ELECTRONIC ENGINEERING

VOL. 31 No. 377

JULY 1959

Commentary

THE National Physical Laboratory has recently held its annual 'open days' and, coinciding with these, it has published its report for 1959. Although these open days are well attended there are a great many engineers who do not take advantage of the welcome that the NPL extends to them and it is, indeed, an opportunity which many, and particularly those engaged in industry, would find both useful and instructive. For those who cannot attend the Laboratory then time spent browsing through the Annual Report would not be time wasted.

Indeed, there are, probably, a considerable number of engineers who are only vaguely acquainted with the general organization and activities of the Laboratory and who tend to regard it merely as an organization for the maintenance of standards and the certification of meters and thermometers. This, in fact, was the avowed object of the Laboratory as set out by a Treasury Committee when its formation was first considered. However, at the official opening of the Laboratory in March 1902 the late King George V, then Prince of Wales, said, "The object of the scheme is, I understand, to bring scientific knowledge to bear practically upon our everyday industrial and commercial life; to break down the barrier between theory and practice; to effect a union between science and commerce". While by no means forsaking the objectives of the Treasury Committee it is very much along the lines set out by His Royal Highness that the NPL has developed. From its inception until 1918, when it became part of the newly formed Department of Scientific and Industrial Research, the Laboratory was controlled by the Royal Society and this body still appoints a General Board and an Executive Committee to supervise the scientific aspects of the work; the President of the Royal Society being the *ex officio* Chairman of the General Board. The Executive Committee consists of nominees of the Royal Society and the principal scientific and technical institutions whose interests lie in the same fields as the work of the Laboratory. When the Laboratory first came into being it had a staff of 31 and was organized under two divisions, Physics and Engineering. It now has a staff of some 1 100 persons organized under ten divisions. The general organization is, of course, continually changing to meet the needs of the day. For instance, a few years ago it was deemed that the Engineering Division and the Radio Division had outgrown, so to speak, their old homes at the NPL, as a consequence of which separate research stations were set up by DSIR; the Mechanical Engineering Research Laboratory at East Kilbride and the Radio Research Station at Slough. This in turn made room for the growth of two new divisions at the NPL; Control Mechanisms and Electronics and Mathematics. As recently as last year further re-organization took place and the Electricity, Metrology and Physics Divisions have been superseded by three new divisions; Applied Physics, Basic Physics and Standards.

While the formulation and maintenance of the fundamental standards of mass, length and time and of the electrical units and temperature etc., remains an extremely important section of the Laboratory's work, over 80 per cent of its effort is now devoted to basic and applied research and, although the NPL must necessarily choose its subjects carefully and concentrate on a comparatively small number of objectives, it covers, in total, a very diverse field of activity and the results of its efforts are applicable to a large range of industries. Indeed, it provides the country with a unique concentration of scientists who can tackle almost any problem involving the application of non-nuclear physics. As an example of this, the Laboratories recent work has covered such widespread subjects as the development, improvement and exploitation of electronic computers, the evolution of better steels for power stations and the improvement of ship and aircraft design. In a great deal of this research the work is directed towards elucidating the basic physical principles underlying key problems and, in this connexion the NPL can accept contracts from industry for pure and applied research in fields where it has unusually good facilities and experience.

The results of the work at the Laboratory and the many discoveries which are made there are, in general, available to industry and it is, therefore, very much in the interest of the engineer engaged in industry for him to take advantage of the facilities which the Laboratory provides for showing and making known its work. A typical example of this is the production of Merton gratings at the NPL which have subsequently been used successfully for the control of machine tools by commercial firms.

To summarize, one may fairly say that the National Physical Laboratory offers industry a considerable service, excellent research facilities and a concentration of scientific manpower. It is up to industry to make the utmost use of what is offered, for to do so cannot but reflect favourably on the individual firms and also on the country's scientific effort as a whole.

An Image Convertor Equipment for the Observation of Self-Luminous Gaseous Discharges

By K. G. Beauchamp* and A. J. Tyrrell†

A high speed camera unit using a pulsed image convertor tube is described. Methods of gating this tube at potentials up to 20kV are detailed and the advantages of deflecting the electronic image across the tube face are discussed.

A considerable reduction in the bulk and complexity of an image convertor equipment is made possible by using a permanent magnet for focusing the tube and the equipment described includes a focusing unit of this type.

(Voir page 440 pour le résumé en français; Zusammenfassung in deutscher Sprache Seite 445)

RECENT work on controlled thermonuclear reactions[1,2] has necessitated the detailed observation of hot gaseous discharge phenomena, using exposure times of as little as 10^{-7}sec. Such methods as that of the Kerr cell shutter[3] have been in use for a considerable period, but suffer from an inherent disadvantage of poor sensitivity in the spectral region of greatest interest. This is a serious limitation when used for gas discharge work owing to the small light output available at these extremely short exposure times.

The image convertor tube allows a high output efficiency to be obtained, and in fact if a demagnification of the image is permissible then a considerable brightness gain

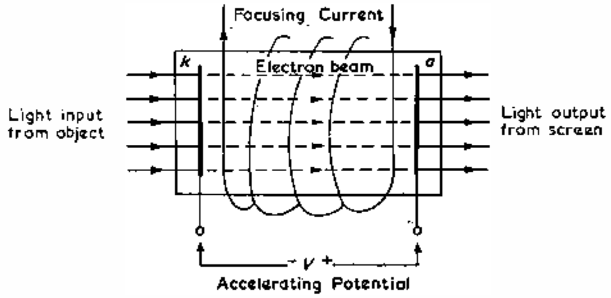


Fig. 1. Principle of image convertor tube

may be obtained, since the screen brightness is proportional to the inverse square of the image magnification. This device has found widespread application in rendering infrared images visible to the eye[4] and also as an X-ray image intensifier[5]. The use of the convertor as a high-speed photographic shutter is, however, possible, and arises from the fact that the electron beam, constituting the intermediate electron image, may be easily controlled and that such control involves extremely little inertia.

The basic principle of the diode image convertor tube is shown in Fig. 1. An optical image of the object is focused on to a semi-transparent photo-cathode *k*. The electrons emitted from the photo-cathode are focused by means of a suitable electron-optical system upon a fluorescent screen *a*, forming part of the anode of the diode tube. A high accelerating potential is applied between anode and cathode, the control of which allows the tube to be operative for the desired exposure period. This control can be accurately synchronized with the discharge under investigation by means of well-known triggering and delay techniques.

The Image Convertor Tube

The tube used in the equipment was the MS 1701 unity magnification magnetically focused diode especially developed for high-voltage operation (Fig. 2). The caesium-antimony photo-cathode has a predominantly blue sensitivity as is shown in the spectral sensitivity curve shown in Fig. 3. This spectrum includes several of the more important spectral lines of the gas under investigation and allows a fairly high conversion efficiency to be realized. A short persistence blue fluorescent screen is incorporated having a diameter of approximately 70mm.

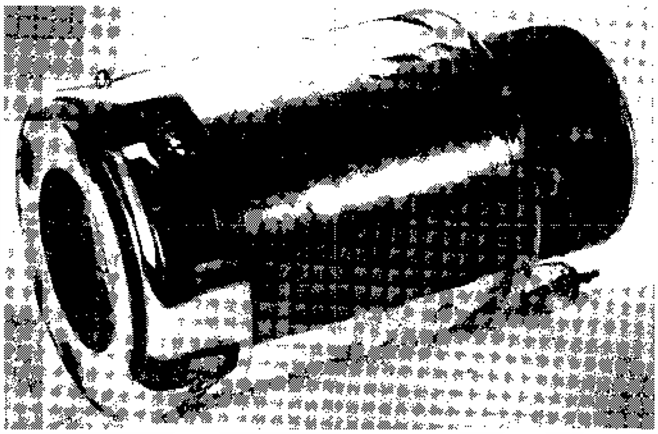


Fig. 2. The image convertor tube

The utilization factor of this screen diameter is fairly good, and this may be seen from the reproduced image shown in Fig. 4. This represents a light dot pattern occupying an area of 30mm square on the screen and was obtained under d.c. operation conditions at an applied accelerating potential of 10kV. The predominant forms of distortion experienced with image-convertor tubes are pin-cushion and *S* bend distortion[6], the latter showing itself as a variable lateral displacement of the image over the screen area imposing an *S*-shape representation of linear images falling on the photo-cathode. This displacement may be expressed as a percentage variation of a maximum, which may be taken as half a picture width. Under these conditions the measured distortion of this image was about 1 per cent, thus allowing an essentially distortion-free picture to be obtained over the central half of the photo-cathode.

The image obtained with the lowest focusing magnetic field suffers from considerable distortion and it is usual to make use of the 2nd or 3rd focus positions with this type of tube. At the second focus, the ampere-turns required for focusing are approximately 8 500 at 20kV, so that the provision of a source of stabilized focusing current represents a fairly large power requirement. In addition, the power dis-

* *formerly with Associated Electrical Industries Ltd.*
† *Mullard Ltd.*

sipated by the copper losses experienced within the coil presents a difficulty since operation of the tube at an ambient temperature in excess of 50°C may damage the photo-cathode. In view of the foregoing it was decided to attempt focus of the tube using a permanent magnet focusing system, a description of which is given later.

A mechanical layout was found convenient for the completed equipment in which the image convertor, focusing magnet, sweep coils, etc., were located in a separate camera head unit, while the trigger circuits, power supplies, etc., were housed in an enclosed rack mounting. This is illustrated in Fig. 5.

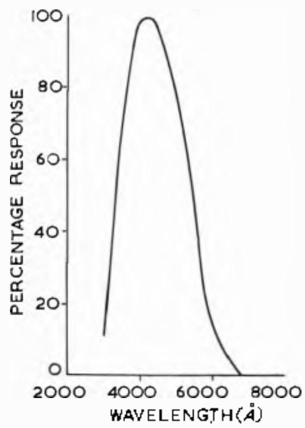


Fig. 3. Spectral response of image convertor tube

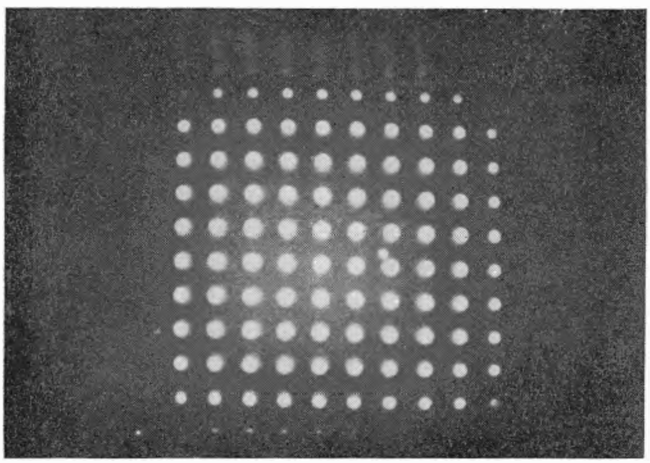


Fig. 4. Light dot pattern reproduced on the screen of the image convertor tube

Focusing

Essentially the problem is merely to immerse the tube in a uniform field of approximately 30mWb/m² parallel to the geometric axis of the tube. The physical limitations imposed by the optical system resulted in some modifications to the ideal design and this was finalized in the form of a hollow ellipsoid with disk shaped iron pole pieces at the ends. These pole pieces, having the minimum size hole compatible with the optical requirements, serve to form equipotential planes between which a uniform field is maintained.

The design was so arranged that the magnetic elements would form structural members of the assembly and also serve to shield the tube from external interference. The magnetic system is symmetrical and circular in section; a central steel plate forms the basis of a sub-assembly having two hollow Ticonal shells bolted to it. Of the two iron pole pieces, one forms part of the camera system and the other forms part of the viewing system. The object of this design is to make use of the fact that the field inside a hollow and correctly designed shell of magnetic material is theoretically uniform and parallel, and with this form of design, external interference is negligible.

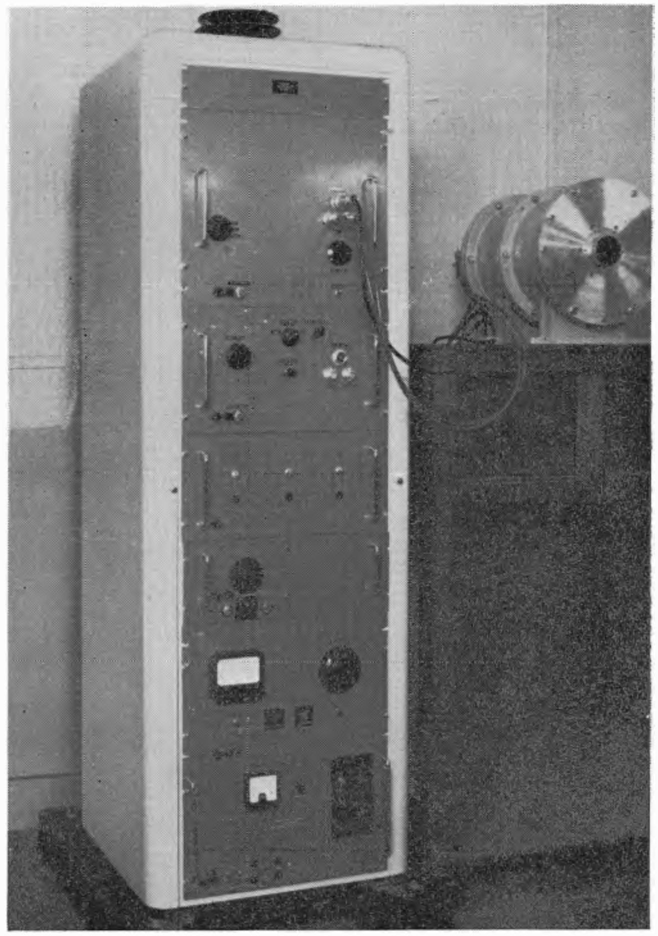


Fig. 5. The complete image convertor equipment

In the near vicinity of the apertures in the pole plates non-uniformity occurs, but these regions are, in this design, spaced sufficiently far from the image convertor that the effect of the apertures is negligible and virtually undetectable.

The completed magnet was calculated to give not only the required field uniformity but a field strength appreciably in excess of the requirements of the image tube. However, for assembly purposes it is convenient to pre-magnetize the sections of permanent magnet and if possible finally assemble the equipment without a further magnetizing or demagnetizing operation.

This technique, when applied to many magnet systems, generally results in the final performance being in the region of 25 per cent of that to be obtained if magnetization of the completed assembly is carried out. In this design it was estimated that even when assembling pre-magnetized magnets the resultant field should be adequate for operation and lie in the region of 30mWb/m². This, in fact, proved to be the case, and in operation the permanent magnet focusing field was sufficient to ensure that the magnification of the image convertor was less than 0·7 resulting in a light gain of two-fold over unity magnification. This method of design results in the unit being extremely stable even under severe external disturbing magnetic fields; at the same time it is of a sound and practical engineering construction.

Triggering and Gating Arrangements

The experimental requirements indicated that a wide range of exposure times would be necessary, ranging from 10^{-7}sec to 5×10^{-5}sec. In order to prevent de-focusing of the image during the operating period of the convertor, the anode-cathode potential must be maintained constant (i.e., the gating pulse must be flat-topped).

A way of achieving this, which is especially suitable for the short exposure times, has been proposed by Dr. R. F. Saxe. This uses two thyratrons in the manner shown in

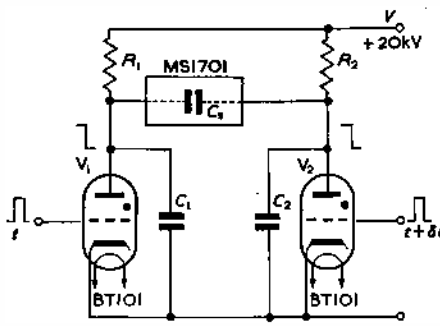


Fig. 6. Principle of thyratron trigger circuit

Fig. 6. Before the application of a triggering pulse the thyratrons are non-conductive and both cathode and anode of the convertor are raised to the supply potential, V volts, and no electron flow occurs through the tube. The tube is triggered when thyratron V_1 is caused to discharge capacitor C_1. This reduces the tube cathode to zero potential and the convertor becomes operative. The accelerating potential is removed by the conduction of thyratron V_2, which occurs δt μsec later and reduces the tube anode to zero potential. The thyratron time-constants C_1R_1 and C_2R_2 are so arranged that the cathode of the tube recovers to h.t. potential first, thus ensuring that the tube remains inoperative after the exposure period.

For the longer exposure times these time-constants would become rather large, and in the event of flash-over taking place within the tube, then the energy storage in C_1 and C_2 would be sufficient to cause permanent damage to the photo-cathode structure. This places a limitation on the maximum size of these capacitors and leads to the use of radar-type hard-valve pulse generators for exposure periods in excess of 1μsec, where no energy-storage problems are experienced. The division of the gating requirements into these two methods for short and long exposure periods enables a flexible arrangement to be obtained and optimum performance secured from each unit.

THYRATRON GATING UNIT

The anode-cathode system of the tube possesses a finite capacitance, and when wiring capacitance is added to this a total figure of about 100pF is obtained. The charge and discharge of this capacitance, C_s, will be obtained from C_1 and C_2, and in order to prevent a change in gating potential during this period, these capacitors will have to be several times greater than C_s. This variation in charge distribution between C_1/C_2 and C_s will not be so great as would be expected from a complete short-circuit of the discharging capacitor, and a fraction of the thyratron current will be diverted to charge C_s. An arbitrary value of $10C_s$ for C_1 and C_2 was chosen and proved adequate to maintain the tube potential constant during the gating period.

In order to charge C_s from a potential of 20kV in, say, 10 per cent of the smallest gating period, i.e., 10^{-8}sec, currents of the order of 200A will be required. However, since the current intensity gradually decreases as the capacitance becomes more nearly charged, the actual peak current requirement will be considerably greater than this. This peak current must be limited by a series resistance to a

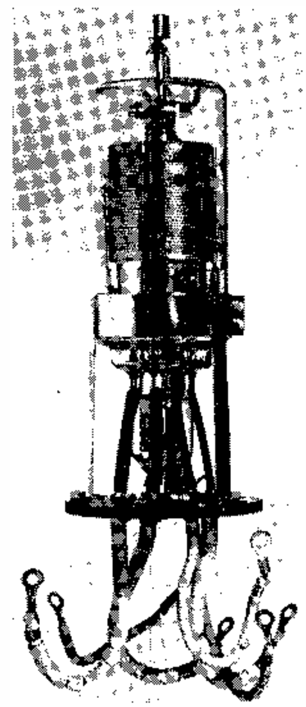


Fig. 7. Type BT101 hydrogen thyratron

value determined by the rating of the thyratrons used. In the gating unit described below, the valves used were B.T.H. hydrogen thyratrons type BT101 (Fig. 7) and resistors were included to limit the discharge to 500A. These thyratrons have a maximum measured 'jitter' of 10^{-8}sec, which is adequate for the shortest exposure time required.

The general arrangement for the unit is shown in Fig. 8. The incoming trigger pulse turns on a small thyratron V_1 which in turn triggers a blocking oscillator V_{2a}, supplying the initiating pulse required by hydrogen thyratron V_5 via a cathode-follower V_{2b}. Simultaneously a pulse is conveyed from V_1 through a delay line, to a second trigger thyratron V_3. This triggers the second hydrogen thyratron V_6 via blocking oscillator circuit, V_4, after a time interval dependent on the characteristics of the delay line used.

Three gate periods are obtainable from this unit, 0·1, 0·5, and 1·0μsec, and in order to simplify the arrangement and reduce the possibility of inductive pick-up occurring in the delay cables, special high-impedance cables were used providing a large delay per unit length. The 0·1μsec delay was obtained using 1·9kΩ Telcon cable, giving a delay of approximately 8·5μsec/100ft, while for the 0·5 and 1·0μsec delay periods, a 4kΩ Hackethal cable was used having a delay of approximately 1μsec/ft.

A peak trigger potential of 500V is required for V_5 and V_6 and is obtained from the blocking oscillator circuit at an impedance level of 200Ω, and having the desired rate of rise of 2kV/μsec specified by the manufacturers.

Care was necessary to ensure that:

(a) A non-inductive path was provided for the discharge of capacitors C_1 and C_2.

(b) The large circulating current set up by the conductors of V_5 did not prematurely fire the second thyratron circuit V_3/V_4. This was largely a question of layout and suitable choice of components, although coupling via supply potentials is possible and to prevent this simple filter circuits have been included in bias and h.t. supply circuits.

Monitor points are shown in Fig. 8, the gate pulse being checked by using a capacitive potential-divider connected at each anode of V_5 and V_6.

Measured overshoot in the gating pulse was less than 1 per cent and no difficulty due to premature firing was experienced. Fall time of the thyratron anode potentials from a maximum of 20kV was also measured and found to be approximately 60mμsec, the thyratron currents being limited to a peak of 500A. Triggering delays are, of course, additive at each stage, but the overall delay between initial triggering and gating of the convertor tube was found to be less than 0·25μsec.

Hard-valve Pulse Unit

For longer exposure times a conventional radar-type pulse modulator was used to provide a negative-going gating pulse at the cathode of the convertor tube. The circuit of this unit is given in Fig. 9 and consists of a small trigger thyratron V_1 which initiates a monostable multivibrator V_3 and V_4, thereby controlling the width of the gating pulse applied via V_5 to the grid of the final output valve V_6. This latter is a high-powered pulse modulated

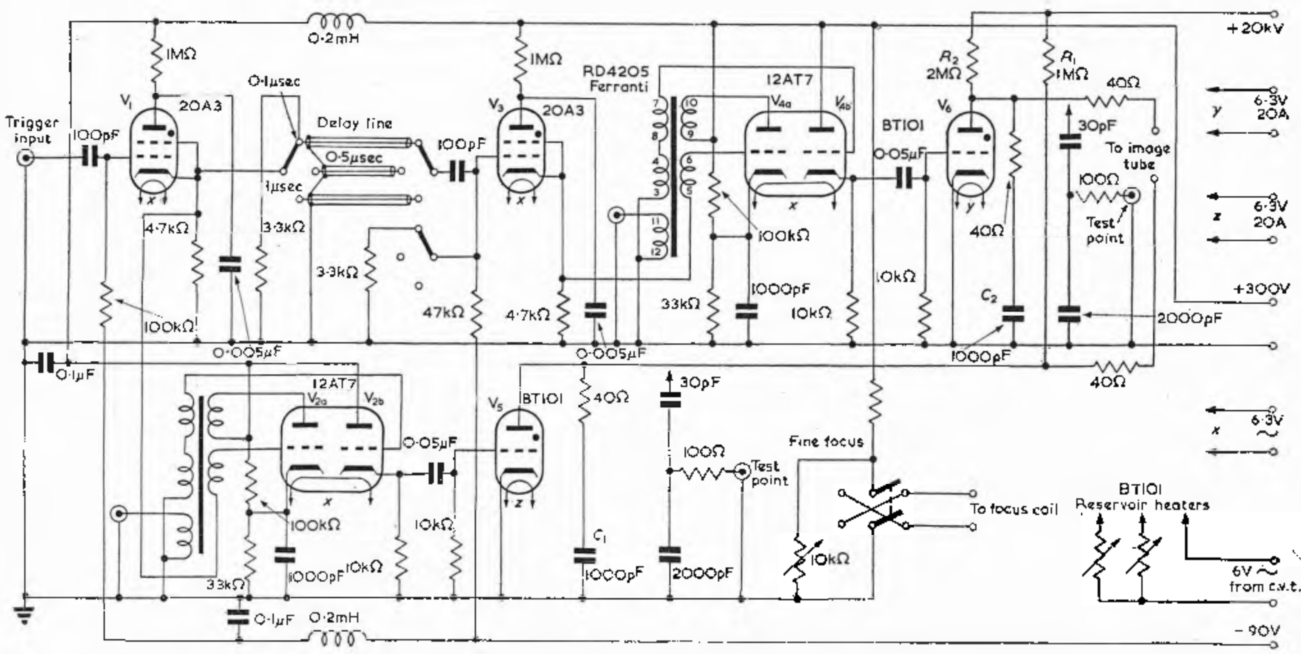
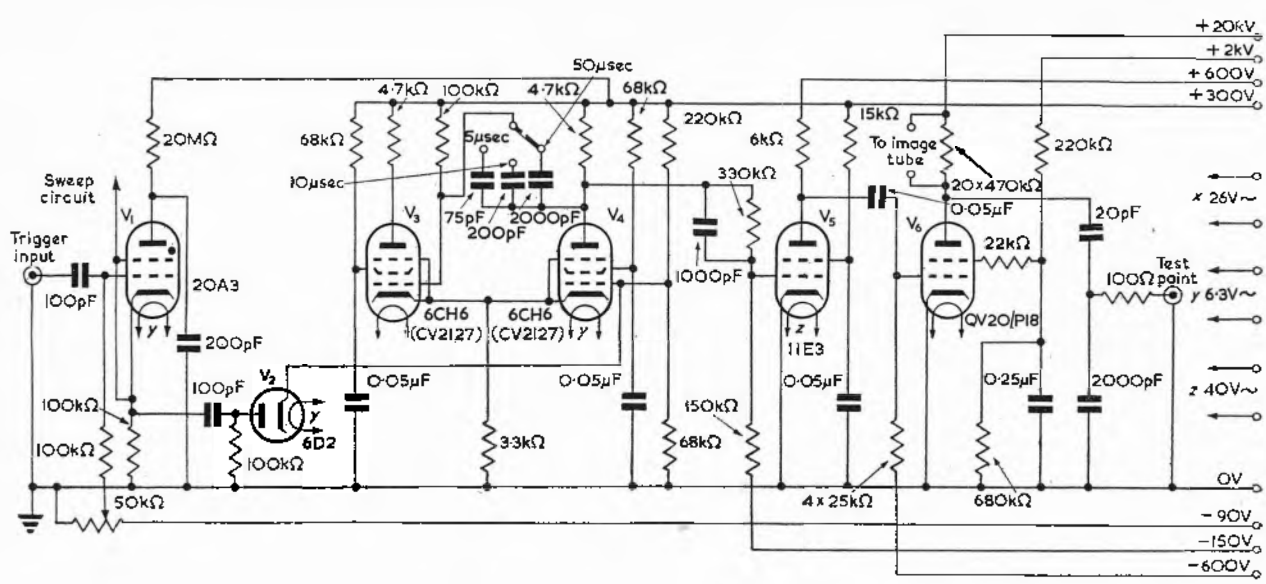


Fig. 8. Thyratron gating unit

Fig. 9. Hard valve pulse unit



tetrode capable of supplying the peak current required to charge C_s.

