for the efficient use of the radio frequency spectrum for communication and other purposes. The Radio Research Organization is participating in measurements on the ionosphere and radio noise during the present International Geophysical Year. Another and entirely separate line of research deals with semi-conductors and transistors. Her Majesty's Stationery Office, Kingsway, London, W.C.2. Price 3s.

BOOK REVIEWS

Electronic Components Handbook

Edited by Keith Henney and Craig Walsh. 230 pp. 80 figs. Demy 4to. McGraw-Hill Book Co., New York and London. 1957. Price 67s. 6d.

THIS book was written as a result of a contract from the U.S. Air Force (Wright Air Development Centre) at Dayton, Ohio, on the McGraw-Hill Book Company, as a guide to U.S. military contractors supplying electronic equipments to the Armed Forces. In the U.S.A. great concern is felt over the unreliability of military electronic equipment and considerable analysis of faults arising in complex electronic equipments has been done. It was found that valves and components were among the prime causes of failure, and while no book is available yet on obtaining more reliability from valves, this book is an attempt to give equipment designers more knowledge of what happens to components when used under arduous military conditions.

The title of this book is, unfortunately, somewhat misleading. The entire contents are devoted to military applications of components, so that the title should preferably read "Military Electronic Components Handbook". For users in this country the title should be "American Military Electronic Components Handbook", and as the book only covers four types of components, the correct title for this country should probably be "Some American Military Electronic Components Data".

The four classes of components covered in this book are resistors, capacitors, relays and switches. No mention is made of plugs, sockets, inductors, transformers, magnetic materials, batteries, wires, r.f. cables, etc. The main purpose has been to give some factual information on each type of those four classes of components for which a coordinated U.S. three-Service specification has been written. The pattern followed is the same for all four components—a general summary of characteristics of the main types followed by detailed descriptions of the military types standardized by the U.S. Armed Forces. Summaries of the military specifications for each component class type are given in the text and the gist of the individual specifications are also given in the appendix.

The first chapter deals with the reliability of military electronic equipment and is a useful summary of known facts. Rather too much time is spent on the statistical approach. One of the major difficulties in fault analysis not stressed is that, too often, components may have been stored for some years before use in equipments. Very little is known of the individual components' histories. However, there are some useful references given such as the U.S. Navy Electronic Laboratory Reliability Bibliography and the McGraw Hill book "Reliability Factors for Ground Electronic Equipment" reviewed in ELECTRIC ENGINEERING (December 1956). Chapter 2 deals with the U.S. Military Specification systems—military standards, U.S. qualified products list, specification writing, etc. The remaining four chapters deal with resistors, capacitors, relays and switches. The nomenclature used in many cases, particularly relays, does not agree with U.K. practice. The opening paragraphs describing the general characteristics of the four classes of components are clear and well written and contain much useful information on American practice. However, the remaining detailed data on U.S. standardized components is useful mainly to the American equipment contractor. There is certainly some useful information to the British reader but it is necessary to sort it out from the U.S. Military specification data. There are several pages of appendices which summarize the test procedures and requirements of military specifications for resistors and capacitors. The reader is warned that the data represents information from the specifications of 1956 and that the specifications themselves are under constant review and change so that designers should always refer to the latest issue of these specifications.

The book is inconveniently large for the average user's bookshelf, measuring 8¾in wide by 11¼in high.

G. W. A. DUMMER

An Introduction to Junction Transistor Theory

By R. D. Middlebrook. 296 pp. 60 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd., London. 1957. Price 68s.

THE large field of application of junction transistors in an ever widening range of electronic circuits must bring a great number of engineers and circuit designers in the position where they would wish to use transistors. Now, as it happens, the junction transistor cannot be used to its best advantage without a fairly detailed understanding of its mode of operation and without good knowledge of the physical significance of its equivalent circuit parameters. There is, hence, a very real need for a book which explains the physical theory of junction transistors to the applications engineer. The present book does just this. Moreover it does it admirably well and is wholly to be recommended.

The very initial chapters contain a few minor errors, but in view of the high general standard of the book these can be ignored.

The book succeeds in building up the theory and physical basis of transistor elements, i.e. semiconductors and pn junctions in semiconductors in the first six chapters forming Part I. Here, as much detail as is required is given, while the text in general flows freely from the more basic and more elementary to the more complicated and derived parts of the subject. The balance of how much detail to present and how much to leave for further study of the interested reader, is very successfully set at just the correct point.

Part II, consisting of six further chapters, is based on the results of Part I. Here the theory of the junction transistor is treated for both d.c. and a.c. operation, and based on this theory the small signal a.c. equivalent circuit is described and discussed.

Part III finally deals with the modifications to the equivalent circuit required in a number of practical applications. It also gives extensions to the physical theory to allow for some more sophisticated second order effects.

The illustrations and printing are clear and the whole book is produced with care. It will surely be a valuable addition to the library of most electronic engineers and many physicists.

K. HOSELITZ

On Human Communication

By Colin Cherry. 333 pp. 45 figs. Demy 8vo. J. Wiley & Sons, Inc., New York. Chapman & Hall Ltd, London. 1957. Price 54s.

THERE is today a tendency to regard the mathematical theory of communication as representative of all that there is to be said on the subject. The mathematical aspect has received close attention because much of our communication is by telephone and telegraph, where we need techniques which enable us to see if our lines will carry the messages offered, either (Fourier analysis) from the point of view of waveshape or (communication theory) of the information content.