Three pulse widths are provided, namely 5, 10 and 50μsec and provision is included for monitoring the output pulse. Measurements gave the fall- and rise-times of the negative pulse as 0·5 and 1·0μsec respectively and an overall delay between initial triggering and production of gating pulse of 0·2μsec.

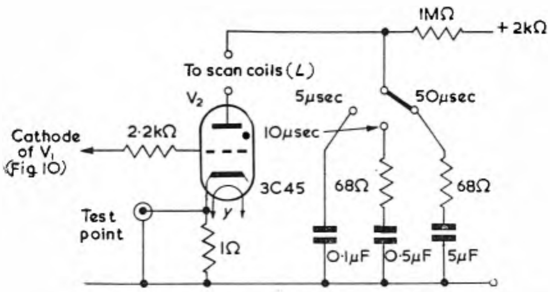


Fig. 10. Sweep circuit

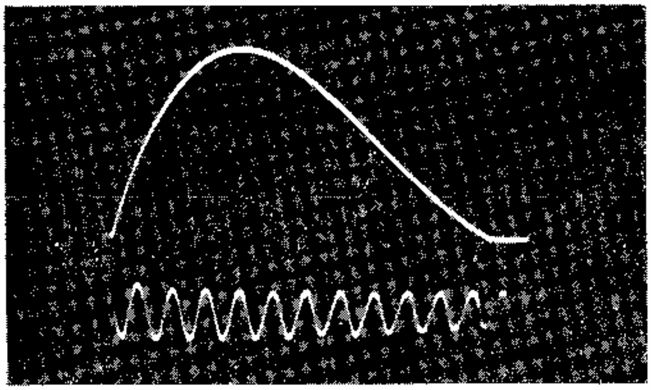


Fig. 11. Sweep current waveform, 10μsec sweep duration
(Calibration waveform = 100kc/s)

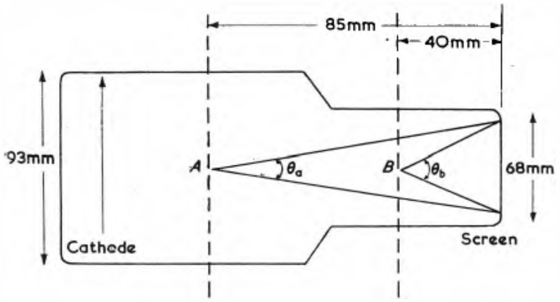


Fig. 12. Two positions for sweep coil

Sweep Facilities

It was considered that the information supplied by the instrument would be considerably enhanced in value if provision were made for a continuous cinematographic record of moving events. The difficulties arising if a series of individual frames of high repetition rate are to be recorded during a single exposure have been previously described by Chippendale and Jenkins[7]. The problem is, however, greatly simplified if the information desired is only required in one linear dimension. This proved to be the case for the problem under investigation; by providing an external field to sweep the electron image across the fluorescent screen at a constant rate, 'streak' photographs are obtained which can yield valuable information as to the variation of the observed phenomenon during the exposure period.

As this facility was only required for the longer exposure periods it was found convenient to include the appropriate circuits in the hard-valve pulse unit. The simple method used is shown in Fig. 10. A triggering pulse taken from V_1 in Fig. 9, is applied to the grid of a small hydrogen thyratron V_2. This allows the charge stored in C to be transferred to scanning coils, L, included in the discharge path. A single half cycle of resonance between coil and capacitor takes place, of which only the first 20 per cent is used. It will be seen from Fig. 11 that under these conditions a fairly linear rising sweep current is obtained. The system is to some extent wasteful of the peak current capabilities of the sweep valve, but has the merit of simplicity.

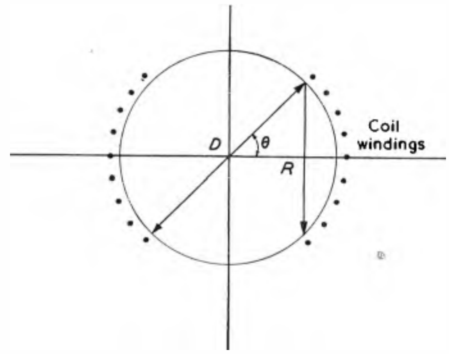


Fig. 13. Winding distribution

Sweep Coil Design

Two positions are possible for the sweep coils used with the convertor tube. These are shown in Fig. 12 and designated A and B for coils situated near the cathode and anode (screen) respectively. Using air-cored coils a figure for the ampere-turns required is given as:

$$IN = \frac{l \,.\, D \,.\sqrt{V}}{0{\cdot}38\, L \,.\, b.}$$

where V = applied accelerating potential (volts)
l = length of magnetic path = 30cm
b = axial length of coils = 4cm
D = diameter of bulb = 9·3cm (position A)
= 6·8cm (position B)
L = length from field centre to screen
8·5cm (position A)
4·0cm (position B)

Substituting figures for position A at an accelerating potential of 10kV then the ampere-turns required will be 1 581. This figure is about $\frac{2}{3}$ of that for similar coils situated at alternative position B.

The construction of suitable sweep coils is very dependent on the geometry of the tube. With a flat-faced screen, pin-cushion distortion would be apparent if a uniform magnetic field were used. In order to ensure that the m.m.f. is proportional to L at all angles θ (see Fig. 13) a distributed winding must be used, such that the total number of turns from the base to θ is proportional to $\sin\theta$. A simple grading, involving two coil positions at 45° and 75° to the vertical, has been described in the literature[8], and a design based on this was constructed. As shown in Fig. 14, two coils per slot were used and interconnected in the manner shown to avoid unequal currents flowing in the two coils. This gave two sets of coils windings, each consisting of 50 turns and connected in series, so that a peak current of 16A was required from V_2. A sweep time equal to the total exposure period was used, giving three sweep ranges of 5 to 10, and 50μsec respectively.

The method described above suffers from the disadvantage that the optical image must be offset on one side of the photo-cathode. In addition, due to the limitation on coil inductance mentioned above, deflexion sensitivity is necessarily poor. The incorporation of deflexion plates within the convertor tube, as described in a recent paper on the design of an experimental tube[9], would allow a far more efficient deflexion system to be arranged and enable normal oscillographic techniques to be used.

Delay Unit

A Cawkell delay unit, type PE 31. V.A.D., modified to give continuously adjustable delay between 1μsec and 1msec, was incorporated in the equipment. The circuit consists of a Sanatron sawtooth generator, supplying delayed pulses to a phase-splitting output stage. Positive delayed pulses are taken from this unit and used to trigger the gating unit concerned.

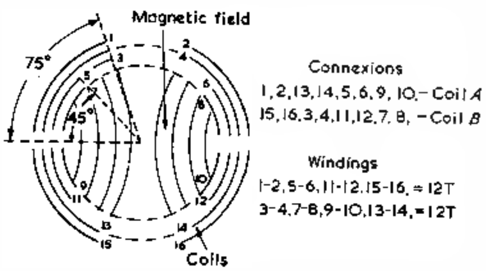


Fig. 14. Coil connexions

Power Supplies

The power supply requirements were fairly conventional. A voltage-doubling e.h.t. circuit was used to provide an accelerating potential, variable up to 20kV. The power input was Variac-controlled and delay circuits were fitted to ensure that the hydrogen thyratrons and pulse modulator valves reached their operating temperature before the e.h.t. was applied. A second power unit was necessary to provide some bias and h.t. supplies not available from the main low-voltage power unit. This latter is a Cawkell P.E. 31 stabilized h.t. unit, and supplies +300V at 350mA, stabilized to 0·1 per cent, together with two separate −150V stabilized bias supplies.

Fairly elaborate precautions were found necessary to ensure that externally produced mains-borne signals were not able to reach the equipment and prematurely trigger the gating circuits. To prevent this a series of filters, (Belling and Lee types 543/15 and L 641/2), were fitted to the main input terminals. These give a 50dB attenuation for incoming signals having frequencies lying between 150kc/s and 250Mc/s.

Performance and Application of the Convertor

With the tubes available for this equipment it was found that when using shutter pulses of 1μsec or less, obtained from the thyratron modulators, the large capacitive currents drawn through the electrodes caused sparking and the resulting cold emission limited the tubes to operating voltages less than 7kV. This figure could be increased to 14kV when 10kΩ resistors were included in both leads close to the tube. The deterioration of the gating pulse rise-times due to the added resistance limited this technique to the 1μsec exposure time.

Under these conditions the photograph of Fig. 15 was taken. This represents a 15kA argon discharge produced by magnetic induction within the 30cm bore of a race-tracked torus of 1 metre major diameter. The length of the race-tracked section was 50cm and the discharges were viewed through a mosaic of holes drilled through the aluminium walls of the section. A mirror allowed orthogonal viewing and the two views of the discharge were recorded simultaneously. The discharge was not stabilized by the application of an external magnetic field, and 'kink' instabilities[3] were clearly seen. The wavelengths of the instabilities are easily measured and these may be related to the eddy-current stabilization of the discharge. Further a measure of the wavelength observed allows an estimate to be made of the instantaneous inductance of the discharge as a whole, so that the conductivity and electron temperature of the discharge may be estimated. Similar photographs have been taken with hydrogen and deuterium discharges in the presence of a stabilizing magnetic field.

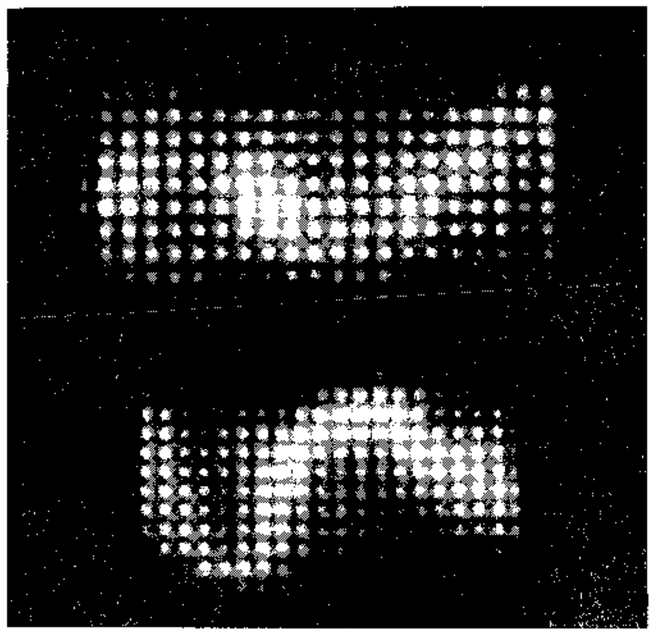


Fig. 15. Discharge in argon viewed through a calibrating mosaic

Acknowledgments

The authors wish to thank Dr. T. E. Allibone, Director of the Research Laboratory, Associated Electrical Industries, for permission to publish this article, and to the staff of the Thermonuclear Section of this Laboratory for their helpful advice and assistance with the experimental work.

Thanks are also due to Dr. P. Schlagen and members of Mullard Research Laboratories for the design of the image convertor tube.

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Wide Range Tuning Cavities for Plug-in Reflex Klystrons

By R. Mather* and J. Sharpe*, B.Sc., A.M.I.E.E.

This article describes a range of external cavities, having a tuning range greater than 30 per cent in the frequency band from 5kMc/s to 12kMc/s, for use with plug-in X-band klystrons. The cavities are of modest dimensions and appear to have advantages over types previously described. Test results are given and applications are suggested.

(Voir page 440 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 445)

OVER the past decade, plug-in reflex klystrons for generating power of 10 to 200mW over the frequency range from 1 000 to 12 000Mc/s have become available in two or three forms[1]. A particularly useful valve is the type R5222 (CV2346) which was designed for a centre frequency around 9 000Mc/s. Various types of cavity have been described for use with this valve[1], having tuning ranges up to 15 per cent which may, however, be of rather large dimensions for the lower frequencies at which the valve may be used (that is, down to 5 000Mc/s). It is the purpose of this article to describe cavities of quite modest dimensions but having a tuning range of some 35 per cent which appear to have advantages over types already described. These cavities employ modified forms of the R5222, types R9525, R9561 and R9562. The basic valve has been described in a previous article in this journal[2].

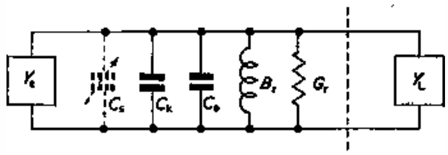

Fig. 1. Simplified equivalent circuit of reflex klystron

Basic Considerations

The reflex klystron may be regarded for the purposes of this article as the resonant system of Fig. 1, where Y_e is the electronic admittance of the beam and Y_L the output or line admittance. The resonant cavity itself is represented by the lumped values B_r, G_r and C, C being shown in two components C_o the inherent capacitance of the cavity and C_k the additional capacitance introduced by the klystron and due to the proximity of the resonator grids. It will be evident that the whole system may be tuned by varying any one of a number of these parameters and in practice several methods are used.

Firstly, the beam admittance Y_e may be varied by altering the reflector potential. This is known as electronic tuning and although convenient in that no mechanical adjustment is involved the frequency range attainable is small, normally of the order 15 to 50Mc/s.

Secondly, the admittance presented to the system by the line, Y_L may be controlled by means of a variable reactance in the output section. This is commonly referred to as 'frequency pulling' and in one form or another is capable of giving a tuning range between 3 and 10 per cent.

The third general method is by mechanical alteration of the resonant cavity, affecting the values C, B_r and G_r. Designs of varying efficacy have involved the introduction of materials of high permittivity into the cavity: the adjustment of the grid gap, thus controlling C_k: the insertion of capacitive or inductive screws into the cavity, affecting C_o or B_r: and the introduction of movable tuning plungers etc., altering the whole geometry of the cavity[1].

The method described below falls into this third category and takes the form of the introduction of a capacitive electrode (represented by C_s in Fig. 1) into the resonator, and represents an extension of an arrangement used in a

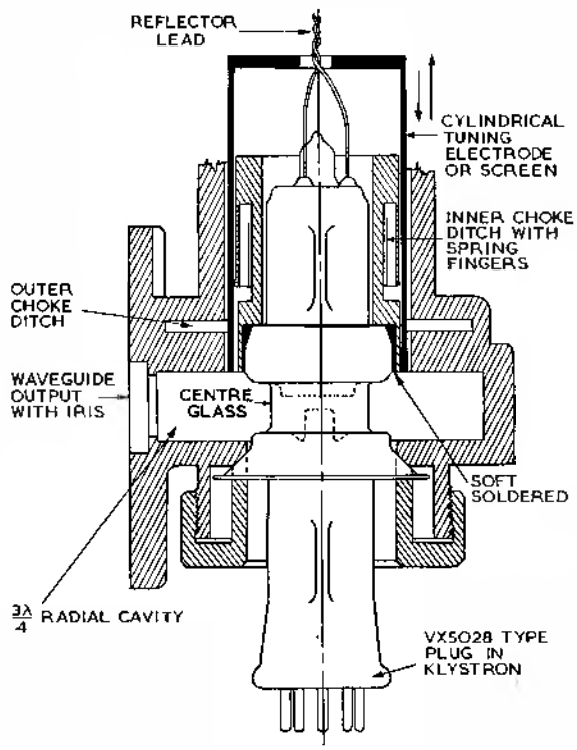

Fig. 2. General arrangement of tuning mechanism

war-time reflex klystron[3], type CV87. As shown in Fig. 2, it consists of the axial insertion of a hollow open-ended cylindrical conducting electrode into a $3\lambda/4$ cavity with waveguide output. This tuning electrode is concentric with the $3\lambda/4$ cavity and also with the tube itself. The resultant increase in capacitance as the tuning electrode is inserted reduces the resonant frequency of the whole system, and proper choice of dimensions enables a 35 per cent tuning range to be achieved. The diagram shows the modified R5222 in position in the cavity, the position of the grid gap within the centre glass being dotted.

Criteria for Essential Dimensions

The diameter of the $3\lambda/4$ cavity is dictated by the upper frequency limit required, as shown in Fig. 3 which gives the relationship between internal cavity diameter and resonant wavelength. The 'ideal' curve refers to a hollow cylinder whose diameter is $3\lambda/2$, while the empirical curve

* *E.M.I. Electronics Ltd.*

relates to a similar cavity with the reflex klystron in position, the longer wavelength at a given diameter being due to the additional capacitance of the grid gap. The optimum diameter of the cylindrical tuning electrode or screen has been found experimentally to be that diameter which, when regarded as a $\lambda/4$ radial line cavity, gives a frequency of oscillation corresponding to the lower frequency limit of the tuneable system. Fig. 3 also shows the relationship between resonant wavelength and internal diameter of a $\lambda/4$ cavity, again showing the effect of the valve capacitance.

Thus, if the required tuning range is 5 000 to 8 000Mc/s Fig. 3 shows the $3\lambda/4$ cavity diameter to be 4·96cm, while the optimum internal diameter of the screen is 2·2cm. In the case of the R5222, the smallest practicable screen diameter is about 1·8cm, owing to the limiting diameter of the centre glass section.

From this description, the conception may be formed of a gradual transition from a certain frequency in the $3\lambda/4$ cavity mode to a lower limiting frequency in the $\lambda/4$ cavity mode as the screen insertion varies from zero to maximum.

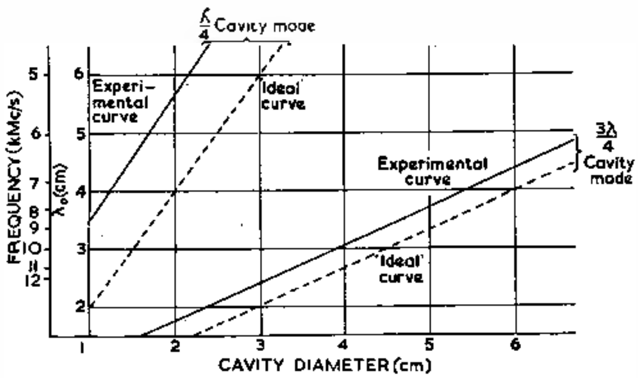

Fig. 3. Resonant wavelength versus cavity diameter

Regarded simply as an additional capacitance which is introduced into an existing resonant system, the screen should have a diameter such that the conducting walls coincide with maximum electric field, at which point its capacitance will be most effective. Again, in view of the break necessary in the wall of the cavity, this diameter should, in order to keep losses to a minimum, be such that the break occurs at or near a voltage maximum. Owing, however, to the considerable variation in field distribution over the tuning range, this voltage maximum is very ill-defined and, in fact, the resultant optimum screen diameters derived from the two concepts are approximately equal.

As shown in Fig. 2 it is found advisable to introduce choke ditches, both at the inner and outer faces of the screen, the latter being of the non-contacting type. The inner contact is, however, important, owing to the necessity of holding the second or reflector grid at resonator potential. A number of spring fingers make contact at a distance of $\lambda_{av}/4$ from the cavity wall and a similar distance from the choke short-circuit, where λ_{av} is the wavelength at mean frequency. The fingers are cut by slotting a beryllium copper ring prior to hardening, and the contact assembly is soft-soldered to the valve copper, cut down where necessary to the appropriate diameter from a standard R5222. Fig. 4 illustrates details of this contact assembly.

Some Typical Results

Cavities covering three frequency bands have so far been developed, the bands being:

Type 25181	5·4 to 8·2kMc/s
Type 25157	7·0 to 10·3kMc/s
Type 25182	8·2 to 11·7kMc/s

As is to be expected, power output is in general greatest at the design frequency of the R5222, in the neighbour-

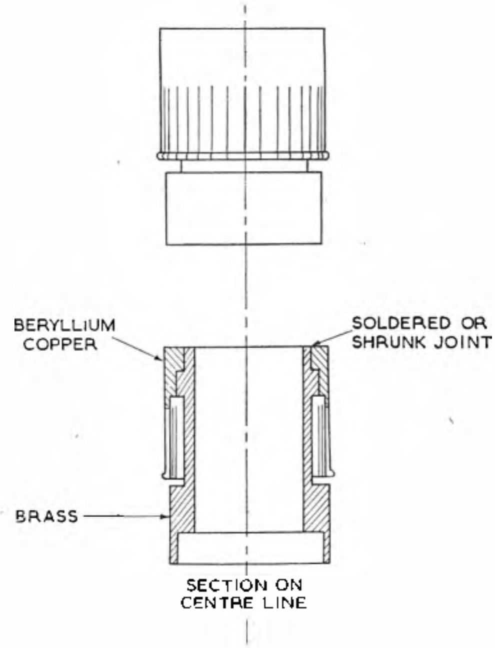

Fig. 4. Choke contact assembly

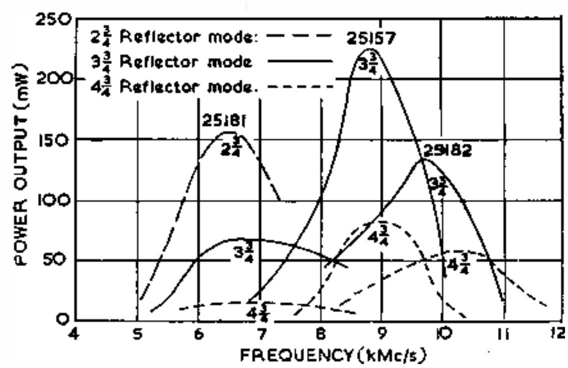

Fig. 5. Power versus frequency—typical curves

hood of 9·0 to 9·5kMc/s and falls off, in a given reflector mode, when well away from this band. Fig. 5 shows typical power frequency characteristics of the three cavity types, and it will be observed that, although for full exploitation of the frequency range a change of mode may be necessary, an extremely useful range is available in a single mode. This fact is of importance when mechanical reflector potential tracking is necessary as, in a given mode, the change required in reflector voltage is approximately proportional to change in frequency. The choice of mode may further be limited by the reflector potential available from the power supply, the range employed in the present case being -50V to -450V with respect to cathode.

As indicated in Fig. 2 the r.f. power is coupled into the load via an iris of such dimensions as to provide good power output at the centre of the frequency range covered by the cavity when a matched load is presented to the valve. An adjustable matching screw is inserted across the iris, firstly, in order to give the coupling as wide a working band as possible and secondly, as a fine load adjustment to give optimum output where a restricted frequency band is in use. Improvement in the output may also be obtained by re-setting this screw when changing the reflector mode. The curves of Fig. 5 were taken with this screw adjusted to give maximum frequency coverage in each case.

The electronic tuning range, specified as the frequency change over a particular reflector mode, between half power points is increased in this type of cavity as the tuning electrode is inserted, and typical values are shown in Table 1 for the more useful modes.

TABLE I
Electronic Tuning

Type	25181		25157		25182	
Mode	$2\frac{3}{4}$	$3\frac{3}{4}$	$3\frac{3}{4}$	$4\frac{3}{4}$	$3\frac{3}{4}$	$4\frac{3}{4}$
Higher Frequency (Mc/s)	—	12	12	23	12	24
Centre Frequency (Mc/s)	10	15	15	30	14	28
Lower Frequency (Mc/s)	12	18	19	35	18	35

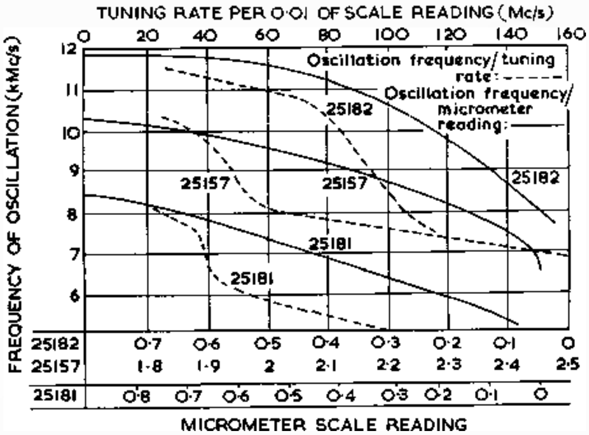

Fig. 6. Frequency of oscillation/Micrometer scale reading tuning rate—typical curves

With regard to the mechanical tuning sensitivity, it is evident that, although this may be controlled to some extent by the nature of the adjusting device employed, the change in frequency for a given increment of insertion of the screen must increase as the frequency approaches its lower limit, owing to the decrease in spacing between the screen and the floor of the cavity. Fig. 6 illustrates the mechanical tuning characteristics, and variation of tuning sensitivity over the tuning range, and it will be seen that, as would be expected, greater re-setting accuracy is attained at the high frequency end of the band. For the curves of Fig. 6 the cavity type 25157 was fitted with a cam movement designed to hold the rate of tuning constant. In the case of the other two types, scale reading may be regarded as being proportional to screen penetration.

Temperature compensation may be readily achieved over a considerable band by suitably designing the mechanical tuning arrangement and may further be controlled by the choice of material from which the screen is turned. If a relatively poor conductor, such as mild steel, is used, copper plating is found necessary in order to keep losses to a low level.

Possible cavity mode interference has been the subject of close scrutiny, since it was suspected that the electrode arrangement would at certain settings tend to promote unwanted modes of oscillation. It has been found, however, that within the working range of the particular cavity, no mode interference is apparent with resonator potentials of between 300V and 350V, using valves of normal production tolerances.

Conclusions

Although the foregoing notes are based on data drawn exclusively from experiments on the R5222, the general

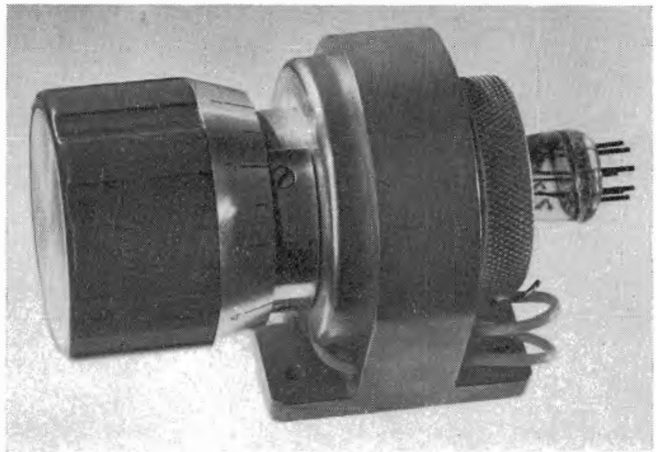

Fig. 7. Complete oscillator type 2518

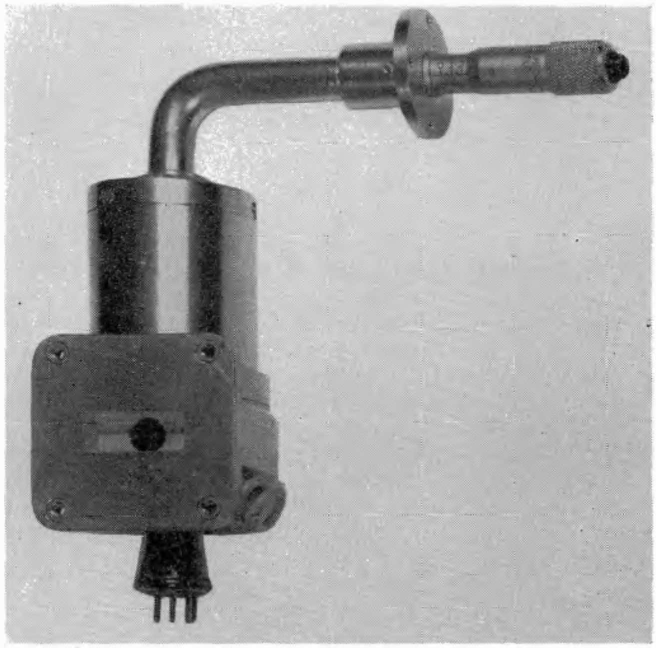

Fig. 8. Type 25182 with remote tuner

method would appear to be equally applicable to any plug-in type of klystron where the resonant cavity is not an integral part of the valve construction. A prototype cavity designed on the above considerations has been produced operating in the S-band (centred on 4 000Mc/s) and initial results, while as yet incomplete, tend to confirm this particular conclusion. The valve used in this latter case is the VX5048, which is still under development.

The concentric arrangement of the tuning electrode lends itself to a variety of tuning mechanisms. Among those which have been developed are a cam drive covering the whole band in half a turn which may be driven continuously by a motor; the micrometer adjustment illustrated in Fig. 7; and the remote micrometer movement of Fig. 8 which was designed for use where the klystron is to be mounted behind a panel.

Applications would appear to be numerous; the wide frequency range and ease and accuracy of re-setting makes these klystrons, in the form illustrated in Fig. 7, particularly suited to general laboratory use and microwave testing. The high-re-setting accuracy (approximately ± 2Mc/s over most of the range) makes the klystron, once it has been calibrated, independent of a wavemeter for many practical purposes. The remote micrometer control of Fig. 8 would be suitable for use in instruments such as waveform analysers or signal generators where the frequency of the r.f. source or local oscillator is set manually. The angle of the driving micrometer related to the plane of the output flange is infinitely variable and the re-setting accuracy again is high. The motor driven cam tuner suggests possible applications in reflectometers, panoramic receivers and automatic microwave test gear. At the present stage of development frictional wear on the spring fingers prohibits continuous running at a high sweeping frequency, but it is hoped to achieve a sufficiently high rate for convenient visual presentation by using a non-contacting choke assembly.

Acknowledgments

The authors are indebted to E.M.I. Electronics Ltd for permission to publish this article.

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THE TOTAL EXCURSION RESISTOR

A New Concept

By R. H. W. Burkett*, B.Sc., A.M.I.E.E.

The increasing demand for electronic equipment to give reliable operation over long periods has made it necessary to examine the behaviour of resistors when loaded for several years. Certain types of film resistor have controlled and predictable characteristics so that it is possible to guarantee that such resistors will hold their tolerance over several years of continuous use. Thus a resistor may be supplied to be within a guaranteed 'Total Excursion' which will include selection tolerance and all changes tending to cause the resistor to vary from its nominal resistance. Maximum excursions of 5, 7 and 10 per cent are shown to be a practical proposition, with the principle extended to 1, 2 and 3 per cent Total Excursions, for special purposes. Data is given showing that it is possible to design resistors to meet such a specification. A proposal is made as to the method of colour coding Total Excursion Resistors.

(Voir page 441 pour le résumé en français; Zusammenfassung in deutscher Sprache Seite 446)

When an equipment contains only a few hundred components and is used irregularly the probability of a component failure is quite small. Thus, such instruments are seldom out of commission and it is possible for the user to be philosophical about an operational failure. With military equipment a much higher degree of reliability is required, though this until recently has generally been associated with a relatively short operational life. The advent of large equipments using large numbers of components and being required for continuous use, equipments which are inaccessible for maintenance, and equipments where operational failure is prohibitively costly or dangerous, has created a demand for components designed for rather different operational needs than heretofore. Examples are electronic computers and electronic telephone equipment, equipment for use in or around nuclear reactors, and submerged repeaters or navigational instruments. The specifications for these equipments must require reliable and predictable operation for many thousands of hours. It therefore follows that components, not only individually but collectively, are required to have a highly predictable performance.

There is some doubt as to the suitability of all components for such purposes and it is not surprising that a great deal of work has been done on establishing the basic reliability of components. As a result of a long programme of study on resistors of various types, a new approach—that of 'Total Excursion'—should prove attractive to circuit engineers. The basic concept would, of course, be equally applicable to other types of component whose characteristics vary with time and operating conditions.

The Problem

In any circuit the nominal resistance value of the resistors will be the optimum value, but designers recognize that a certain variation of resistance must be tolerated if reasonably priced components are to be used. Accordingly resistors at every point are allotted a tolerance, the tolerance being the

* *Technical Director, Welwyn Electrical Laboratories Ltd.*

variation of resistance value which has been found to be acceptable without the apparatus giving a performance below specification limits.

Different selection tolerances are laid down for various types of resistors and, in general, the established tolerances are 1, 2 and 5 per cent for film resistors and 5, 10 and 20 per cent for carbon composition. Of course precision wire wound resistors are customarily offered at much closer tolerances.

Designers have long recognized that they must have something in reserve when specifying selection tolerances. This reserve is necessary in order to provide for any change in resistance value in use, either due to permanent or secular changes, or to cyclic changes such as arise from temperature changes. Accordingly engineers have to specify the type of component to be used so that the amount of deviation from the original value is under some sort of control. Thus a 20 per cent resistance requirement in a non-critical position could be met by the use of a 5 per cent selection tolerance carbon composition resistor which could have an additional instability of between 5 and 15 per cent. On the other hand, where a resistor has to be held very closely to the optimum value one of the more stable types such as the pyrolytic carbon, metal film, or wire-wound precision resistor would be specified, probably at ± 1 per cent selection tolerance.

With the advent of more complicated equipment however, it becomes much more difficult to analyse each component in terms of initial tolerance and stability. A major difficulty is that the designer cannot be familiar with the detailed performance of all resistors to the extent that he could specify those most suitable for his purpose taking into account performance and cost.

The 'Total Excursion' Resistor

The proper approach is that all electronic components should be supplied to meet not only the initial requirements of the designer, but the ultimate requirements under all conditions.

It follows therefore that the resistor manufacturer cannot supply a resistor to an initial selection tolerance, together with data concerning the behaviour of the resistor under various hypothetical conditions from which the user must make deductions concerning its ultimate resistance. The manufacturer should, in fact, collaborate with the user to establish the operating conditions and the expected working life of the equipment, and supply a resistor which will be within the tolerable resistance range for its expected working life. After considerable research in this direction there is now available a 'Total Excursion Resistor'; a component which is manufactured and selected in such a way that throughout its life it will always be within the range of resistance which is guaranteed by the supplier. As is now the case for selection tolerances, a range of 'Total Excursions' should be standardized, and 5, 7, 10 and 20 per cent are proposed as a preliminary range of Total Excursion Resistors for general purpose use, with the expectation that it will be necessary to extend the range to 1, 2 and 3 per cent Total Excursions for special applications.

Using components having a guaranteed Total Excursion, the designer of equipment leaves the problem of behaviour in the hands of the component manufacturer. During the development of his equipment he finds the optimum value of resistance which he needs, ensures that it will be adequately efficient at the maximum Total Excursion and, having established that this is the case, installs the requisite component without doubts as to its ability to meet his requirements.

With a knowledge of the various Total Excursions available, the designer needs only to decide the correct one for his apparatus, rather than attempt to balance tolerance and problematic stability on the one hand against target price for the equipment on the other. If investigation shows that one Total Excursion, at a certain price level, will not quite meet his technical need he knows immediately the cost of going to the next grade up.

While there is no doubt that the availability of such components will relieve the electronic designer of doubts as to the long-term behaviour of his equipment, it places an equivalent responsibility on the component manufacturer.

Design of the Resistor

Although manufacturers may not be willing to admit all the shortcomings of their products, it is necessary to avoid any tendency to self-deception about the characteristics of the unit when they are required to guarantee their behaviour under all possible conditions. Therefore, the preliminary to offering such a product must be a rigorous programme of testing to establish its properties under various operating conditions.

The first thing to decide is over what period the guaranteed Total Excursion must be operative, and it is proposed that the minimum should be 10 000 hours operation for most equipments with 25 000 hours for certain types. This latter period corresponds to three years continuous operation. Accordingly, numbers of resistors have been placed on continuous load and their behaviour has been studied for such periods. It has become apparent that there are a number of factors which must be taken into account to explain the ultimate resistance value of the resistor. It has been found that these factors conform to patterns which are typical (and therefore recognizable) in at least two types of film resistor, namely pyrolytic carbon and tin oxide films.

The factors which have to be taken into account are as follows:

(a) Temperature change
(b) Initial stability
(c) Long-term stability
(d) Voltage coefficient

Voltage coefficient is important only in certain composition resistors of high ohmic value.

Except for military applications the effect of moisture is not an important matter since it is now recognized that electronic instruments are both more efficient and reliable, as well as cheaper, if they are used in conditions which have no extremes of temperature or high humidity.

The results which emerge from the investigation are (a) that the initial load stability can be up to 5 per cent for carbon composition resistors, up to 1 per cent for pyrolytic carbon or tin oxide film resistors and up to 0·25 per cent for metal films; (b) that the long term drift in terms of per cent per thousand hours varies very much with resistance value and between ranges, but with typical values as follows:

(1) Composition resistors 1 per cent per thousand hours.

(2) Pyrolytic carbon or tin oxide film resistors 0·1 per cent or less per thousand hours.

(3) Metal film resistors 0·05 per cent per thousand hours.

It would be interesting to study with complete thoroughness all types of resistor, but as the author has been very much concerned with finding one solution to a problem rather than amassing information for its own sake, it was deduced from the data available that the tightest Total Excursions having most appeal to equipment users could best, and probably only, be met by the use of tin oxide or pyrolytic carbon resistors. On the one hand carbon composition resistors, because of their relatively high long-term drift could be suitable only for a total excursion of 20 per cent or greater, and to meet a figure of 15 per cent, or even 10 per cent, would involve an uneconomically restricted initial selection tolerance. On the other hand, metal film resistors would meet the requirements for the closer total excursions but, at the present time, they would be very expensive, and there is a certain degree of doubt as to their reliability. This subject of reliability will, however, be dealt with later.

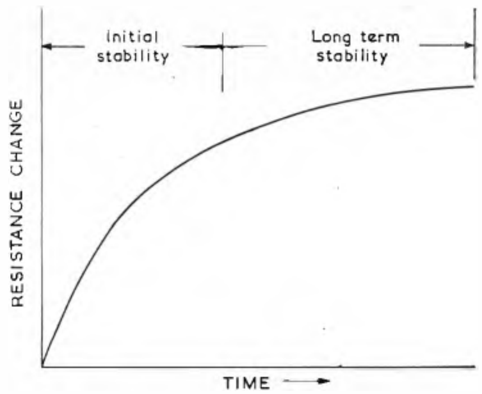


Fig. 1. Typical stability curve for a carbon or tin oxide film resistor

Essential Characteristics of Film Resistors for this Application

Two important features emerge from the study of the behaviour of film resistors under long-term load. The first feature is that there is initially some incidence of resistors giving rise to anomalous changes of resistance or even becoming open-circuit. This soon falls to the point where it is negligible. This aspect has been pursued further and it is found that a pre-conditioning or stabilization process can be devised which can ensure that the incidence of such defective resistors is negligible, even initially. In passing it is necessary to point out that the occurrence of defectives in tin oxide resistors is very much lower than with pyrolytic carbon resistors. The second feature is that film resistors have a characteristic behaviour pattern. It is found that with both pyrolytic carbon and tin oxide resistors there is a settling down period during which the rate of change of resistance is fairly high and this is the 'initial stability' of the component. This period may be as short as 50 hours for low resistance values and up to 500 hours for higher values. After this the changes are very much smaller and they appear to be a linear function of time. Fig. 1 shows the characteristic pattern of stability and it will be clear that there are two parts to the curve; substantial initial change and, after that, a very much slower drift. In order to predict the resistance of a resistor these two effects must be separated. Thus, the initial stability, once it has been achieved, is no longer a factor in the behaviour of the component and thereafter one is essentially concerned with the long-term stability. If the magnitude of this stability is established in terms of per cent per thousand hours it follows that for given operating conditions the total magnitude of this change can be computed.

A further factor which must be taken into account is the load characteristic, i.e. the change of resistance which takes place because of the combined effect of temperature rise and temperature coefficient. As a resistor subject to load will reach its steady state temperature in a matter of a few minutes it is convenient, from the user's point of view, to regard this as an instantaneous change.

Thus, in attempting to predict the behaviour of the component one has to take into account:

(a) The instantaneous change due to load application.

(b) The initial stability which is completed in 50 to 500 hours.

(c) The long-term stability.

Table 1 gives observed values for these parameters on current pyrolytic carbon film and tin oxide resistors.

It can be seen that with a long-term stability of better than 0·1 per cent/1 000 hours one is in a position to predict the effect of 10 000 or 25 000 hours use.

Unfortunately, in any class of resistor the characteristics of the resistor will change with resistance value as well as with resistor dimensions, so that a different figure for these various factors must be adopted for each resistance value. There is, of course, no point of discontinuity in the relationship except that which arises at the critical resistance value, that is the resistance value above which the power dissipation is limited by the voltage limitation, so that it is not necessary to test every resistor in the spectrum. It is sufficient to test only representative values and the curve may be drawn by interpolation.

In order to illustrate the sort of calculation which the resistor engineer needs to perform, two examples are given:

TABLE 1.

	PYROLYTIC CARBON FILM	TIN OXIDE FILM
Temperature coefficient (parts per million)	-180 to -250 for low values -500 to -700 for high values	0 to $+500$ for low values 0 to -500 for high values
Initial load stability (per cent)	0 to $\pm 0·2$ for low values $+0·25$ to $+0·75$ for high values	$\pm$ 0. 5 to $\pm$ 1 for low values ± 1 to ± 2 for high values
Long-term load stability (per cent/1000 hours)	0 to $+0·02$ for low values 0 to $+0·04$ for high values	Less than 0·1 for all values

It should be noted that the figures for Tin Oxide Films relate to bulk production components with no stabilization. Much better performances can be obtained when required (i.e., RCS/114 standards).

EXAMPLE 1

A tin oxide resistor has an average film temperature rise of 50°C, a temperature coefficient between zero and $300/10^6$ positive, a maximum initial stability of ± 1 per cent and a maximum long-term stability of up to $-0{\cdot}06$ per cent/1 000 hours. Fig. 2 indicates the expected relationship between resistance and time. It can be seen that the maximum range of change of resistance is from 2·5 per cent positive to 2·5 per cent negative. Accordingly, it is possible to construct Fig. 3 which shows the changes of Fig. 2 subtracted from a proposed total excursion of ± 7 per cent so as to ensure the resistors remain within the ± 7 per cent of their nominal value at the end of 25 000 hours. This shows that the selection tolerance of the resistor must be between $+4{\cdot}5$ per cent and $-4{\cdot}5$ per cent.

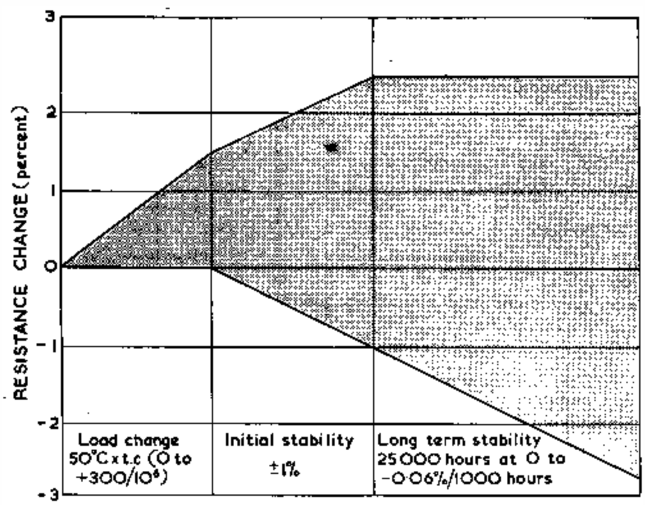


Fig. 2. Maximum range of variation of a typical tin oxide film resistor

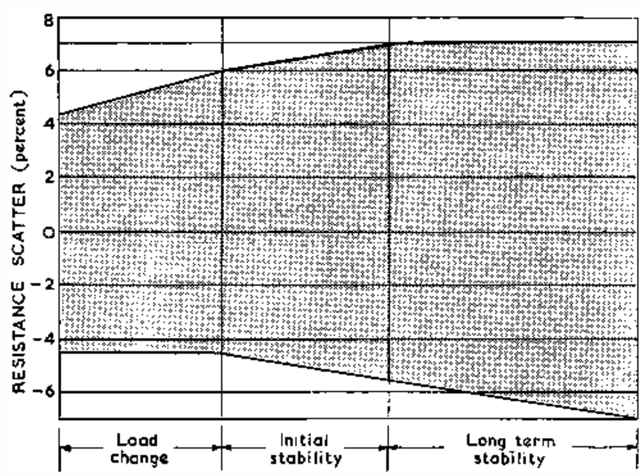


Fig. 3. Range of variation of a 7% total excursion tin oxide resistor

EXAMPLE 2

A pyrolytic carbon film resistor has an average film temperature rise of 50°C, a temperature coefficient of between 250 and $350/10^6$ negative, an initial stability of $\pm 0{\cdot}3$ per cent and a long-term stability of up to $+0{\cdot}04$ per cent/1 000 hours. The calculations are similar and from Fig. 4 it will be seen that the initial selection tolerances must be $+4{\cdot}9$ per cent to $-2{\cdot}9$ per cent if the resistor is to remain within ± 5 per cent for the duration of a 25 000 hour life.

It would appear that if the manufacturer of the resistor can hold these initial selection tolerances, he can then give the total excursion guarantee that the customer requires. This may be simple to the user, but there are many problems for the manufacturer. He must understand his process to the point where he can guarantee that the drift rates are held to the limits which he is specifying, and this guarantee must operate on his bulk production, that is for every resistor. The bulk production must be sufficiently controlled to enable him to hold selection tolerances necessary to produce an economic yield at the guaranteed total excursion. If these resistors are marketed only in the preferred resistance values and the initial selection tolerance is closer than ± 5 per cent, then any resistors outside his initial selection

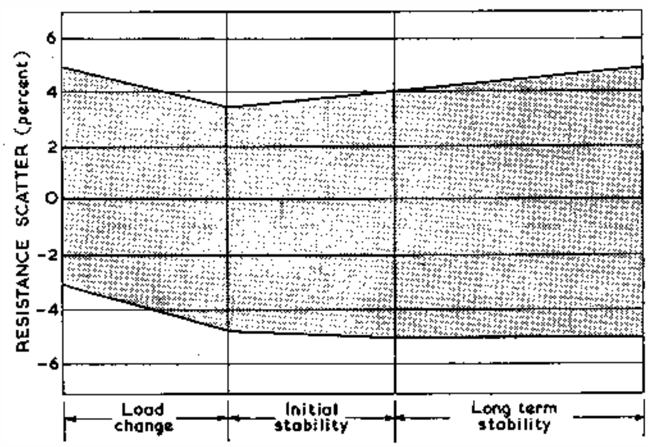


Fig. 4. Range of variation of a typical 5% total excursion carbon film resistor

tolerances are not likely to be sold. This must be remedied by ensuring the accuracy of the control of production or by encouraging the use of a wider total excursion by a suitable price advantage.