But there is far more to the subject than the mathematics of the telecommunications engineer. There is the social aspect, including the importance of correctly-chosen communication paths in large organizations. There is the work of the phonetician: his concept of the phoneme, and his two different ways of analysing (and hence synthesizing) speech—by the physical actions of the larynx and the vocal tract, or by the energies and the frequencies of the emergent sound waves. Matters of language must be considered, including "meaning" as distinct from the mathematical theory's "information". And beyond all these topics is that of the human observer and the problem of recognition—what are the invariants which the brain seizes upon and by means of which it comprehends speech of different dialects, different voices, different degrees of clarity?

Such is the subject matter of *On Human Communication*, which is an introductory volume to a forthcoming series of studies in communication. It is an admirable introduction — informative, entertaining, exhaustive. The chapter on mathematical communication theory might perhaps with advantage have been presented more concisely, at the same time giving more explanation about the obscure matter of statistically matching a message source to its transmission medium. On the other hand, those people whose studies of communication theory have been hindered by the tendency of many writers to emphasize the analogy between information and the difficult thermodynamical notion of entropy will be grateful to Mr. Cherry for deprecating, as being overworked, the use of this analogy.

The finding of cross-references would have been easier if the chapter numbers had appeared at the head of each page, or as the first digit of the Section numbers. Perhaps subsequent volumes in the series will adopt this procedure.

T. L. CRAVEN

Engineering Electronics with Industrial Applications and Control

By J. D. Ryder. 666 pp. 554 figs. Medium 8vo. McGraw-Hill Book Co., New York and London. 1957. Price 71s. 6d.

THIS is the fortieth volume in the McGraw-Hill Electrical and Electronic Engineering Series. Its purpose, like that of so many of the earlier volumes, is to supply the background knowledge required by workers in the field of industrial electronics. While there is little doubt that this purpose has been adequately fulfilled, it is difficult to see what the book has to offer that is not already available.

For the benefit of the newcomer to electronic engineering the first three of the book's eighteen chapters review briefly the physical principles of electronic devices and develop the linear network representation of the thermionic valve. The decibel is introduced and, as seems to be common nowadays, the decibel gain of an amplifier is defined, virtually, as ten times the common logarithm of the indeterminate power ratio zero divided by zero! Chapters IV and V deal with the valve as a small-signal amplifier in earthed cathode, earthed anode, and earthed grid operation. The frequency and time responses of the resistor-capacitor coupled earthed cathode amplifier and the usual techniques for extending the range of frequency response are discussed. Power amplification is the subject of chapter VI and the topics covered include the effects of non-linearity, class A and class B operation, and push-pull input system.

Chapter VII is devoted to negative feedback. Black's equation is derived in the customary way but, as is usual in introductory expositions, it is not stated explicitly that the equation is valid only for the case of the feedback configuration for which it was derived. In the simple examples cited to illustrate the use of this equation the gain expressions, although correct, are wrongly obtained as may be verified by employing the same procedure in cases when the amplifier input impedance is other than infinite. Directly coupled amplifiers and methods of drift compensation are discussed in Chapter VIII. Special reference is made to the directly coupled amplifier connected as an operational amplifier for use in electronic differential analysers. Also considered in this connexion are function generators, limiting systems, and multiplying devices. Chapter IX is concerned with switching circuits and their applications in digital computing. The use of valves as electronic switches, waveform clippers, time-base generators, and relaxation oscillators is explained and the chapter concludes with such topics as storage methods, diode switching matrices, and logical circuits.

The following chapters deal with power supplies, filters, and regulators; class C amplifiers, oscillators, and high-frequency heating; semi-conductors, transistors, and photo-electric devices. Power rectification and power control are treated in Chapters XIV and XV and Chapter XVI has to do with relays, timers, and resistance welding controls. In Chapter XVII some of the basic principles pertinent to the electronic control of motors are considered and the final chapter outlines the fundamentals of servomechanism theory.

The field of industrial electronic engineering covered in the book is broad but the treatment is superficial. The emphasis is on the quantitative aspect of "non-radio" electronics. A list of problems appears at the end of each chapter but no answers are given. According to the information on the jacket, the book is the first to treat the "frequency response of amplifiers from both the sinusoidal and pulse standpoints." This is not so. The book will certainly be of value to advanced technical students and electronic engineers generally but it has no special claim to a significant place in the literature of electrical engineering science.

S. R. DEARDS

Transistor Circuits and Applications

By John M. Carroll. 234 pp. 310 figs. Royal 8vo. McGraw-Hill Book Co., New York and London. 1957. Price 56s. 6d.

THIS book comprises 106 technical articles that appeared in Electronics during the years 1950-1956. All the basic amplifier, oscillator, pulse and switching circuits are shown with typical component values. Many applications of transistors to home entertainment, military, broadcasting, communications, computing, control, industrial, scientific and medical equipment are described.

Digital Computers

By R. K. Livesley. 52 pp. 15 figs. Demy 8vo. Cambridge University Press. 1957. Price 8s. 6d.

THE author gives an introductory account of digital computers, what they are capable of doing and how they are made to do it. Throughout the book the emphasis is on the applications of computers to routine work rather than to advanced research.

ELECTRONIC EQUIPMENT

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

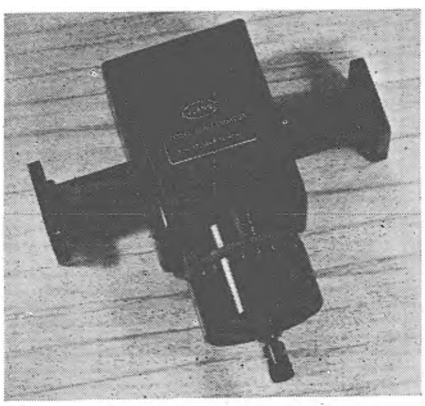
MICROWAVE INSTRUMENTS

(Illustrated below)

Flann Microwave Instruments Ltd, 133 Munster Road, Teddington, Middlesex

A FULL range of microwave instruments is being manufactured by Flann Microwave Instruments Ltd; typical of these are the grade 1 (R.C.S.C.) variable attenuators.