Finally it is necessary for the manufacturer to market such resistors at a price that will encourage their use in the appropriate equipment.

Clearly, the desirable features of a total excursion resistor cannot be achieved unless the manufacturer is in bulk production, and with the necessary organization to control the processes and the quality of the product to the necessary degree

Reliability

As mentioned earlier, reliability is not less important than stability. While a resistor must be regarded as a failure if it has drifted out of accuracy it is also useless if it is open-circuit or has changed in resistance value in a substantial or dramatic fashion at any part of its life. It is therefore necessary to give thought to the material which should be used for total excursion resistors.

With proper stabilization the reliability of pyrolytic carbon film resistors can be rendered better than 1 in 10 000 individual resistors or much better than 1 in 10 000 000 resistor-hours. This performance is adequate for even the most complicated equipments since it must be remembered that defective components will occur in the settling down period of the apparatus possibly even before installation. Thus, one can expect in any apparatus employing 10 000 resistors the chance of one resistor failure in the settling-in period, and thereafter the failure rate due to defective resistors to be much better than once in a thousand hours of continuous use, i.e., very much less frequently than once every six weeks. For part-time use that figure is very much improved. One must, however, give consideration to the acceptability of the product. Because of lack of understanding by users, the pyrolytic carbon film resistor has been given a reputation for unreliability and there are certain engineers who cannot be persuaded that the reputation is undeserved.

The tin oxide film is an excellent basis for the total excursion resistor. With this material the defective rates are very much lower than with carbon films, possibly because of the method of manufacture and possibly because of the affinity that the film has for the substrate. The tin oxide film is much less affected by atmospheric moisture and, what is more important still, should a hot spot develop for any reason there is no fear that it would burn out. In fact, as such films when subject to very high temperatures tend to produce a reduction in resistance value, the formation of a hot spot brings into action certain compensatory changes. It is for these reasons that the tin oxide film is considered to be the best basis for the first total excursion resistors. Field experience indicates that film resistors should have a substantial mechanical protection, which means that they must be moulded with a relatively thick coat of resin. As tin oxide resistors generally run at fairly high temperatures such resins must necessarily be high temperature resins so that the choice is somewhat limited. Nevertheless it is possible to fabricate a resistor of this type with satisfactory insulation.

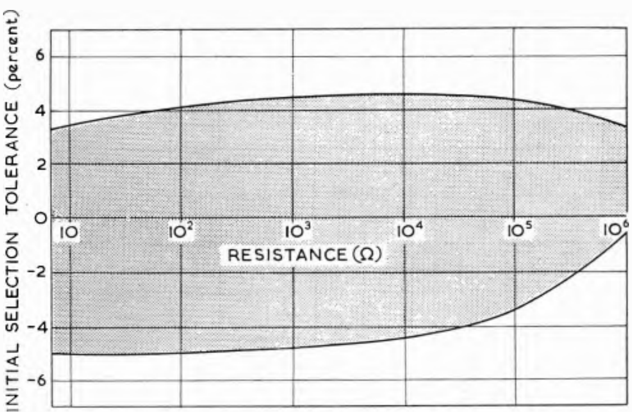


Fig. 5. Range of initial selection tolerance for ±7% total excursion

Size is a matter of major importance and it requires very special processing methods to fabricate resistors small enough to meet current demands.

With a total excursion resistor the initial selection tolerance is bound to depend upon the resistance value since stability and temperature coefficient will vary. Fig. 5 gives an artificial example of the way selection tolerance might vary with resistance on a typical tin oxide film resistor. The band of initial selection tolerance becomes narrower with increasing resistance value, because the 25 000 hour stability deteriorates from about ±2·0 per cent for low values to ±3·5 per cent for high values. It is assymmetrical with respect to the axis on account of the variation in temperature coefficient from 0 to $+300/10^6$ for low values to 0 to $-500/10^6$ for high values. In actual practice the picture would be more complicated because the permanent resistance changes tend to be negative rather than positive. In addition the changes due to the effect of temperature coefficient may be reduced at high resistance values because power dissipation would be reduced because of a possible voltage limitation.

Where an apparatus is likely to operate over a wide range of temperatures then it would be necessary to take this into account and the limits of initial selection tolerance suitably adjusted. Fortunately, the adjustment necessary is bound to be less than expected since low temperatures will be compensated for by the temperature rise of the resistor.

Identification of Total Excursion

It is clearly necessary to distinguish a resistor sold as a 'total excursion' component from one which is sold only to a selection tolerance. It is suggested that a total excursion resistor be indicated by using the normal three band colour

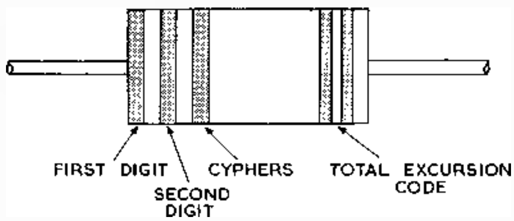


Fig. 6. Colour coding for total excursion resistor

code at one end of the resistor to indicate the nominal resistance value with a double band of suitable colour and situated at the opposite end of the resistor to indicate the total excursion. The double band is considered to convey the conception of a limited range of resistance value and has the advantage that it can be applied by existing methods. The diagram of Fig. 6 indicates the general appearance of a component marked in this way.

The double bands would have colours as follows:

Brown	1 per cent total excursion
Red	2 per cent total excursion
Orange	3 per cent total excursion
Gold	5 per cent total excursion
Violet	7 per cent total excursion
Silver	10 per cent total excursion

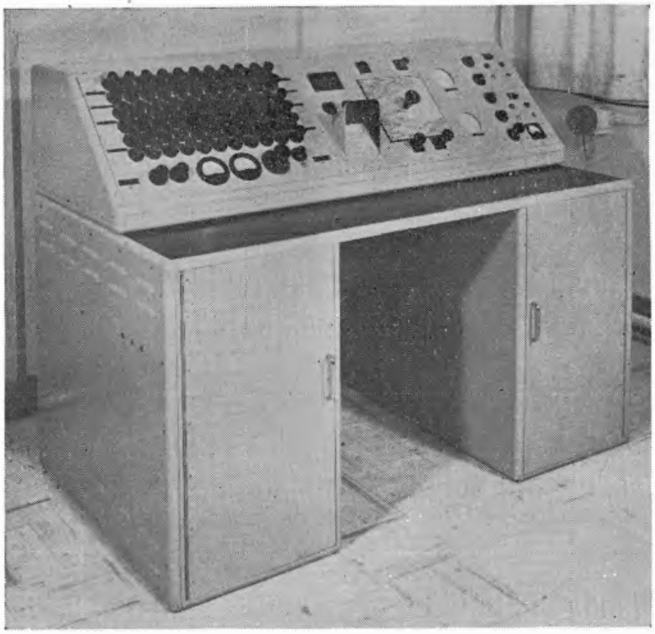
A 20 per cent total excursion might be indicated by a red and black band, though this may prove to be impracticable.

There may be other methods of indicating a total excursion resistor which will find preference to the double band method. There is something to be said for the use of a wavy line or a dotted line as alternative methods. The ultimate decision will necessarily depend upon the ready acceptance by the user and ease of application by the manufacturer.

Acknowledgments

The author wishes to thank his colleagues at Welwyn Electrical Laboratories Limited who have provided the test data upon which this article is based.

An Analogue Computer for the Prediction of Acoustic Propagation in the Atmosphere



By F. A. Key* and W. G. P. Lamb*

An analogue computer designed to enable the prediction of acoustic propagation conditions in the atmosphere is described. Data concerning the variations of air temperature and wind velocity with altitude are fed to the computer and ray patterns representing acoustic propagation in any required direction are presented on the screen of a cathode-ray tube.

(Voir page 441 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 446)

ACOUSTIC propagation over large distances, and consequently the intensity of sound experienced at places remote from the explosion of industrial or experimental charges, is largely determined by the atmospheric conditions prevailing at the time of firing. The main factors affecting this propagation are the variations with altitude of the wind velocity and the air temperature, and if adequate information concerning these parameters is available, ray trajectories and hence sound intensities may be calculated. Since these computations were found to be somewhat lengthy and laborious, the authors decided to construct a repetitive analogue computer which would enable predictions of the variation of sound intensity over the whole of the surrounding area to be made in a matter of a few minutes. The computer was to operate on data supplied by a nearby meteorological station and present the required information in the form of sound ray trajectories displayed on the screen of a cathode-ray tube.

The trajectory of a sound ray travelling in a medium in which the sound speed varies with altitude may be regarded as being determined by the equation for the radius of curvature R of the ray, viz:

$$R = \frac{-c}{dc/dz} \quad \ldots\ldots (1)$$

where c is the sound velocity at height z. In the case of atmospheric propagation the sound velocity is determined partly by the temperature of the air expressed by the well-known relation $c = \sqrt{(\gamma RT)}$, and partly by the wind velocity resolved in the direction of travel.

If attention is confined to ray paths near the horizontal, the effects of temperature and wind variations may be combined to yield an effective sound speed gradient enabling equation (1) to be applied.

From elementary calculus:

$$R = \frac{\{1 + (dz/dr)^2\}^{3/2}}{d^2z/dr^2} \quad \ldots\ldots (2)$$

where r is the range co-ordinate.

For rays close to the horizontal $(dz/dr)^2$ may be neglected so that combining equations (1) and (2):

$$d^2z/dr^2 = (-1/c) \,.\, (dc/dz) \quad \ldots\ldots (3)$$

or since dc/dz is a function of z:

$$d^2z/dr^2 = (-1/c)\, f(z) \quad \ldots\ldots (3a)$$

Integrating twice gives:

$$z = (-1/c) \int \left[\int f(z)\, dr \right] dr + Ar \quad \ldots\ldots (4)$$

where A is a constant determining the initial angle of the ray.

The analogue computer described in this article was designed to trace ray paths by performing this solution for different values of A. A range of 25 miles was covered and data up to an altitude of 6 000ft used in the computations.

Principle of Operation

Fig. 1 shows a block diagram of the method used to solve the differential equation. The height co-ordinate z is represented in the machine by potential (1V equals 1 000ft), and the radial distance r is identified with time (1msec equals 10 000ft). Two integrators arranged in series are used to perform the double integration required to provide z, and this is then displayed against the radial distance r on the screen of a cathode-ray tube.

The above photograph shows the complete computer

* A.W.R.E., U.K. Atomic Energy Authority.

The term dc/dz as a function of height is obtained in the form of a series of discrete steps from meteorological information set up on the so-called 'data handling unit'. This information is tabulated at a number of fixed heights, viz. 100ft, 500ft, 1 000ft, 1 500ft, 2 000ft, 2 500ft, 3 000ft, 3 500ft, 4 000ft, 5 000ft, and 6 000ft and set up as air temperature and components of wind resolved in northerly and easterly directions. These components are obtained with the aid of a built-in resolver on which the wind speed and direction are set by the operator, and the two components read from meters.

The result z, in addition to being displayed, is examined by a 'data switching unit'. This determines in which height interval it lies at any given moment and then obtains from the 'data handling unit' the three appropriate gradients of temperature, wind-north and wind-east. The two wind gradients are fed to a direction control panel, where the required direction may be selected by the use of a control, coupled to two gauged sine-cosine potentiometers, arranged to compute the effective wind gradient using the relationship:

$$dW_\alpha/dz = (dW_N/dz)\cos\alpha + (dW_E/dz)\sin\alpha$$

where the subscripts α, N and E refer to the wind velocity resolved in the direction $\alpha°$ East of North, in a northerly and an easterly direction respectively. The wind gradient resolved in the required direction is then added to the sound speed gradient due to temperature, according to the relation:

$$dc/dz = (dW_\alpha/dz) + 1{\cdot}11\,(dT/dz)$$

where dT/dz is the temperature gradient in degrees Fahrenheit per foot. This summation is multiplied by the factor $1/c$, (taken as a constant since fractional changes in c are small) and fed to the first integrator.

The operation of the computer is synchronized by a master oscillator. This is used to generate the time-base for driving the X plates of the main display tube and also controls two integrator clamps which discharge the integrator capacitors at the end of each computing cycle.

Ten solutions representing the trajectories of rays starting at 1° intervals between 0 to 9° are performed and displayed successively, taking approximately 150msec to build up the whole picture. The initial angle of the rays is changed between each computing cycle by a series of electronic switches, actuated by a Dekatron selector tube triggered from the master oscillator.

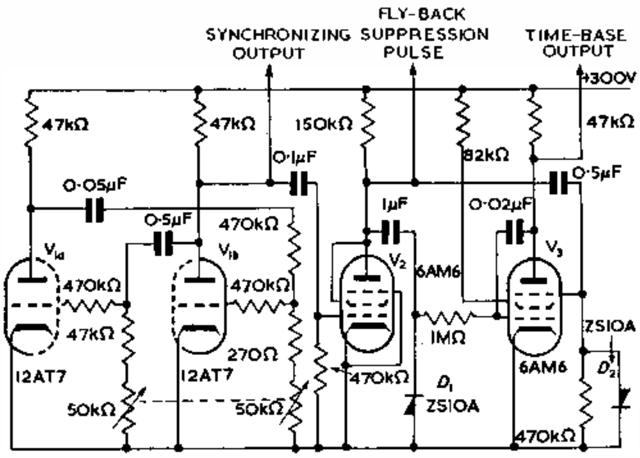


Fig. 2. Master oscillator and time-base circuit

The Master Oscillator and Time-Base

The circuit diagram of this unit is shown in Fig. 2. A double triode, V_1, is connected as a free running asymmetrical multivibrator giving a positive pulse considerably shorter than the negative. The repetition frequency is adjustable by means of two ganged 50kΩ variable resistors, connected in series with the grid leaks.

The output from the anode of the second half of the double triode is fed to a valve V_2 which acts as a limiter and phase invertor. The shorter negative pulse appearing at its anode, is used to initiate the reset process throughout the computer and is also fed to the grid of the display tube to suppress the trace during this time. A Miller integrator V_3 is also fed from the anode of V_2 by the 1μF coupling capacitor, the positive d.c. component being restored by the silicon junction diode D_1.

The values of the series resistor and feedback capacitor of the Miller integrator were chosen so that its anode would fall to a potential of about 100V, at the end of the computing period. The suppressor grid of V_3 is also coupled by a capacitor to the anode of V_2, but during the computing time is held at earth potential by the diode D_2. The negative step which occurs at the anode of V_2 during the reset period, therefore cuts off V_3 by putting a large negative bias on the suppressor grid, thus recharging the feedback capacitor for the next scan. The computing time is about 13msec and the reset time approximately 2msec.

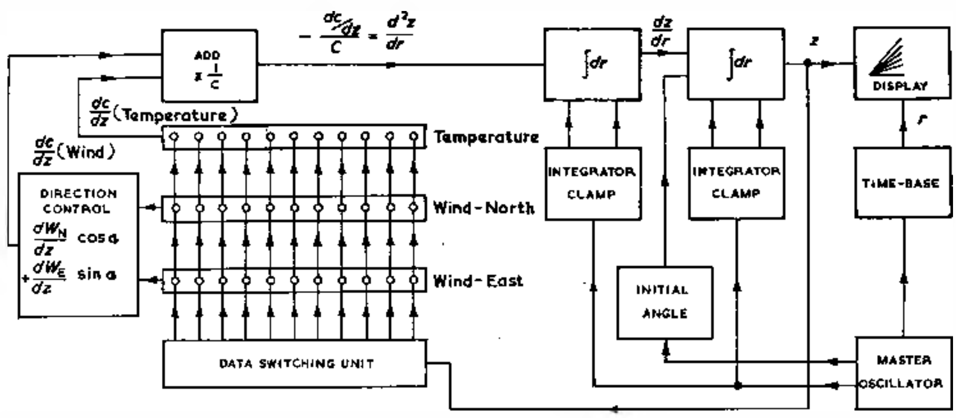


Fig. 1. General arrangement of computer

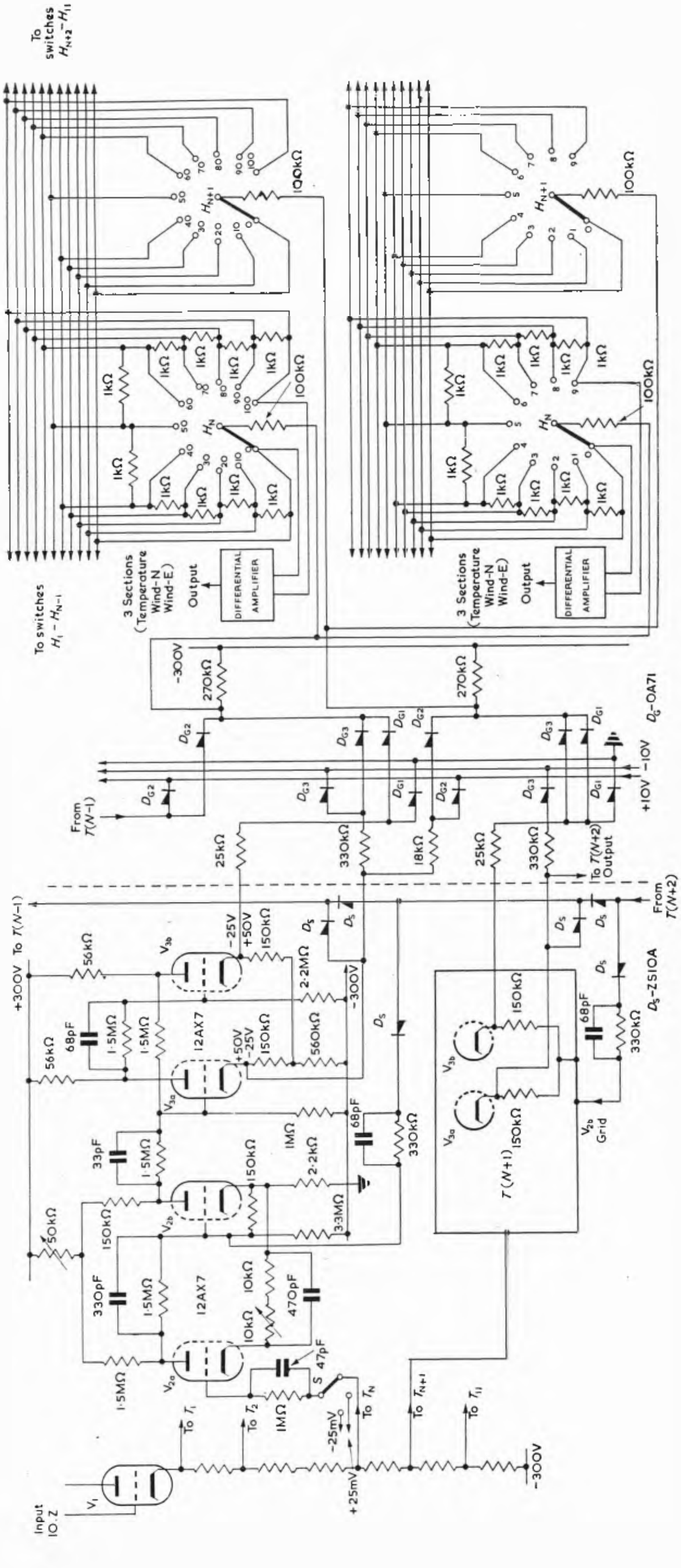


Fig. 3. Switching and data handling units

Integrators

The two operational amplifiers which perform the integration are of conventional design. The input stage uses a 6J6 double triode under semi-electrometer conditions, followed by two high gain pentode stages, and a cathode-follower output stage. Drift-correction is provided by a two stage chopper type a.c. amplifier, the output of which is fed to the second grid of the 6J6. The diode integrator clamps, which operate during the reset period are similar in design to those described by McKay[1] and are capable of completely discharging the plastic film capacitors, even when these are charged to the amplifier limit of ± 50V.

The operational amplifiers used for adding, and sign reversal in the rest of the computer, are of a simpler one stage type, but their effective gain is increased by positive feedback to the cathode of the input stage. Drift due to line voltage variations is reduced to acceptable limits by supplying the valve heaters from a constant voltage transformer.

Switching Unit

The switching unit incorporates a number of modified Schmidt triggers, set to function as the computed solution z passes through the various altitudes at which the meteorological information is provided. The triggers are linked to diode switches which feed potentials to the resistor network of the data handling panel, in order to produce the gradients of temperature and wind velocity.

The circuit diagram of the switching and data handling panel is indicated in Fig. 3. The output z of the second integrator is connected to a sign reversing amplifier having a gain of 10. This amplifier has a cathode-follower output stage V_1 with the cathode load taken to -300V. Eleven tapping points are arranged on this load to correspond to the eleven altitudes at which the meteorological data is available; so that as the solution z passes through any one of these values the potential of the corresponding tapping point is at zero. The triggers are connected, one to each tapping point and are arranged to operate at zero potential. The modified version of the Schmidt circuit uses a 12AX7 double triode V_{2a} and V_{3b}, and has a backlash of as little as ± 25mV. This degree of sensitivity, combined with the effect of multiplying z by 10, results in a switching accuracy equivalent to a height of $\pm$ 2ft 6in.

Bistable multivibrators represented by the valves V_{3a} and V_{3b} are connected to the triggers. The cathodes of these valves provide the potential swing from $+50$ to -25V and vice versa, used to operate the diode switches. In addition, the potential of $+50$V which appears at the cathode of V_{3a} when in the operated condition, is fed to a network of silicon diodes (designated as D_s on the diagram) connected to the grids of the valves V_{2b}, and serves to switch off all the stages driven from those tapping points on the cathode load of V_1 which stand at a positive potential. This ensures that only one section operates at a time.

Adjustment of the trigger circuits is facilitated by the switch S which provides a potential of ± 25mV; a set of neon lamps connected between the cathodes of the valves V_{3a}, and a potential of -50V is used to indicate the state of the triggers.

When a trigger and its associated multivibrator is in the 'off' condition either because the input potential is negative or because a later stage is 'on', the cathode of V_{3b} is at $+50$V and the cathode of V_{3a} is at -25V; the germanium diodes D_{g1} conduct and the potential fed to the data networks is zero. When a stage (for example T_N) is in the 'on' position the V_3 cathode conditions are reversed, and the 50V appearing at the cathode of V_{3a} causes the diodes D_{g3} of the T_N stage, and D_{g2} of the T_{N+1} stage to conduct, thus producing -10 and $+10$V respectively to be applied to the data resistor network.

The height intervals at which the data is supplied are of three values, 400, 500 and 1 000ft. The ± 10V represents the 500ft intervals, $\pm 12{\cdot}5$V is used for the 400ft interval and ± 5V for the 1 000ft intervals. These potentials are obtained from cathode-followers driven from tappings on a resistor chain connected between h.t. positive, earth and h.t. negative.

The Data Handling Panel

The three parameters (temperature and the two wind components) are set up on the controls using a decade system incorporating a row of 'tens' and a row of 'units'. The temperature range covers from 0 to 109°F, and the wind -50 to $+50$ft/sec. For wind speeds greater than ± 50ft/sec a two-position switch is incorporated which alters the sensitivity of the controls and the wind resolver, giving an extended range from -100ft/sec to $+100$ft/sec in 2ft/sec steps.

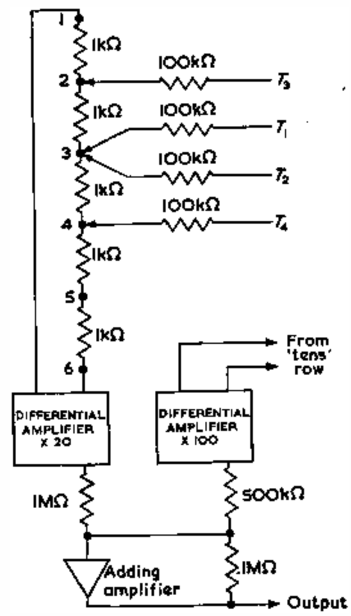


Fig. 4. Circuit of one row of switches

The wind resolver is fitted into the lower section of the panel and the two meter readings, indicating wind-north and wind-east components are transferred by the operator to the corresponding rows of controls. The 'tens' controls consist of 1 pole, 11 position switches, and the 'units', 1 pole, 10 position switches. The switches in each row are wired in parallel, with 1kΩ resistors connecting adjacent contacts. The outputs from the data switching unit T_1 to T_{11} are fed to the movable contact of the switches at the appropriate height, via 100kΩ resistors. The first and last contacts of the switches of each row are connected to a differential amplifier, the output of which is a measure of the difference in settings of the switches. A simplified circuit of one row of switches, for four heights, is given in Fig. 4. This shows six positions of the 0 to 9 units switch, the dial settings for the four heights being 3, 3, 2 and 4 respectively. Consider first the case when T_1 is switched on. This produces $+10$V at T_2 and -10V at T_1; current flows through the two 100kΩ series resistors, and no potential appears at the input of the differential amplifier. When T_2 is switched on, T_3 becomes +10V, T_2 −10V and current flows through the two series resistors, and the 1kΩ resistor between positions 2 and 3, producing a potential drop of 100mV across it. This is amplified by the differential amplifier, which has a push-pull gain of 20, to give a single sided output of −1V indicating a gradient of −1 units. Likewise, when T_3 operates +10V appears at T_4 and −10V at T_3, there is a potential drop between points 2 and 4 of −200mV, producing an output of +2V representing a gradient of +2 units.