Each of these instruments consists of a length of flanged waveguide, upon which is mounted a channel that contains a kinematically supported carriage. One sprung ball-race in each of the two kinematic triangles, and a carriage return spring whose pressure is applied axially in line with the micrometer shaft, ensures a backlash-free system. A plate



mounted upon the moving carriage supports the vane rods that pass through the lossy sections into the waveguide.

The attenuating element is a matched glass vane upon which there is deposited a layer of nichrome covered by a magnesium fluride protective coating.

A close fitting dust cover surrounds the mechanism. A carriage lock screw prevents damage in transit.

Waveguide and flanges, which are silver plated and rhodium flashed, are finally finished on the exterior in a dark green eggshell paint. A polished teak case is supplied with each instrument.

Instruments can be supplied in any intermediate or special waveguide. On some waveguide sizes either 10dB or 40dB elements can be fitted.

Illustrated is the type 18/1. This has a maximum attenuation of 40dB and a maximum dissipation of 0·5W. The minimum resetting accuracy is ·005dB and the v.s.w.r. better than 0·95.

DIFFRACTION GRATING MEASURING SYSTEM

Ferranti Ltd, Ferry Road, Edinburgh, 5

AN extremely accurate measuring device giving instantaneous and continuous indication of distance travel on a Dekatron Counter has recently been made available by the Research Laboratory of Ferranti Limited.

The system utilizes two diffraction gratings which are superimposed, one tilted slightly with respect to the other. The integrated interference effects, caused by the angular intersection of the individual lines on each grating, produce a Moire fringe pattern of approximately sinusoidal distribution of density. When one grating is moved with respect to the other the fringe pattern travels at right angles to the direction of movement, the sense of the pattern movement depending upon the direction of the relative travel of the two gratings.

A photosensitive element integrating light over a section of the gratings can be made to give indication of the passage of each fringe and hence can be used to measure the relative travel of the gratings if the pitch is accurately known.

The accuracy of the system is a function of the number of lines per inch on the gratings, and this is chosen to suit the application. The system is entirely free from friction or wear, and resolution down to 0·0001in can be obtained.

A very important feature of the system is its inherently reversible nature. Change of direction or vibration does not introduce errors.

MINIATURE POLARIZED CHOPPER

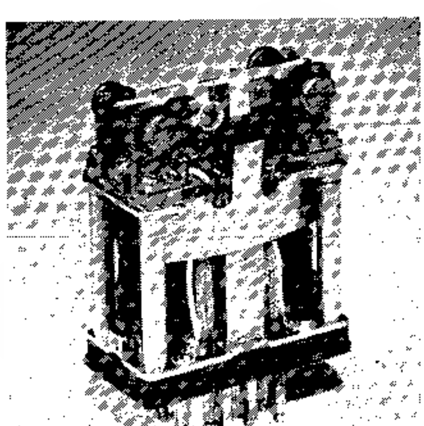
(Illustrated below)

Ericsson Telephones Ltd, Beeston, Nottingham

THIS polarized chopper, believed to be the first in this country incorporating two changeover contacts, has been specially developed for the conversion of low-power signals and/or amplification in devices such as computers and servomechanisms etc.

This two changeover feature will find many applications where the rectification of signals is necessary. The contact system can be either make-before-break or break-before-make.

If required any minor adjustments to suit individual circuit requirements can be easily made by means of the readily accessible contact screws. No further adjustments should be necessary until after several thousand hours when an



additional contact trim will prolong the life of the chopper indefinitely.

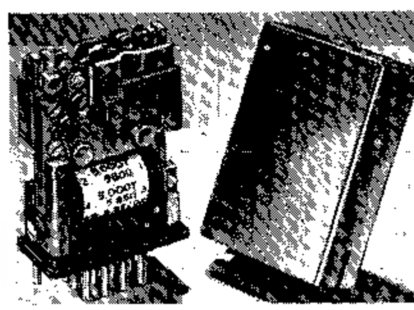
As a result of the Ericsson patented magnetic system, extremely low driving power is required, the effect of which together with the complete isolation of the contact assembly from the driving flux make the chopper practically free from any electromagnetic pick-up. Further, by careful screening, and good space relationship electrostatic interference in rendered negligible.

POLARIZED RELAYS

(Illustrated below)

Telephone Manufacturing Co. Ltd, Hollingsworth Works, Martell Road, West Dulwich, London, S.E.21

NOW available in each-side-stable and centre-stable forms, the Carpenter type 51 relay is an improved version of the well-known Type 5 relay.



At present only suitable for d.c. or low speed switching applications, the 51A each-side-stable relay has a nominal sensitivity of approximately $60\mu W$, while the 51M centre-stable relay, which is already finding uses as a detector in self-balancing bridge circuits, has a nominal sensitivity of approximately $10\mu W$.

Both these relays are suitable for aircraft uses, and comply with the various Ministry specifications.

EXPERIMENTAL CHASSIS SYSTEM

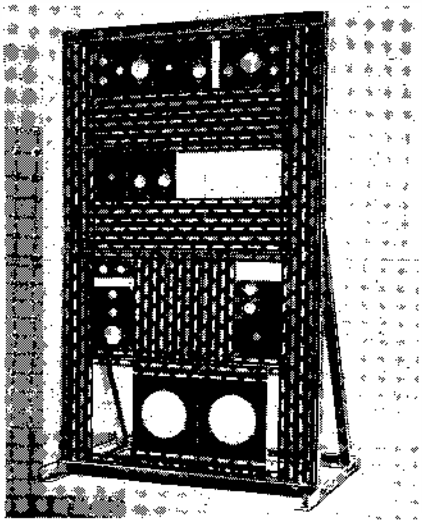
(Illustrated above right)

Shandon Electronics Ltd, 6 Cromwell Place, London, S.W.7

KNOWN as 'Rapikon' this system allows chassis assemblies for experimental and development work to be constructed and modified easily and quickly with the use of a screwdriver only.