The advantage of this method is that since the actual value is not represented electrically, small differences between large or small values are measured with equal accuracy. It also ensures that when the gradient is zero the output is zero. There are, however, two inherent errors. The first is that the potential drop e_n across n sections is given by the expression:

$$e_n = (10 + 10)\left[\frac{1K.n}{2(100K) + 1K.n}\right] \text{ volts}$$

whereas the required value is:

$$20n/200$$

the resulting error is:

$$\frac{100n}{200 + n} \text{ per cent} \simeq n/2 \text{ per cent}$$

The second error is caused by the effect of the switched-off sections, and is due to resistive attenuation of the required potential. It depends on the values of the settings, and therefore varies with each set of conditions. The worst possible condition occurs when the earthed sections are equally divided between the two extremes of the range, but since all the errors are negative they can be kept within ±5 per cent in the majority of cases.

The differential amplifiers are of the a.c. type, designed to have a long overall time-constant, and a gain stabilized at the required value of 20 for the 'units' sections, and 100 for the 'tens' sections by the use of considerable negative feedback. The amplifiers contain two high gain pentode stages followed by a cathode-follower output stage. Single ended outputs are capacitor-coupled to the adding amplifiers. The d.c. level of the output is restored during the 'reset' period, when the pulse from the master oscillator is used to switch off all sections of the data switching unit, and simultaneously to operate a series of diode clamps which discharge the output coupling capacitors to earth. This is simpler and more reliable than using d.c. differential amplifiers with the necessity for elaborate drift correction.

The Bounce Generator

Normally, a ray which returns to earth switches off the T_1 stage, and continues in a straight line off the lower edge of the c.r.t. screen. An additional facility is incorporated, however, to simulate the reflection of a ray at ground level.

The output Z of the second integrator is connected to a sensitive trigger circuit of the same type as used in the data switching unit. When the output potential goes negative (i.e. Z goes below ground level) the trigger operates two diode switches, one of which reverses the sign of the Z co-ordinate, while the other reverses the sign of the effective sound speed gradient. Thus, in effect a mirror image of the gradient, reversed in sign is reproduced for negative values of Z, and the main display is consequently modified to produce a series of bounces. Photographs of a typical display with the bounce generator in circuit, are given in Fig. 5.

The Complete Assembly

The complete computer is shown in the photograph on page 398. All the operator controls are situated on a sloping panel to the rear of a writing surface. The main 4½in square, long persistence display tube is placed centrally, and the whole of the space to the left is occupied by the data handling panel. On the right is the direction control which carries an eight inch diameter Perspex disk. On this is engraved a cursor with calibration marks every five

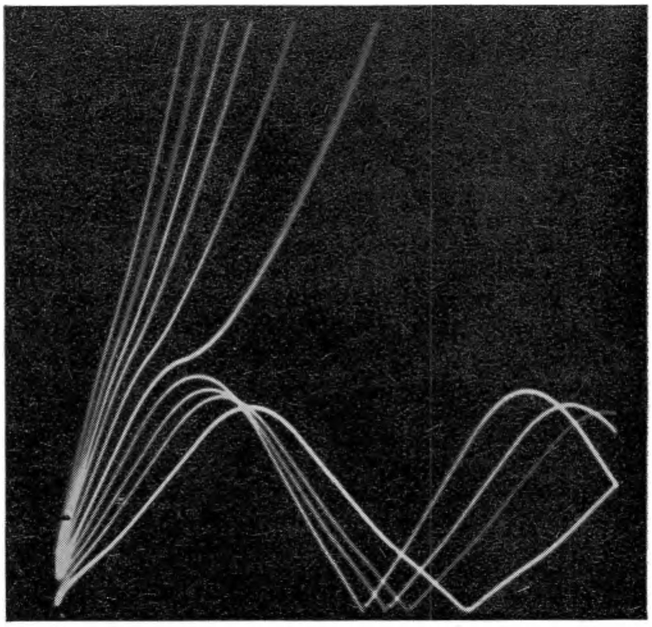


Fig. 5. A typical ray display

miles, corresponding to similar markings on a Perspex shield over the face of the display tube. This enables the displayed ray pattern to be directly referred to a map of the area mounted beneath the direction control disk. There are two small controls above the map and direction control, an on/off switch for the bounce generator, and a selector switch which enables either 1, 5 or 10 rays to be displayed. With the aid of these two controls it is possible to make a detailed examination of a complex ray pattern. The two further controls below the map are the manual adjustment for the initial angle of the rays, and a switch for removing the gradient input.

Further to the right is the gradient monitor panel, which contains two 3½in cathode-ray tubes. The upper tube displays the total effective gradient fed into the first integrator plotted as a function of the height co-ordinate; the lower tube displays any one of the three component gradients chosen by a selector switch. The two integrating amplifiers, the sequencing and control circuits, and the data switching unit are mounted behind the left-hand door. The back of the right-hand section houses the power packs and the front cupboard is fitted with shelves for the storage of meteorological data sheets.

REFERENCE

1. MacKay, D. M. High-Speed Electronic-Analogue Computer Techniques. *Proc. Inst. Elect. Engrs.* 102B, 609 (1955).

Transistor Circuits for a 1Mc/s Digital Computer

By I. Krajewski*, B.Sc.

The requirements of amplifiers used for pulse regeneration are considered and blocking oscillators and regenerative amplifiers employing transistors suitable for reshaping and retiming pulses in a digital computer with a 1Mc/s digit rate are described. Some details of their practical application are given.

(Voir page 441 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 446)

IN certain kinds of digital computer large numbers of amplifiers are used for pulse regeneration. Such computers propagate information as trains of pulses, a train being divided into definite intervals, within each of which can be accommodated a group of pulses. A group of pulses like this is usually called a ' word ', because it may be an instruction to the computer or it may be a factor in a computation. In a serial computer working in pure binary, the word interval can be considered as subdivided into a number of digit intervals; if a pulse occurs during one of these, it represents a binary 1 or 0. The significance of this binary digit depends on the timing of the pulse relative to the start of the word.

Logical operations in computers of this sort involve the comparison of the pulses forming two or more words and the generation of other pulses therefrom according to the logical rules of the operation. Such comparisons of pulses are spoken of as their ' gating together '. The gating may be performed by diodes. To satisfy the rules for a given computer function (e.g. addition) may involve passing the pulse trains through a complex arrangement of gate circuits. So that the waveform distortion and attenuation shall not be cumulative, amplifiers are interposed where necessary between the gate circuits. These amplifiers yield outputs of the desired waveform and amplitude, but they cannot advance the timing of their outputs to offset the delay introduced by the cascade of gate circuits. This difficulty is overcome by retiming the output waveform relative to a reference. The reference is the computer's basic ' clock waveform '. The clock waveform has each of its cycles coincident with the digit intervals of an undelayed word; usually the clock waveform is a square wave. The output from each amplifier is arranged to coincide with a clock pulse; consequently a given digit pulse in a word coincides with successively later clock pulses as the word passes through successive amplifiers. This imposition of delays on words as they pass through the gate and amplifier circuits impinges on the logical design of arrangements needed to satisfy the rules for any chosen function. Such difficulties are most acute where the words which are to be compared pass through different numbers of gating circuits; then it is necessary to introduce an artificial delay in one or more circuits to ensure that corresponding digits in the words are compared.

An amplifier of the kind referred to above has to:

(1) Amplify and reshape its input pulse.

(2) Yield an output of defined waveform and amplitude.

(3) Yield its output coincident with a clock pulse.

(4) Satisfy requirements (1) to (3) at the computer's clock frequency and in such a way that successive digits have no interaction.

It is desirable that it should also:

(5) Produce a high power to allow the output pulse to pass through a great many gate circuits before further amplification is required.

(6) Accept a very badly degraded waveform as its input.

Because a computer will use very large numbers of such amplifiers, each should be economical of both d.c. power and clock pulse power. The amplifier should also be reliable, simple, easy to mass produce and require the minimum of adjustment between being made and being fit for use in a computer.

Blocking Oscillator

A blocking oscillator can be used to meet, to a large extent, the requirements for an amplifier set out above. The

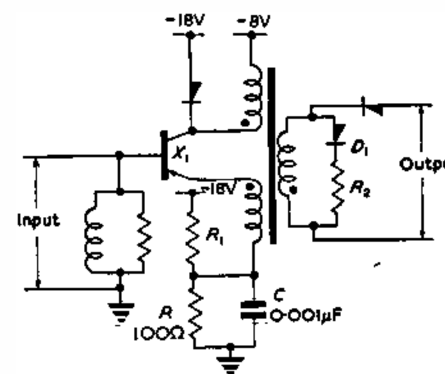


Fig. 1. Blocking oscillator employing collector-to-emitter feedback

only blocking oscillators considered here are those with either collector-to-emitter or collector-to-base feedback; and these are only discussed briefly because they have been fully described before[1,2]. An approximation to the maximum pulse duration is derived and some notes on triggering are given before the application of the circuits is considered. Fig. 1 shows the circuit diagram for a collector-to-emitter feedback blocking oscillator; Fig. 2 refers to a collector-to-base arrangement. In each circuit R and R_1 reverse bias the emitter by 0·1V to ensure stability at elevated temperature. The transformers are critically damped by D_1 and R_2.

PULSE DURATION

The circuits have two time determining mechanisms; one associated with the emitter network and the other with the collector. When such circuits are used in a computer whose clock pulse frequency is 1Mc/s (i.e. 1 digit interval is 1μsec), these mechanisms must determine the pulse duration at about 0·5μsec; furthermore, energy stored in the networks during the pulse must discharge during the second 0·5μsec of the digit interval. The following analysis shows

* *International Computers and Tabulators Ltd.*

the effect of external conditions upon the pulse duration through the time determining network in the collector. The effect of the emitter network is neglected in the analysis. This approximation will result in the calculated pulse duration being greater than that actually obtained. The magnitude of the error will depend upon the relative values of the times determined by the two networks.

The equivalent circuit of Fig. 3 is used to analyse the operation of the time determining network in the collector circuit. If R_{in} is defined as the resistance of the base-emitter junction then:

$$I_b = (V_{be}/R_{in}) = V_f/R_{in}$$

$$= \frac{V_p}{N_1 R_{in}} \text{ since } V_f = (V_p/N_1)$$

and since $I_c = \alpha I_b$

$$I_c = \frac{\alpha V_p}{N_1 R_{in}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (1)$$

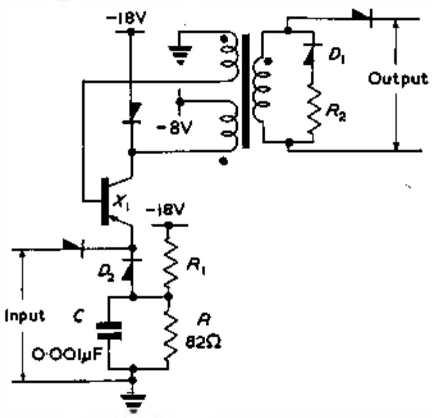


Fig. 2. Blocking oscillator employing collector-to-base feedback

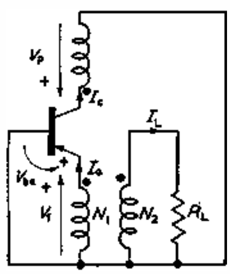


Fig. 3. Illustrating the notations used in deriving the expression for pulse duration

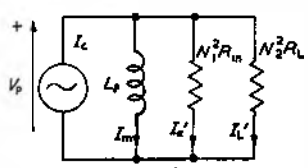


Fig. 4. Simplified equivalent circuit used in deriving the expression for pulse duration

Fig. 3 may be replaced by Fig. 4 where the currents and load resistances are all referred to the primary (i.e. collector) winding.

Then:

$$I_c = I_m + I_e' + I_L'$$

whence:

$$I_m = I_c - I_e' - I_L' \quad \ldots\ldots\ldots\ldots\ldots \ (2)$$

and for the separate arms:

$$I_e' = \frac{V_p}{N_1^2 R_{in}} \text{ and } I_L' = \frac{V_p}{N_2^2 R_L}.$$

Substituting these values and that for I_c from equation (1) in equation (2):

$$I_m = V_p \left[\frac{\alpha N_1 - 1}{N_1^2 R_{in}} - \frac{1}{N_2^2 R_L} \right] \ldots\ldots\ldots\ldots \ (3)$$

Now for the inductive arm in Fig. 4:

$$V_p = L_p (dI_m/dt)$$

integrating:

$$V_p t = L_p I_m + k$$

when $t = 0$, and therefore $k = 0$.

Thus:

$$I_m = (V_p t/L_p) \quad \ldots\ldots\ldots\ldots\ldots \ (4)$$

Equations (3) and (4) can be used to determine the pulse duration in so far as it is controlled by the collector network, since equation (3) defines the condition where the collector current available from the transistor is just adequate to supply the load, feedback and magnetizing currents; and equation (4) shows the latter as an increasing function of time.

Combining equations (3) and (4) and putting $t = T$, the pulse duration:

$$T = L_p \left[\frac{\alpha N_1 - 1}{N_1^2 R_{in}} - \frac{1}{N_2^2 R_L} \right] \ldots\ldots\ldots\ldots \ (5)$$

Equation (5) shows the dependence of pulse duration on loading. The difficulty of designing a blocking oscillator to operate satisfactorily in a computer, where the load may change from open-circuit to full load from one digit period to the next is evident.

Triggering

Triggering blocking oscillators can present a considerable problem in computers, where a mass produced blocking oscillator circuit assembly has to be capable of operating correctly when used in any amplifier position in the machine. It is desirable that the circuit should trigger correctly and operate uniformly regardless of the source impedance of the triggering waveform.

It is desirable that the triggering waveform be fed to a part of the circuit which is not directly in the feedback path. The impedance across which the triggering waveform is developed in the blocking oscillator must not introduce an appreciable time-constant.

Figs. 1 and 2 show two methods which have proved satisfactory in experimental equipment operating at 1Mc/s. The arrangement of Fig. 1 uses an inductance of a few microhenries in the base circuit; the arrangement in Fig. 2 uses diode D_2 across which to develop the triggering waveform.

Pattern Generator

It is convenient to be able to generate a repetitive pulse train which comprises only one word, but each of whose digit intervals can be occupied by a pulse or not at will. It is convenient for experimental work that a pulse pattern generator of this kind should be capable of being synchronized to an external clock waveform. Fig. 5 indicates the circuit for such a pattern generator using blocking oscillators with emitter-to-collector feedback. An instrument of this type with a word length of 25 digits has been in satisfactory use for more than a year; for simplicity a 6-digit version is indicated in Fig. 5 and described below.

Transistor X_1 is a free running blocking oscillator producing a 0·4μsec pulse approximately every 6μsec. The repetition rate is determined by C_1, R, R_1 and the potential to which they are returned. The output of X_1 is delayed for 1μsec by the network D_1, C_1, D_2 and then triggers X_2 which is a triggered blocking oscillator producing a 0·4μsec output pulse. The output from X_2 triggers X_3 in the same way as before, and so on until X_6 produces its output pulse 6μsec after X_1 produced the first pulse. The pulse from X_6 is fed back to the base of X_1, whose operation is thereby synchro-

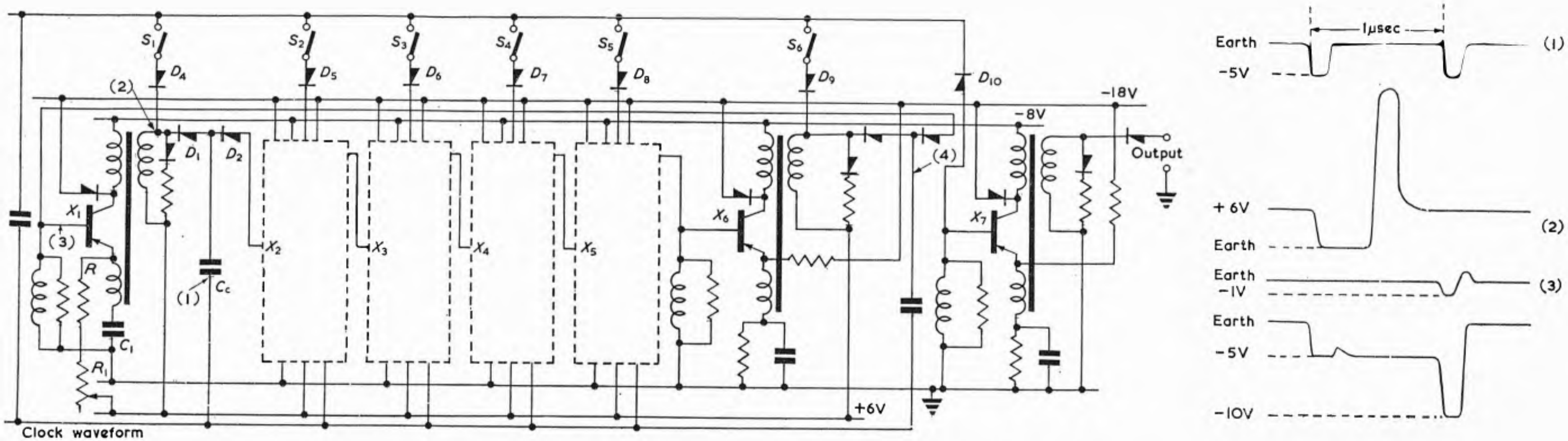


Fig. 5. Six-bit-word pattern generator. Idealized waveforms are shown to clarify the action

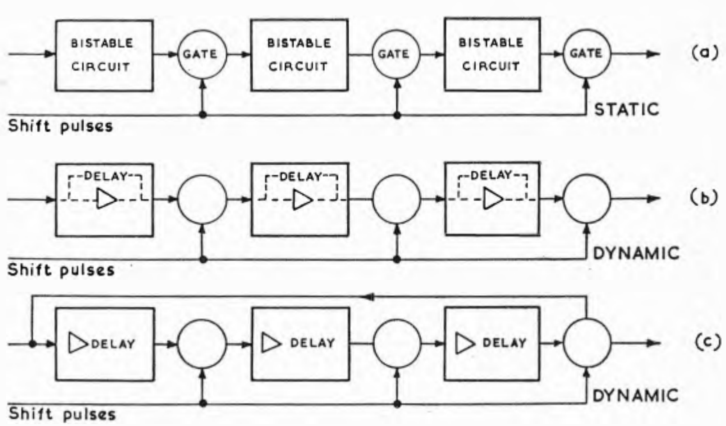


Fig. 6. Contrasting a static method with two dynamic methods of storage to show the need for a circulating loop in dynamic storage

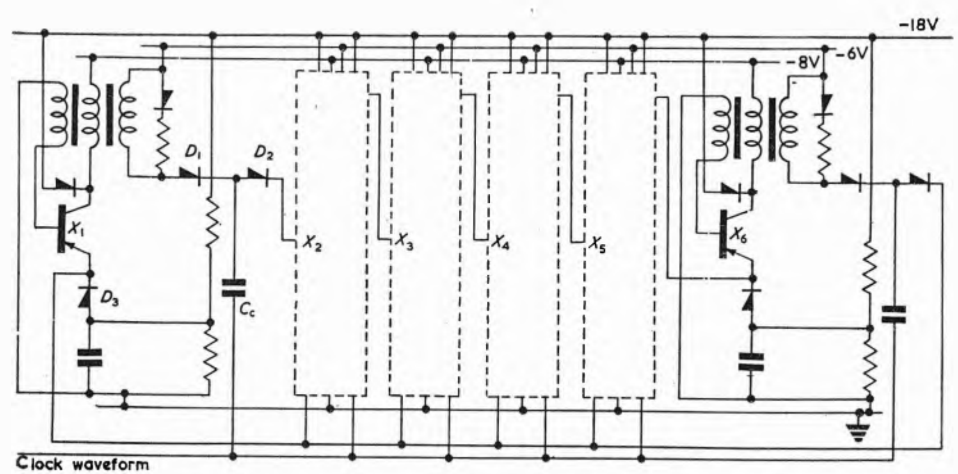


Fig. 7. A six-stage shifting register employing the storage method indicated by Fig. 6(c).

nized. The action can now be considered to be a single pulse of 0·4μsec duration circulating in a loop which it takes 6μsec to travel round. As the pulse 'passes' each of the switches S_1 to S_6, it can be fed via the combining diodes D_4 to D_9 to trigger X_7. The circuit of X_7 is triggered or not in each of the 6-digit intervals of the word according to how the switches are set.

Shifting Register

A shifting register is a cascade of storage devices, each capable of holding one digit. The digits can be shifted from one storage device to the next by applying a pulse to a gate circuit between the storage devices. Usually the gate circuits employ diodes; the storage devices are usually thought of as being of two kinds; static (e.g. a bistable circuit) or dynamic. Dynamic circuits store the pulses representing the digits by circulating them in a loop; sometimes there is a loop for each storage device, as in Fig. 6(b); sometimes there is one loop for the whole register, as in Fig. 6(c).

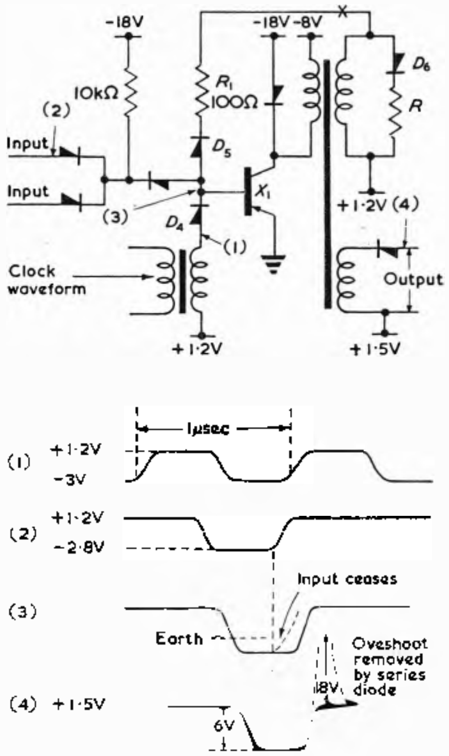


Fig. 8. Regenerative amplifier with input gating circuits

The idealized waveforms show how the output waveform's duration is determined by the clock waveform and not by the input waveform's timing

Fig. 7 shows the circuit diagram for part of a shifting register using dynamic storage of the kind indicated in Fig. 6(c). Each storage element uses a blocking oscillator with base-to-collector feedback. Pulses representing digits are delayed and transferred to the next stage by the same diode-capacitor arrangement as used in Fig. 5. Pulses fed into the register at the base of X_1 will be preserved by circulation and can be fed out as desired. If necessary for logical reasons to satisfy the rules of some function (e.g. forming the complement of the number stored), the preserve loop between the two 'end' transistors can include gate circuits.

Regenerative Amplifiers

Regenerative amplifiers are an alternative to blocking oscillators for reshaping and retiming pulses after their passage through gate circuits. The amplifiers described below are derived from the SEAC circuit, so called after the computer in which it was first used.

Much has been written[3] about regenerative pulse amplifiers for computers using point contact transistors. The details below relate to such amplifiers using junction transistors and intended for use at a digit rate of 1Mc/s. All of the amplifiers described use the common emitter configuration, because it results in the highest power gain. To increase the gain and stabilize the pulse duration, external positive feedback is applied. Maximum efficiency is ensured by operating the transistor from cut-off to bottoming.

All of the amplifiers use transformer coupling, whose advantages over resistance-capacitance coupling include:

(1) Ready provision of both polarities of output.

(2) Easy impedance matching.

(3) Greater freedom in arranging the transfer of energy. In turn this allows a better control over the d.c. levels of variable length trains of pulses.

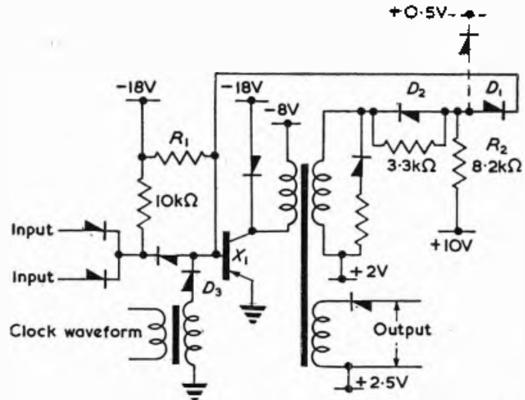


Fig. 9. Regenerative amplifier employing gated feedback

The diode shown dotted can be used to eliminate breakthrough of the clock waveform to the output

This last advantage can only be fully realized if the transformer is critically damped, i.e., the energy stored in the mutual inductance must be completely withdrawn before the next pulse is fed to the amplifier. With a 1Mc/s digit rate using 0·4μsec pulses, it is desirable to withdraw all the energy in the 0·4μsec following the pulse, to allow the circuit a stable 0·2μsec before the next digit interval starts.

Fig. 8 shows the circuit of a simple regenerative amplifier whose input reaches it via an AND gate followed by an OR gate. In a practical computer, the amplifier input circuit would usually embody several such gate circuits. The amplitude of the output is controlled by the circuit parameters; the timing is controlled by the clock waveform fed to the base via D_4. During the positive half cycles of the clock waveform X_1 is cut off; during the negative half cycles X_1's potentials allow it to amplify.

Inputs are timed always to coincide with negative half cycles of the clock waveform. If an output is produced only when the input coincides with the negative half cycle of the clock waveform, the duration of the output pulse will vary in sympathy with the input timing. To obviate this, the circuit arranges that the transistor conducts for the whole of any negative clock half cycle during which an input appears. The feedback path embodying D_5 and R_1 ensures this. When an input causes an output, this output is fed via R_1 and D_5 to the base where it simulates a continuation of the input until the transistor is cut off by the next half cycle of the clock waveform. Thus, the output duration is controlled by the clock. Diode D_5 isolates the input pulse from the feedback path; R_1 limits the feedback current.

Because the output circuit of Fig. 8 presents a high impedance at 1Mc/s, the circuit suffers from breakthrough of the clock waveform to the output. This can be minimized by clamping the point x with a diode whose cathode is returned to the + 1·2V rail.

Another approach is to use gated feedback. This is shown in Fig. 9, in which the dotted diode should be ignored at present. The action of the circuit depends on switching the current in R_1 between D_1 (between pulses) and the base (during pulses). Between input pulses, the current in R_1 flows by the path R_2, D_1, R_1. This current holds the base at

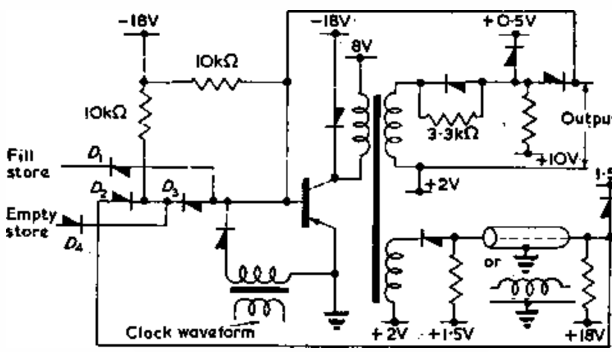


Fig. 10. The circuit of Fig. 9 is used here in a serial store

a positive potential. When an output of amplitude enough to overcome the reverse bias on D_2 is produced, D_1 is cut off; the current previously flowing from R_1 to D_1 is now diverted to the base. This circuit as so far described suffers from some breakthrough of clock waveform to the output. The diode shown dotted clamps the output level, and eliminates this.

Using OC44 or XA102 transistors, the circuit of Fig. 9 has been used to deliver 20mA, 5V pulses of 0·4μsec duration to an external load. The maximum dissipation in the transistors is well within their power ratings, although the currents and voltages are higher than the small signal ratings. New types are becoming available which allow operation entirely within rating. A Mullard ferrite pot core type FX1011 is used for the transformer, which has 10-turn feedback and output windings and a 13-turn collector winding. All of the windings are single layer, one on top of the other.

Serial Store

One application of the circuit of Fig. 10 is in the construction of a serial store in which pulses representing digits are circulated in a length of delay cable by the amplifier. Fig. 10 shows such an arrangement. The pulse train representing one or more words to be stored is fed to the base, as in Fig. 10. After reshaping and retiming the pulses appear across the output windings; thence to the delay cable whose length determines the capacity of the store. The deformed pulses from the output of the delay cable are fed via the gate D_2D_3 to the base. After reshaping and retiming, the cycle repeats—the pulse train circulates in the loop and is stored.

If D_4 conducts, the output of the delay cable will be shunted and pulses will not reach the base. This is used to clear the store. If a positive pulse is fed to the anode of D_4 during every digit interval, the store will be completely cleared. The store can be partially cleared by feeding the desired number of positive pulses to D_4; alternatively, a positive rectangular waveform can be applied, starting and ending coincident with the first and last pulse to be cleared from the store.

Acknowledgments

Acknowledgment is made to International Computers and Tabulators Ltd. for permission to publish this article. The contributions of Messrs. J. S. Walker and J. Redrupp and the encouragement of Mr. M. L. N. Forrest are acknowledged as is the assistance of Mr. S. C. Murison in the preparation of this article.

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An Electronically Controlled Drilling Machine

A new electronically controlled drilling machine capable of accurately positioning and drilling holes up to 2in. in diameter in thick steel in a fraction of the time taken by a skilled operator working a conventional radial drill, has been developed by E.M.I. Electronics Ltd and Wadkin Ltd.

Known as the E.M.I./Wadkin Electronic Positional Drilling Machine Model TCD 1, the equipment has a large worktable measuring 3ft 6in by 5ft 6in and has been designed specially to exploit the high degree of accuracy provided by the E.M.I. control system.

One of the features of the machine is the fact that the drill head is mounted on a cross beam supported on twin vertical columns. This provides extra strength and enables drills up to maximum diameter to be fed directly into steel without the 'breakthrough' problem normally experienced.

Unlike conventional drills, it is not necessary to go through the lengthy procedure of drilling a small pilot hole, followed by progressively larger ones, until the required size is reached. Large holes can thus be drilled to very fine tolerances in a fraction of the normal time.

As with E.M.I.'s other control systems, this machine is automatic and operates from either punched tape or dial settings. No marking out or drilling jigs are required. Nor is an external computer needed to carry out even complicated programming, as the E.M.I. system incorporates what is in effect a 'built-in' computer.

In operation a table of x and y axes dimensions is prepared. The first tabulated x and y dimensions are the alignment points necessary to enable the accurate location of the workpiece with respect to the table datum. These are followed sequentially by the x and y dimensions of all the holes to be drilled. When practicable, all holes of the same dimension should be grouped to minimize the number of tool changes.

The table of dimensions is typed-out on a modified teleprinter which produces the punched paper tape used to control the machine.

The workpiece is then located on the table relative to the table datum using the programmed alignment points, these may be machined faces, cast bosses or specially machined flats.

The correct tool is fitted, speed, feed, and depth set and the start button pressed, the machine will then commence and automatically cycle through the first group of the programme. After the first cycle has been completed the machine automatically stops to enable the next tool to be fitted before the next series of operations.

For simple 'one off' jobs, it may be considered impractical to produce a tape and for this reason dials are provided to individually set the co-ordinates required.

The machine represents a good example of collaboration between the electronic and machine tool industries, experience having shown that it is only by developing a machine tool to meet the requirements of electronic control that full advantage can be taken of the new technique's potentials.

Approximate Methods for Calculating the

Behaviour of Square Loop Magnetic Cores in Circuits

By D. A. H. Brown*, M.A.

The switching time in a square loop magnetic material is inversely proportional to the excess of driving m.m.f. over the coercive m.m.f. This well-known result is combined with the m.m.f. equation to calculate the switching time and secondary current when the core has a loaded secondary winding. The switching time is shown to be the sum of (1) the open circuit switching time with the primary drive applied, and (2) the ratio of flux change to applied voltage. Two applications of the theory (1) to interconnected cores and (2) to conditions when the drive is not from a pure current source are then given.

(Voir page 441 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 446)

THE small ferrite core has become at the present time a versatile circuit element both as a means of data storage and for switching applications. The makers' data for these cores usually specify certain voltage outputs when subjected to current pulses of given amplitude and duration. However, the cores are often used in circuits where the conditions are very different, and it is then important to have some guide as to the probable behaviour. Qualitatively the general effects may be obvious, but a quantitative theory is desirable and a simple theory is developed here. It is recognized to be approximate only, but it has served to co-ordinate a number of different cases of core circuits and to show that the effects follow from basic principles.

Standard transformer theory cannot be applied directly to a square loop core since it is no longer true that flux density and magnetizing current are proportional. First, the simple modification necessary to take the square loop characteristic into account, and the resulting switching time of a core having a loaded secondary winding are considered.

The switching time is known to vary with the driving current and the modifications to take this fact into account are discussed. The results of some simple experimental tests are given and two cases are considered where the theory has been applied to practical problems.

Symbols Used

B	= flux density
B_s	= saturation flux density of square loop material
F_c	= coercive m.m.f. corresponding to H_c
H	= magnetizing force
H_o	= coercivity
H_c	= threshold magnetizing force for switching
I_1	= primary current
I_2	= secondary current
I_3	= programme current pulse (Fig. 5)
l	= mean length of magnetic circuit in a core
N_1	= primary turns
N_2	= secondary turns
R	= load resistance
S	= switching constant (equation (5))
α	= cross-sectional area of a core
τ	= switching time
ϕ	= flux
ϕ_s	= saturation flux
ϕ_1	= flux in storage core (Fig. 5)
ϕ_2	= flux in switch core (Fig. 5)

* *Royal Radar Establishment.*

The Loaded Core

Consider a core having the ideal square loop B-H curve of Fig. 1, and used in the circuit of Fig. 2 where it is loaded by a resistance across its secondary winding. Assuming for the moment that flux follows magnetizing force instantaneously then, during any flux change, H is always equal to $\pm H_c$. One can write therefore for Fig. 2,

$$N_1I_1 - N_2I_2 = F_c \quad \ldots\ldots\ldots\ldots\ldots (1)$$

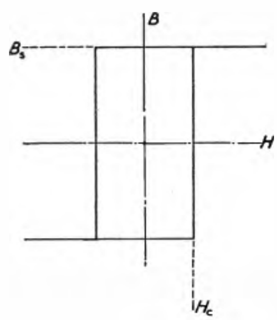

Fig. 1. Ideal square loop B-H curve

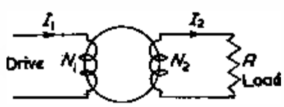

Fig. 2. Core with loaded secondary winding

In a normal pulse transformer the net magnetizing m.m.f. $N_1I_1 - N_2I_2$ is zero at the start of the pulse but builds up with time. In this ideal square loop case there will be no time variation of magnetizing current while the flux is changing. This is equivalent to the statement that the magnetizing current is constant and equal to F_c during any change of flux. Equation (1) contains no reference to the load and shows that the output current is determined solely by the core characteristic F_c, winding N_1 and N_2, and the drive current.

The second equation which must be applied is

$$N_2\,(d\phi/dt) = I_2R \quad \ldots\ldots\ldots\ldots\ldots (2)$$

By eliminating I_2 from equations (1) and (2) the switching time can be found (from $-\phi_s$ to $+\phi_s$) as

$$\tau = (2\phi_s/R)\,\frac{N_2^2}{(N_1I_1 - F_c)} \quad \ldots\ldots\ldots\ldots (3)$$

Equations (2) and (3) show that, as R becomes large, the rate of flux change increases and the switching time decreases, the limiting case being $\tau = 0$ for $R = \infty$. It is in fact known that there is a lower limit to the switching time of about one microsecond on open-circuit. The equations of this section must therefore be approximate, but they do form a useful basis for calculation when the switching time is long compared with the open-circuit switching time, under the conditions of small R. This is the condition of the switching being controlled by the external circuit rather than by the limits of the internal domain changes. The following section considers a method of taking the limiting switching time set by the internal domain changes into account.

An Equation for the Internal Switching Mechanism

One of the main characteristics of a square loop magnetic core, whether of ferrite or of nickel iron alloy, is the linear relation between the reciprocal of the switching time on open-circuit and the driving current which may be written as:

$$1/\tau \propto (H - H_0) \quad \ldots\ldots\ldots\ldots\ldots (4)$$

H_0 represents a threshold below which no switching takes place. It is closely related to the coercivity H_c though the dynamic value of H_0 obtained from experimental curves of $1/\tau$ against H may not be exactly the same as the d.c. coercivity. For many purposes it is sufficient to regard H_c and H_0 as the same and this will be done here.

It is remarkable that this relation holds for both ferrite and iron alloy cores since the switching mechanism is quite different in the two cases. In metal tape cores the switching is entirely eddy current controlled[1], but in ferrites the resistivity is much too high for eddy currents to be important and the switching is thought to be limited by the movement of domain walls[2,3]; the velocity of a domain wall is proportional to $H - H_0$.

An equation based on this domain wall motion is

$$(dB/dt) = S(H - H_c) \quad \ldots\ldots\ldots\ldots (5)$$

This implies equation (4) provided H is considered constant in the core; in practice, when H is provided by current in a straight wire threaded through the hole in the core, H will vary across the core inversely as the radius. Ridler and Grimmond[4] have taken this effect into account in their work but it leads to complicated expressions. If H is considered constant at the mean value in the core, equation (5) leads to quite simple and useful results.

If a step function of H ($> H_c$) were applied to a core obeying equation (5) then it would give a constant rate of change of flux and therefore a rectangular voltage pulse. This is clearly an idealization.

The open-circuit switching time can be calculated, using equation (5), as

$$\tau = \frac{2\phi_s}{(\alpha S/l)(N_1I_1 - F_c)} \quad \ldots\ldots\ldots\ldots (6)$$

Equation (1) must be modified since the net magnetizing m.m.f. is no longer restricted to F_c; it can, and in fact it must, rise above F_c and it is this excess which determines the switching speed. Instead of equation (1) therefore:

$$N_1I_1 - N_2I_2 = lH \quad \ldots\ldots\ldots\ldots (7)$$

This allows one to calculate the general switching time of a loaded core (Fig. 2). Eliminating H from equations (5) and (7) gives

$$d\phi/dt = (\alpha S/l)(N_1I_1 - N_2I_2 - F_c) \quad \ldots\ldots (8)$$

and eliminating I_2 between this and equation (2), which still holds good:

$$d\phi/dt\,[1 + (\alpha SN_2^2/lR)] = \alpha S/l\,(N_1I_1 - F_c)$$

and therefore the switching time is:

$$\tau = 2\phi_s\left[\frac{1}{(\alpha S/l)(N_1I_1 - F_c)} + \frac{N_2^2}{R(N_1I_1 - F_c)}\right] \ldots (9)$$

Equation (8) is interesting as showing that the switching time is the sum of the two terms; the first is the natural switching time on open-circuit and the second is that calculated for the loaded core by the simple theory given earlier.

The secondary current which will flow for a given primary drive N_1I_1 may be found by eliminating H between equations (5) and (7). The result is

$$N_2I_2 = N_1I_1 - [F_c + (l/\alpha S)\,.\,(d\phi/dt)] \quad \ldots\ldots (10)$$

and shows that for the purposes of current output the core may be regarded as having an effective coercivity of H_c plus a term depending on the rapidity of turnover. The simplest way of considering it is that part of the drive overcomes the static coercive force, part provides the excess over this necessary to switch the core, and the rest is available for output.

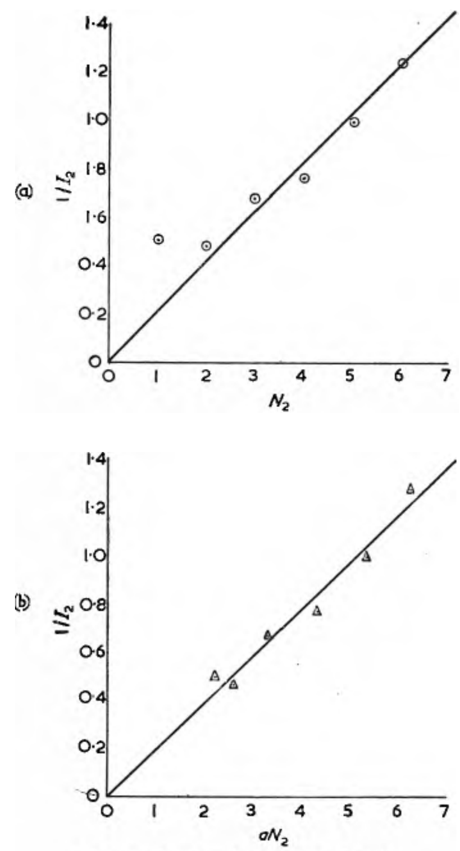

Fig. 3. Results of experimental tests

Experimental Results

The results of two experimental tests of the equations given above are shown in Figs. 3 and 4. Fig. 3(a) shows $1/I_2$ against N_2 for a core loaded by a one ohm resistor on its secondary winding in the circuit of Fig. 2; the 'core' used was actually a pair of cores type FX1396/D1, 8mm o.d. According to equation (1):

$$1/I_2 = \frac{N_2}{(N_1I_1 - F_c)} \quad \ldots\ldots\ldots\ldots (11)$$

and Fig. 3(a) shows a good linear relation except for the point at $1/I_2 = 0{\cdot}5\text{A}^{-1}$, where the effective load is so small that the core turnover is limited by the internal mechanism, and not by the load. The slope of Fig. 3(a) is 0·21 against 0·18 calculated from:

$$N_1 = 25 \text{ turns } I_1 = 0{\cdot}3\text{A}, \ F_c = 2\text{A}.$$

If now one uses the more exact theory and uses equation (7) instead of equation (1) then, eliminating $d\phi/dt$ between equations (8) and (2) one finds for $1/I_2$ the equation:

$$1/I_2 = \frac{N_2\,[1 + (R/N_2^2)\,.\,l/\alpha S]}{(N_1I_1 - F_c)} \quad \ldots\ldots\ldots (12)$$

$$= \frac{aN_2}{(N_1I_1 - F_c)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (13)$$

where

$$a = 1 + (R/N_2^2)\,.\,l/\alpha S$$

From the switching data on D1 cores, $\alpha S/l$ is calculated to be 0·38Ω for a single FX1396, or about 0·77Ω for a pair.

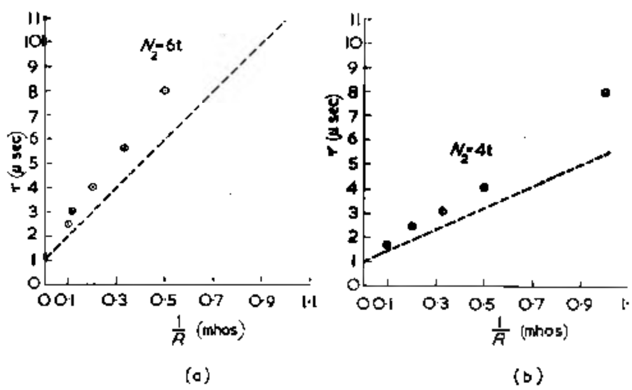

Fig. 4. Results of experimental tests

Using this value $1/I_2$ has been replotted in Fig. 3(b) against aN_2 and it can be seen that there is now much less discrepancy between the point at $1/I_2 = 0{\cdot}5\text{A}^{-1}$ and the line. Further the slope of Fig. 3(b) is 0·19 which is in rather better agreement with the calculated value.

The switching time given by equation (9) may also be used to test the theory. Fig. 4 shows τ plotted against $1/R$, for two cases, of $N_2 = 6$ turns and 4 turns. The linear relationship is clear. Using the values of $\phi_s = 76 \times 10^{-8}$Wb, $\alpha S/l = 0{\cdot}77\Omega$ and $F_c = 2$A (all for a pair of FX1396 cores) and $N_1I_1 = 7{\cdot}5$At, the calculated intercept on Figs. 4(a) and (b) is 0·36μsec. The 'best' straight lines, give an intercept about 1μsec; this cannot be considered surprising because the actual core output pulse in open-circuit has a long tail; the experimental times of turnover were measured between the points at 10 per cent peak value on the rising edge and the falling tail, and this will obviously be a considerably longer time than if all the flux change were concentrated into a rectangular pulse of constant amplitude. The calculated slopes (shown by the dotted lines which have been given an intercept of one microsecond) are lower than the experimental points by about 20 per cent for the 6-turn case and 40 per cent for the 4-turn case. Further work would be needed to resolve this discrepancy.

Applications

Switch Core Driving a Line of Storage Cores

Fig. 5 shows a circuit where the equations may be applied. It represents conditions in a part of the R.R.E. computer. Here a line of 15 storage cores (FX1508/D2) are set to the '1' state by a programme current pulse I_3. They are also connected to a winding on a switch core (which is in practice a pair of FX1396/D1 cores), and R represents the resistance of this part of the circuit. Later the storage cores are set back to '0' state by a 'read' pulse, I_1, on the drive winding of the switch core. Consider first the programme pulse; during this time the switch core is held biased hard over into saturation and any current I_2 will produce negligible flux change in the switch core, so that its inductance will be small and only the resistance need be considered. Also I_1 is zero at this time. The fifteen storage cores act as a transformer with I_3 as primary current and I_2 as secondary, the primary and secondary being single turn windings. Equation (9) for the switching time then becomes:

$$\tau = 2\phi_s\left[\frac{1}{(\alpha S/l)\,(I_3 - F_c)} + \frac{15}{R(I_3 - F_c)}\right]$$

The first term representing the switching time on open-

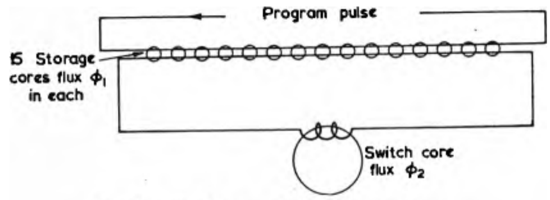

Fig. 5. Switch core driving storage cores

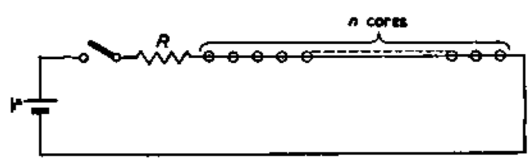

Fig. 6. Cores driven from imperfect current source

circuit may be taken as a fixed one microsecond. The second term, with $\phi_s = 4 \times 10^{-8}$W, $I_3 = 1$A and $F_c = 0{\cdot}4$A (for FX1508 cores) gives an $R\tau$ product of 2Ωμsec. With $R = 1\Omega$ this gives a 'circuit' switching time of 2μsec, and a total time when added to the fixed 1μsec of 3μsec. It has been found in practice that most cores do switch completely with these conditions, but some are not completely switched. Conditions are marginal. To give that factor of safety which is essential to ensure reliable working in the computer R has been set at 2Ω by careful choice of the wire gauge and wire length used; this gives a theoretical pulse length of 2μsec and in fact a 5μsec pulse has been used.

Having chosen the resistance R the design of the windings on the switch core can be considered. When the drive is applied the programme pulse wire is open-circuit and then for the switch core:

$$N_2\,(d\phi/dt) = I_2R + 15\,(d\phi/dt)$$

Integrating gives $2N_2\phi_{2s} = \int I_2R\,dt + 15{\cdot}2\,\phi_{1s}$.