Pre-punched holder plates take such components as valves etc. These plates are individually placed where required. One end is prolonged so that they can be slipped into position behind tag panels and components—and just as easily moved if necessary. Wiring can consequently be kept short. The slotted plates and girders are mounted vertically on the rack to any required layout.

Because of this vertical layout, components are always accessible for wiring, replacement or measurement. Layouts



can easily and rapidly be modified. Several chassis or circuits can be simply connected to one another without taking up bench space and with no flying leads.

TRANSISTOR TEST SET

Standard Telephones & Cables Ltd, Connaught House, Aldwych, London, W.C.2

THE 74163-A transistor test set, is a small portable set for measuring the current gain of transistors of the pnp type in the common emitter configuration. It will measure gains ($\alpha/(1-\alpha)$ or α') of 10 to 100, and in some cases gains of less than 10 may be measured, depending on the characteristics of the transistor. An indication of the collector current with the base open-circuited (I_{c0}) is also given, up to 2mA. The current gain is measured with an emitter current of approximately 2mA and with a collector voltage of approximately $-1{\cdot}5$V. The measuring accuracy is $\pm$ 10 per cent.

The set is completely self-contained in a light alloy box with dimensions approximately 10 × 7 × 4½in it operates from four 1·5V dry cells (U2 or equivalent) which are housed in the box.

PROCESS TIMER

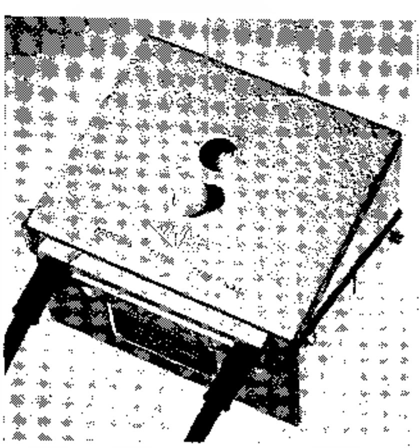
(Illustrated below)

Airmec Ltd., High Wycombe, Buckinghamshire

THE process timer type N237 provides an accurately timed switching facility for industrial control purposes over the range 1 to 10 and 10 to 100sec.

The starting signal may consist of a connexion or disconnexion of a pair of contacts and may be either a short pulse or a contact longer than the timing period. The versatility of the timer also allows it to be connected in a number of different ways, e.g. (a) two timers back to back, (b) two or more tandem, (c) three or more in a ring circuit.

The unit is housed in a small robust case and may be supplied with the time control adjustable by means of a knob on the front panel or a sealed screwdriver slot to prevent tampering.

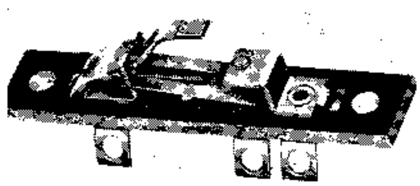


SNAP ACTION SWITCH

(Illustrated below)

Arcolectric (Switches) Ltd., Central Avenue, West Molesey, Surrey

THIS sensitive snap action switch, catalogue number T280, is of skeleton construction and is designed on the micro-gap principle. It is intended for building into equipment such as calculating machines, coin operated machines, process timers, etc. It is rated at 3A 250V a.c. and has a life expectancy of 1½ million operations. It has five silver contacts and a single-pole changeover action which returns to normal when pressure is released.



FLEXIBLE CASTING RESIN

Murphy Radio Ltd, Welwyn Garden City, Hertfordshire

MURPHY Radio have now placed on the market a formula which produces a slightly flexible casting resin, eliminating pressure on the encapsulated components and immune from cracking down to temperatures of at least − 70°C.

This material, under the registered trade mark 'Castoflex,' is a castor oil-urethane resin of excellent electrical properties. It is supplied in the form of thin liquid components in which the toxic hazards usually associated with the processing of urethanes have been eliminated.

Murphy Radio are manufacturing a range of transformers cast in this material. It is also being used for the protection of precision wire-wound resistors and silvered-mica capacitors, as well as the conventional "potting" of subminiature circuit assemblies for missiles.

Other applications for which 'Castoflex' is attractive are the protection of flexible delay lines and the moulding of cable terminations for low temperature.

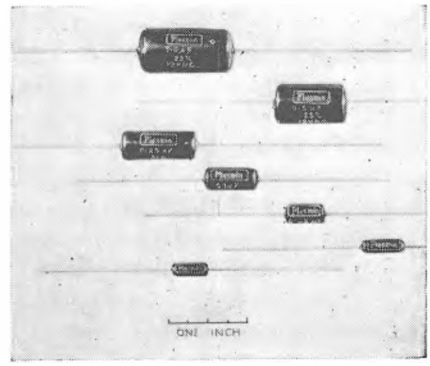
The maximum ambient temperature recommended is 120°C, but internal temperatures up to 150°C are permissible.

MINIATURE PAPER CAPACITORS

(Illustrated above right)

The Plessey Co. Ltd, Ilford, Essex

THE advent of the transistor with its low operating potentials and small size has necessitated reductions being made in the physical size of paper capacitors, without, as far as practicable, impairing their electrical performance. To meet this need The Plessey Company have recently introduced the 'Plesmin' range of paper capacitors. The small size and high standard of performance has been achieved by entirely new methods and processes of manufacture.