For a 1A pulse, 2μsec long with $R = 2\Omega$ the $\int I_2R\,dt$ is 400×10^{-8}Wb, while $30\phi_{1s} = 120 \times 10^{-8}$Wb (with $\phi_{1s} = 4 \times 10^{-8}$). For a pair of FX1396 cores $\phi_{2s} = 76 \times 10^{-8}$, N_2 is therefore 3·5. Taking the next integral value gives $N_2 = 4$ turns. The required drive is found from equation (10); with $\alpha S/l = 0{\cdot}77\Omega$ for a pair of FX1396 cores and the mean $d\phi/dt = 152 \times 10^{-8}/2 \times 10^{-6} = 0{\cdot}76$, the turn added to the coercive m.m.f. is 1A. With $F_c = 2$A and I_2 required to be 1A it is found that $N_1I_1 = 7$A turns. The actual values used are $N_2 = 5$ turns and $N_1I_1 = 7{\cdot}5$At (compared with 8At calculated for $N_2 = 5$); these were the results of earlier experiments but the later developments of the theo-

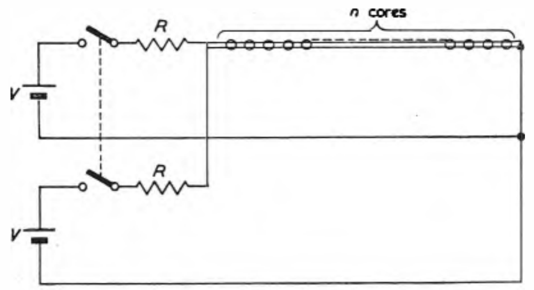

Fig. 7. More practical arrangement of Fig. 6.

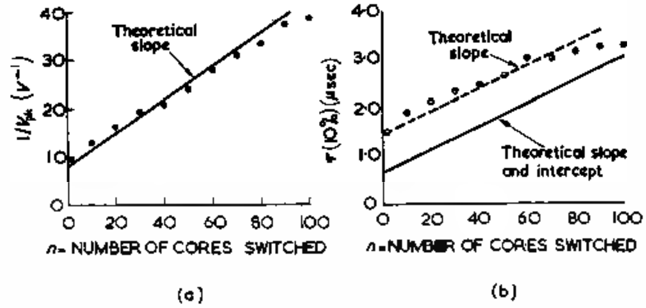

Fig. 8. Experimental results obtained with circuit of Fig. 7.

retical background indicate that the secondary winding could be reduced to four turns with a slight saving of input drive.

Driving Many Ferrite Cores from an Imperfect Current Source

Consider the circuit of Fig. 6 in which a set of n cores are arranged in series with a voltage source and a resistor. When the switch is first closed current $I = V/R$ flows. Then as the cores start to switch the back e.m.f. reduces the effective voltage acting in the circuit and reduces I. If this back e.m.f. is comparable with V the current will be significantly reduced leading to a reduction in switching speed and hence of output voltage.

The case was examined in the more practical form shown in Fig. 7 in which two generators each supply $I/2$; this approximates to conditions occurring in a coincident current matrix store. The equations governing the system of Fig. 7 are:

$$(dB/dt) = S(H - H_0)$$

$$(I/2) = (1/R)[V - n(d\phi/dt)]$$

Writing $H = (I/l)$; $H_0 = (I_0/l)$ and $\phi = \alpha B$

then eliminating I:

$$1/(d\phi/dt) = \frac{1}{(\alpha S/l)(2V/R) - I_0} + \frac{Z_n}{(2V - I_0R)} \quad (14)$$

$$= P + nQ$$

This shows that a linear relation should hold between the reciprocal of the output voltage and the number of cores switched. Further, the switching time should be:

$$\tau = 2\phi_s(P + nQ) \quad \ldots\ldots\ldots\ldots\ldots \quad (15)$$

showing a direct linear relation with n. The slope and intercepts of the straight lines, equations (14) and (15), should be related.

Fig. 8 shows the experimental curves obtained for the coincidence circuit of Fig. 7. High frequency transistors replace the switches and the voltage V is the collector supply voltage. The values used were $V = 6\text{V}$, $R = 15\Omega$, $I_0 = 0{\cdot}4\text{A}$ (for FX1508); these give a calculated slope of $\frac{1}{3}\text{V}^{-1}$ for $1/(d\phi/dt)$ against n (equation (14)) and a line of this slope on Fig. 8(a) fits the data well. There is also good agreement between the calculated slope of the τ against n line (equation (15)) on Fig. 8(b). The absolute value of the τ intercept on Fig. 8(b) calculated from the intercept of Fig. 8(a) is about a factor of two out, but this is certainly due to the over simplification made in ignoring the long tail of the actual pulse.

Conclusion

The behaviour of square loop ferrite cores under various conditions of loading and drive has been shown to be in agreement with that predicted from a simple theory.

Acknowledgment

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The Ophitron

The General Electric Company Limited have announced that a compact microwave generator embodying a new focusing principle has been developed at its Research Laboratories at Wembley. The programme of research which has led to this new type of valve was carried out by the G.E.C. for the Admiralty, under the programme for Co-ordination of Valve Research and Development.

The valve is an electrostatically-focused backward-wave oscillator which has been named the 'Ophitron', after the Greek: Oφis, a serpent, the word being suggested by the undulating path of the electron stream flowing along the structure.

The most striking advantages of the new oscillator are its small size (6in long × ¾in diameter) and low weight (7oz), and in addition the 'Ophitron' system has been designed to be simple to construct and to operate. A single stamped-out periodic structure and two flat focusing plates form the propagating path for the r.f. wave, and set up the periodic electrostatic-field which focuses the electron beam. The system has the fundamental advantage that the crests of the undulating electron beam are brought into the region of maximum r.f. field. This feature gives good coupling between beam and wave, and leads to greater bandwidth than is obtained with the equivalent magnetically focused backward-wave oscillator.

The present 'Ophitron' tunes electronically over at least a 40 per cent band in the 10 000Mc/s region. A range of such valves is envisaged, covering most important centimetre wavelength bands. It is expected that the noise performance will be better than that of magnetically focused backward-wave oscillators due to the ion drainage from the electron beam inherent in the focusing method.

This novel device is ruggedly constructed and when fully developed will have many applications in military and commercial fields. It will be manufactured and sold by the M.O. Valve Co. Ltd, Hammersmith, a wholly owned subsidiary of The General Electric Co. Ltd, and it is believed that it will be the first microwave oscillator incorporating electrostatic focusing to be commercially produced.

Transistor Invertors and Rectifier-Filter Units

By F. Butler, B.Sc., M.I.E.E., M.Brit.I.R.E.

Following an elementary account of the theory of transistor invertors, details are given of two practical designs suitable for operation from a 12-volt supply. One of these is a symmetrical push-pull circuit incorporating some refinements to improve the output waveform. The other employs four transistors in a bridge connexion which is adaptable to give a very high output power.

(Voir page 441 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 446)

A TRANSISTOR d.c. to a.c. invertor consists essentially of a transformer with some form of electronic switching mechanism which causes a recurrent interruption or reversal of current in the transformer primary winding. The desired a.c. output is taken from a suitable secondary winding. A conventional rectifier-filter unit may be added if the required output is smoothed d.c. In principle any standard oscillator can be used to effect d.c.-a.c. conversion but the normal sinusoidal operation results in low output and efficiency. For reasons which will be discussed later, square wave operation leads to an improved performance and all practical invertors use this mode.

Two basic types of circuit have been developed. In one of these the battery is connected to the invertor transformer through a single transistor which is used as an electronic switch. The supply current is periodically interrupted. The second type employs two or any multiple of two transistors in a symmetrical arrangement which allows the battery current to be reversed in the transformer primary winding. It is effectively a square wave push-pull oscillator and is the type normally used when a high output power is required. Square wave push-pull invertors are further divisible into types which are distinguishable by the nature of the switching mechanism used to effect the required current reversals. In one of these types the switching frequency is governed by magnetic saturation of the core of the invertor transformer. To a first approximation the operating frequency is independent of the connected load. In the second type the instant of current commutation is determined by the characteristics of the transistors used in the circuit. In this case the operating frequency varies with a change in loading.

In what follows the intention is to give a simplified theory of these modes of operation and to describe some circuit refinements which result in an improved switching waveform and which give good starting characteristics.

Finally a bridge-connected transistor circuit will be described. A low power version serves to illustrate the principle involved but the circuit appears easily adaptable to give a very high output power.

Invertor Switching Principles

Fig. 1 gives the circuit diagram of a transformer coupled relaxation oscillator with a load resistance R_L connected across the transformer primary. On closing the switch S the full battery voltage E appears across R_L and a linear rise of flux occurs in the transformer core. This in turn induces a constant voltage in the transformer secondary winding connected between base and emitter of the transistor. If the winding polarity is such as to make the base negative with respect to the emitter, the transistor draws an increasing collector current I_c which is the sum of the load current I_L and the transformer primary magnetizing current I_m.

The current I_m continues to rise until the transformer core becomes saturated or until the transistor collector current is limited by the available base circuit drive voltage which in turn defines the maximum base current. Collector

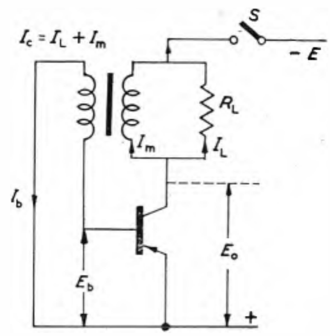

Fig. 1. Basic transformer switching circuit

current limiting due to this second effect is aided by the fact that power transistor current gain falls off seriously at large collector currents.

Fig. 2 shows waveforms at various points on the circuit of Fig. 1. A negative step in the base voltage causes the transistor to conduct. Its collector current I_c consists of a steady component I_L plus a linearly rising magnetizing current I_m which serves to sustain the constant base voltage. When I_m reaches a value sufficient to cause core saturation in the transformer the induced secondary voltage (the base drive voltage) cannot be sustained. The falling negative voltage now tends to reduce the base current and hence the collector current of the transistor. The consequent reduction in I_m results in the production of a reverse base voltage and a cumulative action ensues as a result of which the transistor is rapidly cut off and remains in this state until the energy stored in the transformer has been dissipated or usefully transferred to the load. Regenerative switch-on then takes place and the cycle is repeated. The collector output voltage has a rectangular waveform with limits E and E_0. While the transistor is conducting its collector voltage is bottomed at E_0 (normally a small fraction of 1 volt). While cut-off, the collector voltage rises to E (the supply battery voltage). When current commutation is initiated by transformer core saturation at some definite value of I_m the instant of switching is unaffected by the load current I_L. This feature of the circuit is illustrated in Fig. 2(b) and 2(c). In both cases the transistor remains switched on for the same time T and although the load currents are widely different the peak magnetizing current I_m is the same in each case.

Consider now the case in which the maximum collector current I_c is insufficient to saturate the core of the transformer. As before, after switching on, the total collector current has a fixed component I_L due to the load and a linearly rising component of magnetizing current I_m. The time rate of change of I_m is fixed by the supply voltage and the transformer constants but is independent of the load. The amplitude of the base drive voltage is fixed by the supply voltage and the turns ratio of the transformer.

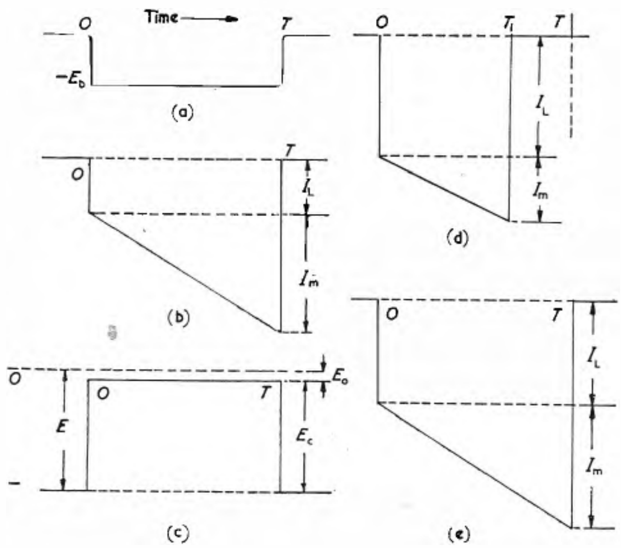

Fig. 2. Circuit waveforms of Fig. 1

(a) Base voltage
(b) Collector current
(c) Output voltage
(d) Collector current with transistor saturation
(e) Collector current with transformer core saturation

The transistor base current is in turn determined by the base voltage and the input impedance of the transistor while the maximum possible collector current which can be drawn from the supply battery is the base current multiplied by the current gain factor of the transistor. Referring again to Fig. 1 the initial action after switching on is as already described for the saturable core transformer but switch-off occurs not as a result of transformer saturation but due to the inability of the transistor to draw an increasingly large collector current when the base driving power is limited to some prescribed maximum value.

The total collector current I_c is the sum of I_L and I_m and if I_c is fixed it is clear that an increase in load current I_L must be compensated by a reduction in I_m. But I_m is a linear function of time and a reduction of I_m results in a proportionate reduction of the conducting phase of the transistor operation. The switching frequency is thus dependent on the connected load, whereas with transformer saturation the frequency is almost independent of the load. Fig. 2(d) shows a reduction in switching period from T to T_1 as a result of current saturation effects in the transistor. Switch-off occurs at maximum I_c since at this point there can be no further rise in I_m to sustain the requisite base drive voltage.

Both modes of operation have their special merits and both have certain limitations. The choice of a practical circuit must be made after considering user requirements. The arrangement using saturable core transformer switching operates at a frequency which is almost independent of the load. The high operating flux density calls for the use of low-loss core materials and a good switching waveform is best achieved by the use of alloys having rectangular B-H loop characteristics. The base circuit drive power requirement is rather high, resulting in relatively low efficiency on light loads. The arrangement is self-protecting against heavy overloads, starts readily with normal loads and its performance is not markedly dependent on transistor characteristics.

Invertors using unsaturated transformers operate at a switching frequency which is a linear function of the output load power. For a fixed load this is no problem and there are many applications in which a change of frequency can be tolerated. Normal core materials can be used, transistor drive power is not excessive and the overall

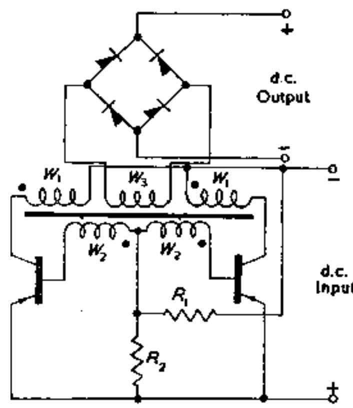

Fig. 3. Circuit diagram of simple invertor and rectifier

efficiency is high and is maintained high over a relatively wide variation of load resistance. Against this, the circuit requires protective measures in case of heavy overload. There is a critical load which will cause the frequency to rise to a very high value. In this condition, transformer losses become excessive and the collector current rises to a value which may destroy the transistors. The switching frequency is dependent on the transistor characteristics and if two dissimilar types are used in a push-pull circuit the output waveform is asymmetrical. Replacement of transistors by others with different current gain factors will result in a change of operating frequency. In both modes of operation the frequency is sensitive to supply voltage changes. On balance, experience shows that the saturable core switching mechanism is more desirable and the practical invertors to be described all use this mode of operation.

The description so far given ignores many theoretical and practical considerations which must be taken into account during development work on invertor circuits but these points are more conveniently discussed later in the text.

Invertor and Transformer Theory

A push-pull invertor has advantages over the single-transistor prototype circuit of Fig. 1 and the circuit diagram of a practical symmetrical invertor is shown in Fig. 3.

In this the transformer has a centre-tapped collector winding W_1, a centre-tapped base winding W_2 and an output load winding W_3, connected to a bridge rectifier if a d.c. output is required. The action after switching on the battery supply is that a current pulse is passed through W_1, inducing a voltage in W_2 and causing one of the transistors to conduct collector current. The starting action can be improved by the deliberate addition of circuit elements which disturb the initial symmetry of the arrangement. The rising collector current in one transistor induces a constant base circuit voltage and if the relative winding polarities are favourable a regenerative action sets in which causes a continued rise of collector current until the transformer core becomes magnetically saturated. Up to this point there has been a constant time rate of change of flux in the core but after saturation there is no further increase so that the induced base circuit voltage falls to zero. There is a corresponding fall of collector current and in core flux which drops to the remanent value. The changing flux induces a reverse voltage in the base circuit winding which rapidly cuts off the collector current. The second transistor, previously cut off by a positive base voltage, now begins to conduct. The whole action is repeated but with a reversal of the roles of the two transistors. It should be noted that, due to auto-transformer action, the peak voltage across the whole primary winding of the transformer is twice the battery voltage and that when one transistor is bottomed the collector of the other is subjected to the whole of this inverse voltage.

In Fig. 3 the resistors R_1, R_2 serve to provide the requisite fixed bias for the two transistors. The dots on the ends of the transformer windings indicate their instantaneous relative polarities.

The switching frequency may be calculated from elementary electromagnetic principles. It is well known that a changing magnetic flux in the core of a coil will induce a voltage in the coil which is proportional to the time rate of change of flux. Numerically:—

$$e = -(1/10^8) \cdot d\Phi/dt \quad \ldots\ldots\ldots\ldots (1)$$

where e = instantaneous induced voltage,

Φ = total flux linkage with the coil,

t = time in seconds.

The production of a constant e.m.f. requires only a constant rate of change of flux, i.e., a linear rise or fall with time. Conversely, if a battery of fixed voltage is suddenly connected to any coil with a magnetic core, the flux will change at a constant rate until saturation is reached.

For a coil of N turns wound on a magnetic core of cross section A in which the saturation flux density is B_o the maximum flux linkage Φ is given by:

$$\Phi = NAB_o \quad \ldots\ldots\ldots\ldots (2)$$

If a battery of voltage E is connected to the coil equation (1) becomes, (ignoring the negative sign):

$$E = (1/10^8)\, NA\, (dB/dt) \quad \ldots\ldots\ldots\ldots (3)$$

Where B is the instantaneous flux density. Integrating this and assuming that $B = 0$ when $t = 0$:

$$Et = (1/10^8)\, NAB,$$

or

$$t = (1/10^8)\, NAB/E \quad \ldots\ldots\ldots\ldots (4)$$

This equation gives the time in seconds required to reach any prescribed flux density in some particular coil and core combination. A numerical example is instructive.

Let N = 10 turns,

B = 18 000 lines/cm²,

A = 2cm²,

E = 12V,

$$\text{Then } t = \frac{10 \times 2 \times 18\,000}{10^8 \times 12},$$

$$= 0{\cdot}0003\text{sec},$$

$$= 300\mu\text{sec}.$$

If now the battery voltage is reversed when the saturation flux density is established in one direction in the core the first action will be to reduce the flux to zero and then to reverse its direction, eventually reaching saturation in this new direction. The total flux change is 2 × 18000 and the time to effect the flux reversal is clearly 600μsec. This time corresponds to one half cycle of the switching frequency of a practical invertor. With this amount of basic theory one can proceed to consider the practical design of an invertor transformer. Having chosen the battery supply voltage E, selected the transformer core material of cross-section A

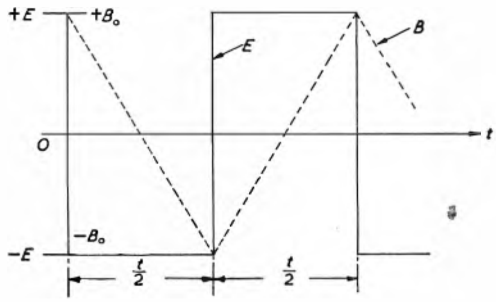

Fig. 4. Voltage and flux changes in invertor transformer

with a saturation flux density B_o and defined the required switching frequency f it is normally required to calculate the number of turns N on each half of the collector winding of the transformer.

Since the operating frequency is f the time duration of one cycle is $t = 1/f$. Starting with the condition that one transistor is drawing maximum current, the applied voltage across one half-primary winding is E and the flux density is the saturation value B_o. It is now required to switch over the transistors and, in the time of one half cycle (i.e. $t/2$), establish saturation flux density in the reverse direction. The rate of change of flux is thus $4B_o/t$, (see Fig. 4). Substituting $4B_o/t$ for dB/dt in equation (3):

$$E = (1/10^8)\, NA\, (4B_o/t)$$

But $1/t = f$, so that:

$$E = (1/10^8)\, 4\, NAB_o f,$$

or

$$N = \frac{10^8 E}{4\, AB_o f} \quad \ldots\ldots\ldots\ldots (5)$$

This is the fundamental design equation for invertor transformers. It may be written:

$$E = \frac{4\, NAB_o f}{10^8}$$

It is of interest to compare this with the well-known transformer equation:

$$E = \frac{4{\cdot}44\, NABf}{10^8}$$

Apart from the fact that in the second case B is the working flux density and not the saturation flux density B_0 the only difference is in the constant in the numerator, 4 in one case and 4·44 in the other. The ratio 1·11 is the form factor of a sine wave, defined as the ratio of the r.m.s. to the average value. In the square wave case the form factor is clearly unity.

From equation (5) it might appear that even when E, B_0 and f are fixed there is scope for an independent and arbitrary choice of N and A provided their product is held constant. In practice this is not the case. The core volume and hence its cross-sectional area A is obviously determined by the total output power required. The requisite turns ratio between collector and base circuit windings will be discussed later. Clearly it will depend on the transistor characteristics and in particular on the current gain factor in the common emitter connexion.

Practical Design Considerations

Square wave transistor oscillators are of high efficiency because the transistors alternate between two conditions in which the internal energy dissipation is very low. If one transistor is cut off there is no collector current and although the collector voltage is then twice the supply voltage there is no power loss. When this transistor is conducting maximum current the collector to emitter voltage is very low, commonly no more than a fraction of 1 volt. Here again energy dissipation is small. The greatest energy loss occurs during the transition from one state to the other. For high efficiency the transit time should be made as short as possible. Even a poor approximation to a square waveform shows a much higher overall efficiency than can be achieved by sinusoidal operation for which the maximum figure is probably less than 75 per cent. Unfortunately there are inherent defects in transistors and transformers which make it difficult to generate an ideal output waveform.

When a negative flat topped pulse having steep leading and trailing edges is applied to a power transistor in the common emitter connexion the output pulse is slightly delayed in time, shows a finite rise-time, persists for a longer time than the input pulse and has a sloping trailing edge. In addition the output pulse is not flat topped but commonly shows a step coincident in time with the trailing edge of the input pulse. A complete explanation of these effects leads one into the realm of semiconductor physics. The process of conduction in the base of a transistor is mainly by diffusion of holes from the emitter to collector and the hole transit time across the base is relatively long in high power low frequency transistors.

When a transistor is driven from cut off to normal conduction by a negative base voltage, base-to-emitter current flows in the emitter diode and initially the base current is equal to the emitter current. Emitted holes now diffuse across the base region and delayed collector current starts to flow.

As the collector current rises the base current falls until steady conditions are established under which the base, emitter and collector currents are related by the d.c. current gain factor of the transistor.

When the base is driven positive the emitter diode is cut off and the emitter current falls to zero. But holes are still diffusing across the base and collector current continues to flow, being now equal to the base current.

Still more marked effects are observed when a transistor is rapidly taken from cut-off to the fully bottomed condition in which it is drawing saturation current. When the base is driven negative the conditions are at first as previously described but the final value of base current is much greater than that required to maintain the collector current at its bottomed value. When the emitter diode is subsequently cut off by the application of a sudden large positive base voltage the emitter current ceases to flow and the excess holes in the base diffuse to the collector, thus maintaining collector current for an appreciable time at its bottomed value before it falls normally. If a large positive voltage is applied to a transistor previously carrying maximum current the emitter will act as a collector for holes remaining in the base region. A large transient reverse current flows in the emitter-base circuit before normal cut-off takes place. This reverse current pulse tends to cause ringing or shock excitation of the invertor trans-

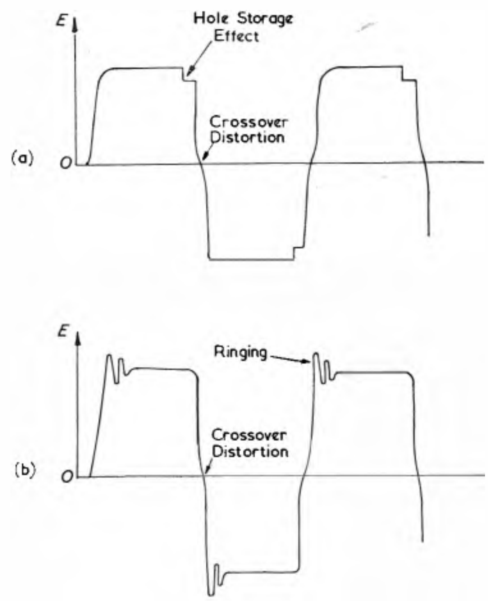

Fig. 5. Defects of invertor waveforms

former, the leakage inductance of which resonates with the winding, circuit and transistor capacitance and the effect manifests itself as a damped oscillatory voltage superimposed on the output waveform. In high power invertors the transistors are normally operated close to the manufacturers' limits of allowable peak current and inverse voltage. The additional oscillatory voltage may well result in destruction of the transistors. Two remedial measures may be applied. As regards the transformer, leakage inductance may be reduced by using close coupled windings. In cases where high capacitance can be tolerated, bifilar windings may be used. A toroidal core, although difficult to wind, has a lower leakage flux than stacked cores of E and I laminations and alloys of the nickel-iron type should be used in preference to conventional silicon steel material.