Physical sizes are considerably smaller than those made entirely by orthodox methods, the smallest capacitor of the new range measuring only ⅜in long and ⅛in diameter and the largest 1¼in long and 19/32in diameter. The range extends from 0·001μF up to 1·0μF with a normal tolerance of ±20 per cent for 0·01μF and above, but below this the normal tolerance is ±25 per cent.

Voltages normally covered by 'Plesmin' capacitors are 12, 25, 50 and 100V but working voltages up to 150V d.c. are also obtainable if ordered specially. An insulation resistance of 1kΩ/F minimum at 20°C at the rated d.c. working voltage can be obtained. Moreover, like metallized paper capacitors. They have self-healing properties if circuit parameters are suitable.

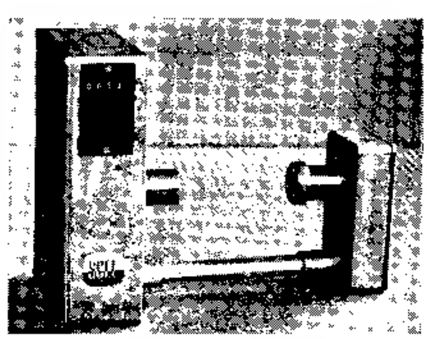
PHOTOCELL COUNTING UNIT

(Illustrated below)

Sutton Coldfield Electrical Engineers, Reddicap Trading Estate, Coleshill Road, Sutton Coldfield, Warwickshire.

THIS new range of counting equipment works at speeds up to 400 per minute and has been designed for easy installation.

The composite unit consists of three basic units, namely (1) relay unit, (2) cell holder and (3) light unit; thus provision is made for the remote fitting of any one of these items where necessary. In many instances the relay unit and cell holder can be combined as an integral unit to form a composite item as illustrated.



Short News Items

The Royal Society announces the awards of grants by the Paul Instrument Fund Committee, as follows. £2 700 to Professor G. F. J. Garlick, professor of physics in the University of Hull, to enable him to carry out studies of phosphors and photoconductors with a view to the development of solid state light image amplifiers or convertors. £2 000 to Dr. H. H. Hopkins, senior lecturer in physics (technical optics), Imperial College of Science and Technology, London, for continuation of the construction of a static scanning fibrescope, the aim of which is to provide a flexible unit capable of transporting optical images along its length. £100 to Dr. D. Gabor, Mullard reader in electronics, Imperial College of Science and Technology, making a total grant of £1 900 for work on a new type of Wilson cloud chamber in the form of a resonant microwave cavity in which time marks are applied to the particle tracks at equal intervals. The Paul Instrument Fund Committee, composed of representatives of the Royal Society, the Physical Society the Institute of Physics and the Institution of Electrical Engineers, was set up in 1945 'to receive applications from British subjects who are research workers in Great Britain for grants for the design, construction and maintenance of novel, unusual or much improved types of physical instruments and apparatus for investigations in pure or applied physical science'.

The Institution of Electrical Engineers' Radio and Telecommunications Section visit to the Netherlands took place from 18-22 September. Some 100 members attended, the Chairman of the section being Dr. R. C. G. Williams. There were visits to the Zuider Zee Reclamation Scheme (Eastern Flevoland); the Netherlands Testing Station for Electrical Materials (KEMA); Philips Research Laboratories and factories at Eindhoven; Philips Telecommunications Factory at Hilversum and Transistor factory at Nijmegen. A wreath was also laid in the Airborne Forces cemetry at Arnhem.

The BBC have sent their Head of Engineering Training Department, Dr. K. R. Sturley, to the United States to study methods of training in broadcasting and radio engineering. He will see the work of the chief American broadcasting networks in New York and Chicago. The training methods used in industry and selected technical institutions in Washington, Cleveland, and Boston, will be examined. Dr. Sturley will be presenting three technical papers written by other members of the BBC Engineering Division for the Audio Engineering Society's Annual Convention in New York and for the Society of Motion Picture and Television Engineers' Convention in Philadelphia.

The British Computer Society are pleased to announce that Dr. M. V. Wilkes, F.R.S., has accepted their invitation to become the first President of the Society. Dr. Wilkes was a pioneer of the design and construction of automatic electronic digital calculating machines in this country. Membership of the Society is open to anyone who has an interest in the use of electronic computers and computing techniques. Application may be made to the Secretary at 29 Bury Street, St. James's, London, S,W.1. Annual subscriptions are £3 3s. for Ordinary Members (with Entrance Fee of £1 1s.); Associated Members (aged 21 to 25) £1 1s; Institutional Members (corporate bodies, institutions, etc.) £10 10s. Details of meetings of the Society will be given in the Meetings this Month columns of Electronic Engineering.

The 4th International Instrument Show will be held at Caxton Hall, Westminster, London, S.W.1. from 24 to 29 March, 1958.

The next British Plastics Exhibition and Convention to be held at Olympia, London from 17-27 June 1959 is to be renamed the International Plastics Exhibition and Convention.

The British Electrical Conference 1958 will be held in Brussels on 16 and 17 May. Further information may be obtained from The Organizing Secretary, 1958 British Electrical Conference, 36 Kingsway, London, W.C.2.

Auckland University Engineers' Association has recently been formed, membership of which is open to members of staff, past and present, graduates and past students of the School of Engineering, Auckland University College, New Zealand. There are many who have passed through the School since its establishment in 1906 and cannot be traced. Those interested in further details should write to J. H. Percy, Secretary, Auckland University College, School of Engineering, Ardmore College P.O., Auckland, New Zealand.

The Southampton Harbour Board has awarded a substantial contract to Marconi's Wireless Telegraph Co Ltd for equipping the Port of Southampton Operation and Information Service with v.h.f./f.m. radio. The equipment will conform to the standards laid down at The Hague Convention of January last and will form the basis of a new system of port communications that may eventually have a world wide application. The order includes five 25W f.m./v.h.f. transmitters, Type HX86A, and five f.m./v.h.f. receivers, Type HR86A, together with associated remote control equipment. The transmitters and receivers are capable of operation on any one of six adjacent channels, any of which may be selected at will from a control centre sited at a convenient point in the area. Three seven-channel f.m./v.h.f. mobile equipment, Type HP86B, and two automatic mains/standby engine-alternator sets are also among the items to be supplied.

Aveley Electric Ltd have been awarded agency in this country for The Narda Corporation, Minneola, New York; Gosen of Erlangen, West Germany, and Electronic G.M.B.H. Munich.

Boots Pure Drug Co. Ltd have recently placed an order with E.M.I. Electronics Ltd. for an EMIDEC Electronic Data Processing System to be installed in their head office in Nottingham.

Sir Thomas Spencer, Chairman and Managing Director of Standard Telephones and Cables, has recently completed half a century of service to the telecommunications industry and to Standard Telephones and Cables Ltd. Internationally recognized in the field of telecommunications, Sir Thomas Spencer has for many years been a leading figure in the industry.

Mr. E. E. Jones, Group Commercial Director of The Solartron Electronic Group Ltd., recently left for a stay of some months in the United States. He will review the progress of Solartron Incorporated over the first year of its existence.

Errata. The following amendments should be made on page 506 of the October issue (The S.B.A.C. Exhibition). Under the Ultrasonic Generator of Dawe Instruments Ltd, for "total effective radiation surface 350^{2}", read " . . . $350in^2$". Weight should be 175lb, not 1 751 lb. Under the Double-Beam Oscillograph manufactured by A. C. Cossor Ltd, for " . . . not less than 2cm to peak . . .", read " . . . not less than 2cm peak to peak . . . ". For " . . measurement of time along the X axis are carried out in a similar manner to some accuracy", read " . . . measurement of time along the X axis is carried out in a similar manner to the same accuracy."

Notes from ——————

NORTH AMERICA

National Symposium on Reliability

The 4th National Symposium on Reliability and Quality Control will be held January 6, 7 and 8, 1958 at the Hotel Statler in Washington, D.C.

This Symposium will cover fields of reliability in the electronics industry, and will encompass the following topics: Reliability Organization and Management; Theory and Mathematical Techniques; Application of these Techniques; Design Information; Education and Training for Reliability.

Oscilloscope Shadow-Screen

This is a new oscilloscope accessory which, it is claimed, eliminates the need for subdued lighting or an oscilloscope hood.

Known as an oscilloscope shadow-screen, the new device is a thin sheet which fits directly in front of the scope face. The screen blocks off the ambient light that blurs the cathode-ray tube image when the instrument is used outdoors or in a brightly lit laboratory.

Manufactured by Van-Dee Products, Laguna Beach, California, the oscilloscope shadow-screen consists of hundreds of small, hexagonal openings that serve as individual shadow boxes, shutting off the glare that would otherwise hit the face of the cathode-ray tube. The shape of the openings, however, permits full observation of the scope image from any point within 45° of the face-on position.

The clarity of a cathode-ray tube image depends on both the brightness of the electron trace and the darkness of the surrounding glass. The oscilloscope shadow-screen acts by sharply reducing reflected light from the face of the tube. This greatly increases the contrast between the face and the electron image.

Shadow-screens are available to fit any standard 3in or 5in oscilloscope. Similar units are being designed for use with radar screens.

Automatic Circuit Assembly

An automatic assembly machine which takes orders from itself was announced at the end of last month by IBM Corporation in New York. The new device is used to fasten resistors, capacitors, diodes, and the like—automatically—on to printed circuit panels. When a change in the layout and components of a panel is called for, the machine gets instructions from IBM cards and then automatically resets itself for the new assignment.

The new device, the Programmed Component Assembly System, was developed and built by manufacturing engineers at IBM's Poughkeepsie (U.S.) plant and currently is at work assembling wiring panels for the Company's data processing equipment. IBM engineers say assembly speed is better than ten times faster than is possible by manual methods and a more uniform product is obtained.

The system was built for the Company's own use, but IBM's new Special Engineering Products Division is considering making similar machines available to industry. The system now in use at Poughkeepsie inserts one component every second and a half. The proposed market version of the system is a one-a-second machine with probable price in the \$100 000 range. IBM believes the extreme simplicity and economy of the changeover operation makes the new system particularly adaptable to low production runs of a great variety of board assemblies.

Key to the Programmed Component Assembly System's successful design and operation was IBM's decision to standardize many characteristics of printed panels as well as dimensions of electronic components. Now, for example, all components are packaged in one of two ranges of size regardless of their individual characteristics. Prior to this, the great variety of shapes and sizes of these components resisted automated assembly methods.

Components are grouped according to their electrical values, mounted on masking tape 'belts', and wound on 'cut-off' reels, much in the same way that machine gun bullets are loaded in drums. Several thousand components are stored on each reel in this fashion. The system has a capacity of twenty reels, although it can be expanded to accommodate many times that number if necessary. The reels are set in place on the selection 'rack' and the lead-taped components are fed into cut-off stations.

In order to attach a component on a printed panel, three instructions must be provided simultaneously by the system's card reader. Two columns of an IBM card contain the three instructions needed to insert any one component. The first tells the device which one of twenty different stored components to select, cut-off from its reel, and deliver to the insertion system. The second and third instructions position the panel so that the selected holes are accurately aligned under the insertion mechanism. When all three instructions have been obeyed, the insertion system automatically inserts the selected component and staples it to the panel. This process continues repeatedly until all instructions have been followed and the printed panel is completely assembled.

The Programmed Component Assembly System will accommodate any size panel up to 10in square and will operate to accuracies of better than 0·002 of an inch.

Solid State Battery

The Patterson, Moos Division of Universal Winding Company has developed a miniature, high voltage, solid state battery that is capable of delivering higher current drains for longer periods of time than other comparable batteries.

Called the 'Dynox 95', the first model of the new battery is $1\frac{1}{4}$in long, $\frac{5}{8}$in in diameter, and has a potential of 95V in 0·14in^3. It can supply a steady current of 1×10^{-9}mA for 176 000 hours at 70°F. with only a 10 per cent voltage drop, or a flash current of 20μA. Also, it can be stored for 20 years without losing its power.

Other characteristics include resistance to extreme shock or vibration, performance at extreme temperature ranges, including the ability to operate after temperature cycling between −100 and 170° F. It is hermetically sealed, which makes it resistant to the effects of humidity. Batteries can be assembled to meet many special dimensional and electrical requirements. For example, batteries that can furnish 625V/in^3 and can be drained continuously for 5 years at 5×10^{-9}A are also available.

Semiconductor Prices

Prices are lowered as much as 22 per cent on silicon transistors and 35 per cent on silicon glass diodes according to an announcement made today by Texas Instruments, the nation's first and largest manufacturer of silicon transistors.

This is the second significant silicon transistor price reduction made by TI during 1957. Last January Texas Instruments reduced the prices on silicon transistors and rectifiers purchased in large quantity lots.

In addition, Texas Instruments also announced that its warranty on its entire line of silicon diodes and rectifiers is now extended to a full year instead of the 90-day warranty previously in effect. This means that the company now guarantees the electrical characteristics of all semiconductor products—silicon and germanium transistors, silicon diodes and rectifiers—for a full year starting with the date of delivery.

Silicon transistor prices to original equipment manufacturers in 100 and up quantities now begin at \$6·90 for small signal devices, \$14·05 for medium power units, \$15·50 for switchers, \$40·00 for power semiconductors, and \$16·00 for high frequenty transistors.

Quantity prices of TI's 400mA and 225 to 600V glass diode line now start at \$2.10 to manufacturers of original equipment.

Change of Name

The Narda Corporation, Mineola, New York, has changed its name to The Narda Microwave Corporation. The new name was chosen to better reflect Narda's greatly increased activities in the microwave and u.h.f. test equipment field, such as the recent acquisition of Kama Instrument Corporation, and a general expansion of the company's research and production facilities.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 27 November. Time: 6 p.m.
Annual General Meeting, followed by
Lecture: Transmission Standards and Signal Distortion in Television and other Communication Systems.
By: A. van Weel.

West Midlands Section

Date: 13 November. Time: 7.15 p.m.
Held at: The Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Cold Cathode Switching Techniques.
By: J. H. Beesley.

North Eastern Section

Date: 13 November. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Lecture: Electronic Control of Machine Tools.
By: H. Ogden.

South Midlands Section

Date: 1 November. Time: 7 p.m.
Held at: The Winter Gardens, Malvern.
Lecture: Marine Radar.
By: H. R. Whitfield.
Date: 29 November. Time: 7 p.m.
Held at: The North Gloucestershire Technical College, Cheltenham.
Lecture: Automatic Programming.
By: P. M. Woodward.

THE BRITISH COMPUTER SOCIETY

Date: 18 November. Time: 6.15 p.m.
Held at: L.C.C. County Hall, London, S.E.1.
Lecture: Applications of a Computer to the work of Norwich Corporation and Plans for Future use.
By: A. J. Barnard.

THE INSTITUTION OF NAVIGATION

Date: 15 November. Time: 5.15 p.m.
Held at: The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.
Lecture: Doppler Navigation.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution commencing at 5.30 p.m.

Ordinary Meeting

Date: 7 November.
Discussion of the report on The Supply and Training of Teachers for Technical Colleges.
Introduced by: Professor Willis Jackson.
Date: 28 November.
Lecture: The Digital Computer Applied to the Design of Large Power Transformers.
By: W. A. Sharpley and J. V. Oldfield.

Measurement and Control Section

Date: 5 November.
Lectures: The Design of the Control Unit of an Electronic Digital Computer.
By: M. V. Wilkes, W. Renwick and D. J. Wheeler.
A Decimal Adder using a Stored Addition Table.
By: M. A. Maclean and D. Aspinall.
An Accurate Electroluminescent Graphical Output Unit for a Digital Computer.
By: T. Kilburn, G. R. Hoffman and R. E. Hayes.

Radio and Telecommunication Section

Date: 13 November.
Lectures: Broad-Band Slot-Coupled Microstrip Directional Couplers and: Re-Entrant Transmission Line Filter using Printed Conductors.
By: J. M. C. Dukes.
Date: 25 November.
Informal Evening: Problems of Sound and Television Broadcasting Coverage.
Talk by: G. Millington.

Cambridge Radio and Telecommunication Group

Date: 12 November. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: Some Radio Aids for High-Speed Aircraft.
By: J. S. McPetrie.

East Anglian Sub-Centre

Date: 4 November. Time: 6 p.m.
Held at: The Crown and Anchor Hotel, Ipswich.
Lecture: Colour Television.
By: L. C. Jesty.
Date: 19 November. Time: 8 p.m.
Held at: The Cavendish Laboratory, Cambridge.
Lecture: Earth Electrode Systems for Large Electric Stations.
Date: 25 November. Time: 7.30 p.m.
Held at: The Assembly House, Norwich.
Lecture: Germanium and Silicon Power Rectifiers.
By: T. H. Kinman, G. A. Garrick, R. G. Hibberd and A. J. Blundell.

North-Eastern Radio and Measurements Group

Date: 4 November. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Informal Evening: The Importance of Research in Hearing and Seeing to the Future of Telecommunication Engineering.
By: E. C. Cherry.
Date: 18 November. (Time and place as above).
Lectures: An Analogue Computer for Nuclear Power Studies and: The Application of Analogue Methods to Compute and Predict Xenon Poisoning in a High-Flux Nuclear Reactor.
By: G. J. R. MacLusky.

North Midland Centre

Date: 12 November. Time: 6.30 p.m.
Held at: The Technical College, Bradford.
Discussion: The Teaching of Radio and Television Servicing.
Opened by: G. N. Patchett.

North-Western Centre

Date: 5 November. Time: 6.15 p.m.
Held at: The Engineers' Club, 17 Albert Square, Manchester.
Lecture: 138kV Submarine Power Cable Interconnexion between the Mainland of British Columbia and Vancouver Island.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

North Scotland Sub-Centre

Date: 14 November. Time: 7 p.m.
Held at: The Electrical Engineering Department, Queen's College, Dundee.
Sub-Centre Chairman's Address.
By: E. G. Cullwick.
Date: 15 November. Time: 7.30 p.m.
Held at: Robert Gordon's Technical College, Aberdeen.
Sub-Centre Chairman's Address.

South-East Scotland Sub-Centre

Date: 27 November. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Servo-Operated Recording Instruments.
By: A. J. Maddock.

South-West Scotland Sub-Centre

Date: 6 November. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lecture: Recent Developments in Low-Voltage High-Rupturing-Capacity Fuse Links.
By: R. H. Dean.
Date: 12 November. (Time and place as above).
Lecture: 138kV Submarine Power Cable Interconnexion between the Mainland of British Columbia and Vancouver Island.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.
Date: 26 November. Time: 7 p.m.
Held at: The Royal College of Science and Technology, George Street, Glasgow, C.1.
Lecture: Servo-Operated Recording Instruments.
By: A. J. Maddock.

South Midland Centre

Date: 4 November. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Film entitled: The Construction of the Super Grid.
Introduced by: H. S. Davidson.
Date: 14 November. Time: 6 p.m.
Held at: The Midland Institute, Birmingham.
Lecture: Some Engineering Problems in the Industrial Development of Nuclear Energy.
By: Sir Claude Gibb.
(Joint meeting with the Midlands Association of The Institution of Civil Engineers and the Midland Branch of The Institution of Mechanical Engineers). (Admission by ticket).
Date: 25 November. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Magnetic Materials.
By: Professor R. Brailsford.

South Midland Radio and Measurements Group

Date: 25 November. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Magnetic Materials.
By: Professor R. Brailsford. (Joint meeting).

North Staffordshire Sub-Centre

Date: 18 November. Time: 7 p.m.
Held at: Duncan Hall, Stone.
Lecture: 138kV Submarine Power Cable Interconnexion between the Mainland of British Columbia and Vancouver Island.
By: T. Ingledow, R. M. Fairfield, E. L. Davey and K. S. Brazier.

Rugby Sub-Centre

Date: 6 November. Time: 6.30 p.m.
Held at: The Rugby College of Technology and Arts.
Lecture: Financial Structure of the Electrical Manufacturing Industry.
By: The Hon. Alexander L. Hood.
Date: 26 November. (Time and Place as above).
Lecture: Transistor Circuits and Applications.
By: L. P. Morgan.

Southern Centre

Date: 6 November. Time: 7 p.m.
Held at: Southampton University.
Lecture: Frequency Modulated v.h.f. Transmitter Technique.
By: A. C. Beck, F. T. Norbury and J. L. Storr-Best.
Date: 20 November. Time: 6.30 p.m.
Held at: The S.E.B., Salisbury.
Lecture: The Control of Nuclear Reactors.
By: R. J. Cox and J. Walker.
Date: 29 November. Time: 6.30 p.m.
Held at: The South Dorset Technical College, Weymouth.
Lecture: Equivalent Circuits of Transistors and their Application.
By: L. E. Jannson.

West Wales (Swansea) Sub-Centre

Date: 14 November. Time: 6 p.m.
Held at: The Conference Rooms, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: The Measurement of Earth-loop Resistance.
By: G. F. Tagg.

Reading District

Date: 25 November. Time: 7.15 p.m.
Held at: The George Hotel, King Street, Reading.
Lecture: Cathodic Protection.
By: J. H. Gosden.

THE INSTITUTION OF ELECTRONICS

Date: 28 November. Time: 6.30 p.m.
Held at: The Assembly Hall, University of London Institute of Education, Malet Street, London, W.C.1.
Lecture: The High Fidelity Reproduction of Sound.
By: C. Brown.

Northern Ireland

Date: 19 November. Time: 7.30 p.m.
Held at: The Union Hotel, Belfast.
Lecture: A Low Frequence Decade Oscillator.
By: A. W. S. Gilham.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 12 November. Time: 5 p.m.
Held at: Waterloo Bridge House, London, S.E.1.
Lecture: Recent Developments in the Cabling and Wiring of Telephone Exchanges.
By: T. F. A. Urben.

THE TELEVISION SOCIETY

Date: 15 November. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Industrial Television.
By: I. M. Waters.
Date: 29 November. (Time and place as above).
Lecture: Some Aspects of Waveguide Technique.
By: J. C. Parr.