To minimize the effects of hole storage, experience shows that there is some advantage in restricting the amplitude of the base drive voltage but inadequate drive will introduce starting difficulties. A satisfactory compromise is to use an asymmetrical drive in which the positive half cycle is smaller in amplitude than the negative swing. Fig. 5 shows some of the waveform defects which may be observed in practice; (a) shows the output voltage from an ideal transformer while (b) illustrates ringing due to leakage inductance and shunt capacitance.

Invertors are commonly required to produce a high voltage output which calls for well insulated transformer windings. The switching transformer core need only be of very small size to perform the current commutation from one transistor to the other and the core window area is often so small that it is difficult to accommodate all the windings. One way out of this difficulty is to use an entirely separate output transformer, the primary of which is connected from collector to collector of the invertor in parallel with the switching transformer which then requires no secondary load winding. All the core winding space is thus available for coils which serve only for current commutation. Satisfactory operation can be achieved provided only that the magnetizing current of the switching transformer is appreciably larger than the full load current drawn by the auxiliary output transformer.

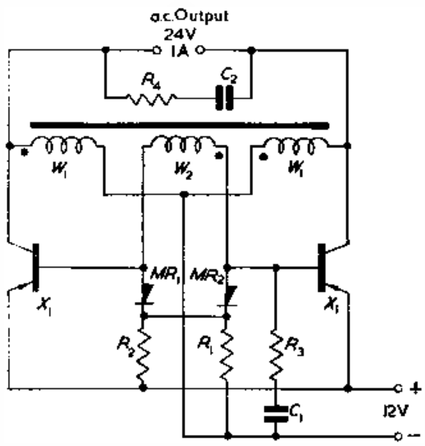

Fig. 6. Practical circuit of 12V 24W invertor

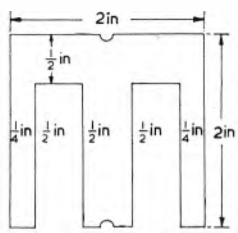

Fig. 7. HCR core lamination, size of switching transformer (stack depth-½in)

It may well be the case that the use of a separate output transformer adds little to the overall cost of the invertor. Switching transformers use expensive core materials and a reduction in core size can do much to offset the cost of a second core. The additional output transformer is operated under normal conditions and does not run into core saturation. Ordinary grades of magnetic materials, including ferrites, can be used and there is no problem in finding sufficient winding space for e.h.t. secondaries with thick insulation. Moreover, a single invertor can be used to supply a number of parallel-connected output transformers. Practical tests show very little difference between the overall efficiency of invertors with a single transformer and those with separate switching and output transformers. In the two circuits now to be described, only the switching transformers are included and it is assumed that supplementary output transformers will be employed. Standard design techniques may be used, though it may be advisable to operate such transformers at a conservative flux density.

Invertor Using Centre-Tapped Transformer

The circuit diagram of a 24-watt push-pull invertor is shown in Fig. 6. It incorporates a number of design features already discussed and differs from the standard circuit in respect of the base drive supply. In particular there is no centre-tap on the base circuit winding of the switching transformer and two germanium junction diodes have been included in the circuit. The resistors R_1 and R_2 provide sufficient base bias voltage to ensure reliable starting. The components C_1 and R_3 also improve the starting characteristic without affecting the output waveform.

The action of the circuit is most easily understood by assuming that R_2 is short-circuited. When the base of X_1

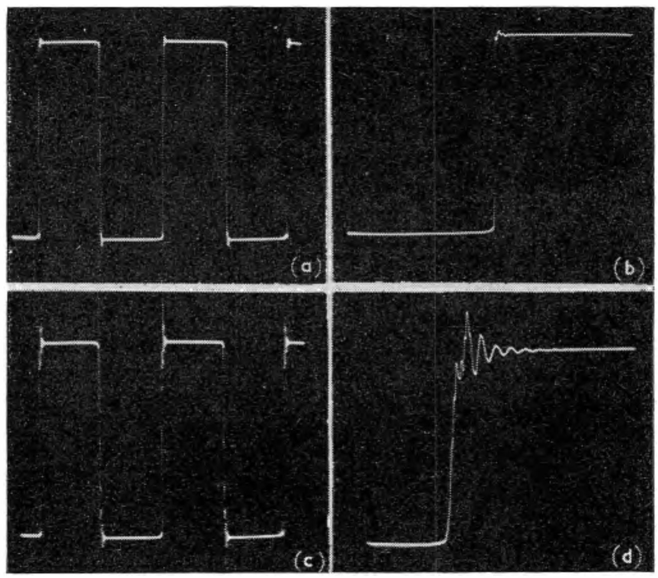

Fig. 8. Oscillograms

(a) Output waveform across a resistive load
(b) Switching transistion at a much higher time-base speed
(c) and (d) Waveforms across the secondary of a transformer with normal leakage reactance and with no compensation

is driven negative, this transistor conducts current and the diode MR_1 is inoperative. The diode MR_2 conducts a current which is equal to the base current of X_1, but the forward resistance of MR_2 is so low that the base of X_2 is virtually short-circuited to its emitter and cannot be driven positive. This limits the amplitude of the transient reverse current, due to hole storage effects, which mars the performance of the invertor shown in Fig. 3.

On the next half cycle when X_2 conducts current the functions of the two diodes are reversed. The action is practically the same when R_2 is in circuit except that this resistor is effectively in series with the conducting diode.

The components C_2 and R_4 provide additional suppression of transformer ringing by combining the load resistance and transformer leakage inductance to form a non-reactive network. A circuit of inductance L and resistance R can be made purely resistive at all frequencies by connecting in parallel with it a capacitance C in series with a resistance R, provided only that $L = CR^2$. For example if $L = 10\mu$H and $R = 10\Omega$, then $C = 0{\cdot}1\mu$F.

The switching transformer employs a core of HCR magnetic alloy. The core consists of 100 laminations of thickness 0·004in assembled to form a ½in stack having the dimensions shown in Fig. 7. The window area available for winding is 1in by ½in. The collector windings W_1 comprise 30 + 30 turns of 18 s.w.g. wire, bifilar wound, with the two sections connected series aiding. The base winding W_2 has 14 turns. This number of turns appears to be excessive and tends to overdrive the oscillator. It would be worthwhile to use a multi-tapped winding and to select the optimum number of turns by experiment so as to give the best possible output waveform while preserving reliable starting characteristics.

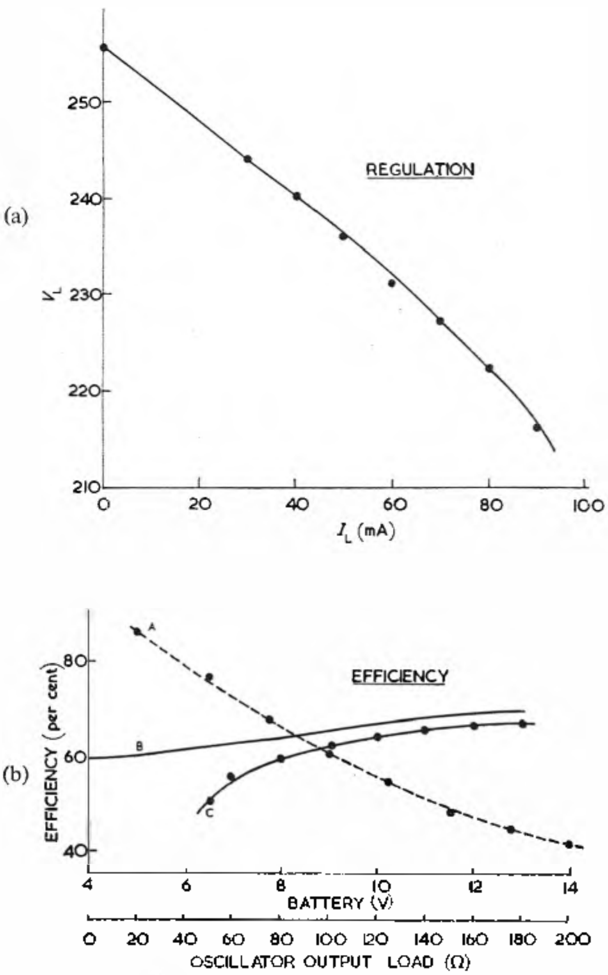

Fig. 9. Measured performance figures of invertor with rectifier filter unit
(a) Regulation
(b) Efficiency
Curve A *Oscillator section only (efficiency/R_L)*
Curve B *Efficiency/battery voltage (fixed load = 3kΩ)*
Curve C *Overall (efficiency/I_L)*

Remaining circuit details are tabulated below:

Transistors X_1 X_2	Newmarket, type V30/30P,	
Diodes MR_1, MR_2 . . .	B.T.H. germanium junction rectifiers, type GJ3-M,	
R_1 . . . . . 470Ω 1W,	C_1 . . . . 0·05μF,	
R_2 . . . . . 10Ω 1W,	C_2 . . . . 0·1μF,	
R_3 . . . . . 470Ω ½W,		
R_4 . . . . . 20Ω 1W,		

Operating frequency: 500c/s.

The output waveform across a resistive load is shown in Fig. 8(a). The oscillogram in Fig. 8(b) shows the switching transition at a much higher time-base speed. Current commutation takes place completely in 20μsec with negligible ringing or overshoot. Figs. 8(c) and 8(d) show the waveforms across the secondary of a transformer having a normal leakage reactance, without compensating resistance and capacitance, the primary being connected to the invertor output terminals. Pronounced ringing is visible.

Some measured performance figures for the invertor with an external h.t. rectifier-filter system giving a nominal output of 220V 80mA are shown in Fig. 9. The upper curve shows the output voltage regulation while the lower figure shows:—

Curve A

Efficiency of oscillator unit only when supplying a pure resistance load, variable between 20 and 200Ω.

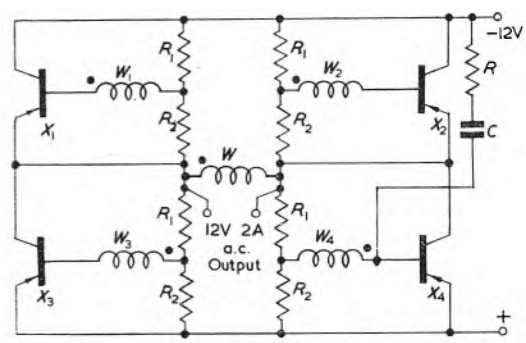

Fig. 10. Bridge connected transistor invertor

Curve B

Efficiency of invertor plotted against battery supply voltage when a fixed resistance of 3kΩ is connected across the h.t. output terminals of the rectifier-filter unit.

Curve C

Overall efficiency of the complete unit plotted against the d.c. output current.

The low efficiency of the invertor unit when lightly loaded shows that the transistor input drive power is excessive.

The same circuit can be used to give a higher output power by using parallel-connected transistors in each half of the push-pull circuit. If two matched pairs are used the only circuit change required is to halve the values of the base bias resistors. There is some risk of exceeding the peak current rating of the two germanium diodes but extra diodes may be connected in parallel with the existing pair to achieve the desired peak current capability.

Bridge-Connected Transistor Invertor

When a very high output power is required there are many advantages in using a bridge-connected arrangement of transistors. A suitable circuit is shown in Fig. 10. Four transistors are used in association with a transformer having a single primary winding and four separate secondaries each of which provides the base drive power for one transistor. The transformer winding polarities are such that the transistors conduct alternately in diagonally opposite pairs. Thus X_1 and X_4 will both conduct when their bases are driven negative and a current will flow through the a.c. load in one particular direction. On the next half cycle X_1 and X_4 are cut off while X_2 and X_3 conduct and load current flows in the reverse direction.

As before, dots indicate the relative polarities of the transformer windings. Base bias to facilitate starting is provided by the resistors R_1 and R_2. These should be so chosen that all the transistors draw equal currents when supplied from a battery of one half the voltage of that which will be used to operate the invertor, since in this circuit two transistors are always connected in series across the full supply voltage.

Fig. 11 gives a diagrammatic comparison of the switching actions in the push-pull and bridge-connected invertor circuits.

A practical invertor has been built to give 24W output from a 12V supply, using Newmarket transistors type V15/10P. Operation from a 24V supply would be feasible using V30/30P transistors, in which case the output would

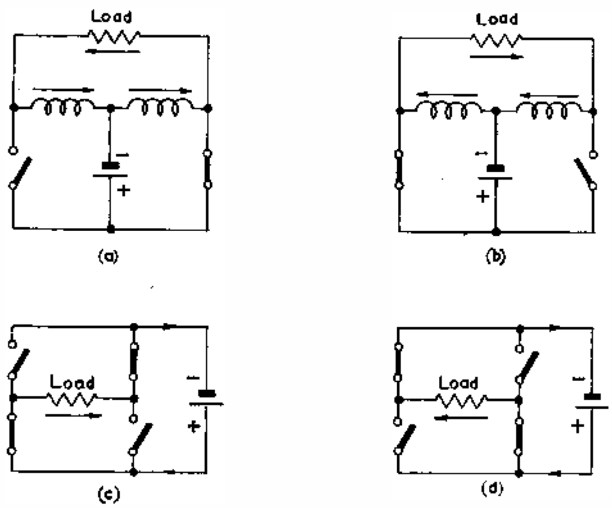

Fig. 11. Invertor switching sequences

be at least 50W. Still greater output could be obtained by parallel connexion of two or more transistors in each arm of the bridge.

At first sight the transformer design appears rather complicated, but although four separate secondary windings are required they have so few turns that the construction is very simple. The same type and size of core is used as for the transformer employed in the push-pull invertor already described. The primary winding has 32 turns of 18 s.w.g. enamel and cotton covered wire. Each secondary has 10 turns of 24 s.w.g. To reduce the leakage inductance, four separate lengths of wire are loosely twisted into cable form, 10 turns of this being wound on the core and the ends separately connected to the secondary terminals. In this way close magnetic coupling is achieved. This type of construction, together with the bridge connexion of the transistors which permits the use of an untapped primary winding, is sufficient to produce a very good switching waveform without further circuit complications. The components R and C serve to assist starting against full load.

Component values in the actual invertor are:

R_1 . . . . 330Ω, 1W,	R . . . . 560Ω, $\frac{1}{2}$W,
R_2 . . . . 6Ω, 1W,	C . . . . 0·05μF

As in the case of the push-pull invertor the bridge connected version in intended for use with a separate output transformer and bridge-connected rectifier filter system of conventional design. The invertor transformer is so simple in construction that a toroidal core could be used and the windings put on by hand. Core losses would thus be reduced to a minimum and the switching waveform would be almost ideal. Magnetostriction noise is a nuisance in transformers which are driven hard into core saturation. The toroidal construction leaves no exposed surfaces to radiate sound and more nearly silent operation would be achieved.

Oscillator-Amplifier Invertor Systems

The operating frequency of a saturable core invertor is directly proportional to the d.c. input voltage and is slightly dependent on the value of the connected load resistance. The load-dependent change of frequency can be removed by making use of a low power invertor followed by an amplifier using an unsaturated output transformer. Alternatively, a self-oscillating invertor can be synchronized by an external pulse generator driven by a low power precision oscillator. The oscillator-amplifier arrangement is ideally suited for supplying high frequency power to a magnetic amplifier. A low frequency version is suitable for operating standard magnetic tape recorders from batteries in special cases when an a.c. mains supply is not available.

When a very large output power is required there are again advantages in using an oscillator-amplifier combination. A simple and highly efficient circuit could consist of an oscillator using the circuit of Fig. 7 with four separate secondary windings on the output transformer, arranged to drive a four-transistor bridge circuit. No output transformer would be required in the amplifier if the a.c. load was arranged to present the correct impedance for direct connexion. Alternatively a very simple transformer could be designed to match any specified load impedance. Overall efficiency would be very high for a number of reasons. The driving oscillator need produce only a moderate output power; starting would be easy and it would be possible to use a low forward base bias voltage to reduce the normal quiescent collector current of the oscillator. Fewer turns would be required on the transformer base circuit windings. The reduced drive voltage would in turn cause less internal energy dissipation in the transistors. As regards the amplifier, this could be operated under strict class-B or even under class C-conditions. In this case, no bias resistance network would be needed and another source of energy loss would disappear. Finally, the output transformer could be operated at a flux density well below the saturation level and core losses would be correspondingly reduced.

In conclusion it is worth remarking that the four-transistor bridge circuit could be used as the output stage of a high power audio frequency amplifier. With a 28V power supply, four V30/30P transistors would provide more than 25W output with acceptably low distortion.

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Some Unusual Devices for Electronic Music

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

The conventional electronic keyboard instruments are now widely known, but instruments for producing or modifying musical tones by other means continue to occupy the attention of investigators. In this article a number of unusual devices of this nature are described.

(Voir page 441 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 446)

THE conventional keyboard instruments are now widely known, but instruments for producing or modifying musical tones by other means continue to occupy the attention of investigators.

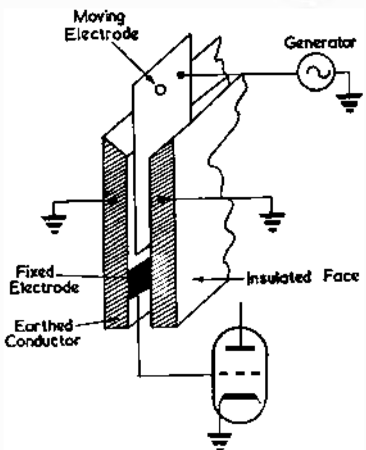


Fig. 1. Touch sensitive key coupling

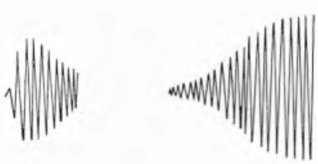


Fig. 2. Different types of envelope control from arrangement of Fig. 1.

The natural limitations of an equally tempered scale, the inability of keyboards to produce gliding tones, and the known difficulties of regulating the attack and decay of sounds as well as the impossibility of mixing with infinite gradation form the subject of several less well known electronic instruments. None of the apparatus for ' Musique Concrète ' is included under this heading for the instigators of this art insist that the sounds shall be formed by distorting concrete sounds, e.g., rain, footsteps, railway locomotives, falling drops of water and a similar assortment of apparently non-musical sounds. Nevertheless, ' music ' can be actually scored for this art, and since a good deal of electronic apparatus is involved, six references are appended[1].

One of the main defects of keyboard sustained tone instruments is that they are not touch sensitive. To a limited extent, the Baldwin organs are expressive in this way, as are old tracker action pipe organs. A method of controlling the attack of an oscillator is shown[2] in Fig. 1. By applying several of these variable electrostatic couplings to each key, and feeding them with related harmonics to form a composite tone, many interesting effects are possible. A very high attenuation can be attained[3], as can

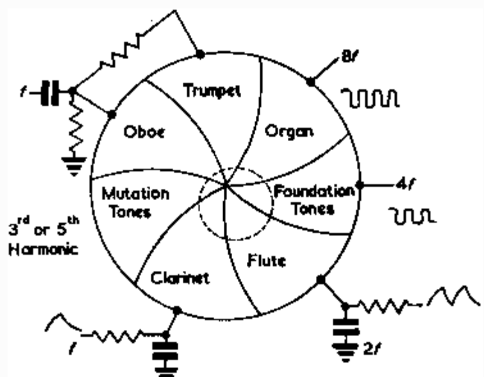


Fig. 3. Tone mixing device

be seen in Fig. 2, which shows two of the different attack envelopes available. While one of the many gating or storage circuits can give the same kind of attack control, it cannot produce graduations of attack at the will of the performer after the manner of a piano. The system can be applied to polyphonic instruments.

Turning to timbre control, such as exists in all orchestral instruments and is, of course, responsible for their versatility, an ingenious method of mixing several different waveforms[4], is shown in Fig. 3. The circular plate is divided into several separate conducting segments which are supplied with waveforms from a frequency divider chain. These waves, square in origin, are modified by external networks to increase the complexity of the harmonic structure.

The above photograph shows a multivibrator being played at the Philips electronic music studios

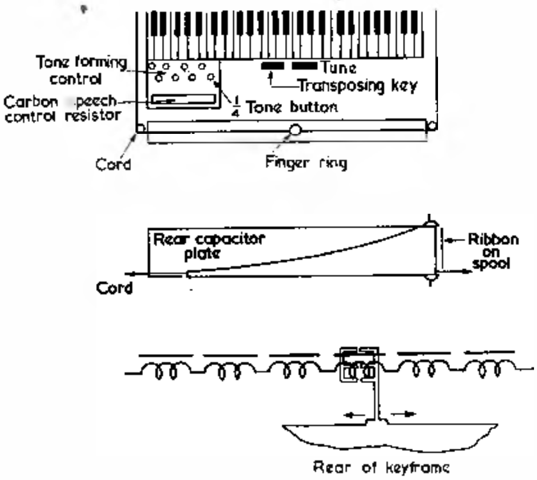


Fig. 4. Controls of Ondes Martenot

A conductive disk, shown dotted, can be moved over the surface of the segments, and is capacitively coupled thereto. The mixing of the signals can evidently be carried out exactly as the player requires, but it should be remembered that this is only part of the effect and the attack and decay circuits must be adjusted to suit the kind of sound simulated. Since one hand is required to operate the mixer, the device is more suited to a monophonic generator.

An attempt to perform the same operations with a keyboard instrument is exemplified in the Bode Melochord[5]. The generators in this are multivibrators, but the point under discussion is the 'travelling formant' circuit. This tone filter is continuously re-tuned by the playing keys, in step with the pitch of the generators. In this way, some correspondence is achieved with the same effect which obtains in orchestral instruments. To do this, contacts are arranged on the keys in such a way that all of the formant tuning capacitors which are on the left-hand side of any key depressed are separated from the main tuning capacitors; thus the sum of all capacitors (for they are in parallel) remaining to the right of the key played is available for tuning the parallel *LC* resonant circuits. There are 37 steps in the capacitors, corresponding to the 37 keys of the instrument; but the inductors are also tapped, so a very wide range of tonal effects is possible.

One of the most ingenious melodic instruments is the Ondes Martenot[6]. A four octave keyboard has, as an alternative frequency control, a continuously variable capacitor. Fig. 4(a) shows the general arrangement of the playing desk. The circuit uses a beat frequency oscillator, and a set of iron cored inductances. They are not used together. By means of a tape with a ring attached for operation by the finger, a logarithmic ribbon capacitor changes the oscillator frequency with a linear movement of the tape, corresponding very nearly to the position of the keys above. But, of course, there are no steps so that gliding tones are readily produced. On the other hand, when the keys are used, the inductances are progressively cut out of circuit (as in the Hammond Solovox[7]) and the circuit is re-tuned in this way. Every fourth or fifth coil has a movable iron core projecting and attached to the keyboard frame, which rests on rollers. Therefore by slightly pressing the keys from left to right, the frequency is modulated to produce a vibrato. This is an extremely expressive method, affording the performer the same facilities as he would have if playing a 'cello. A light spring ensures that the instrument will return to correct pitch.

The attack or speech control consists of a bar operating a noiseless carbon resistor. This is controlled by the left hand. A small button nearby changes the basic pitch of the oscillator so that, with sufficient adroitness, it is possible to play in quarter tones on the keyboard. There is also a transposing control, to allow of playing in any key in the usual way.

Possibly the most remarkable part of this apparatus is the loudspeaker. Fig. 5 shows a wooden sound chamber, similar to a mandolin body, strung on both sides with twelve steel wires. These are tunable, and pass over bridges near the top, where they are rigidly anchored. At the bottom, they are all collected at a special type of moving-coil loudspeaker movement. The action of this is lateral, as shown by the arrows, and at resonance the strings vibrate forcibly and a loud sound is produced. Since harmonics also energize the strings, a continuous and sonorous sound spectrum is heard. The starting and stopping characteristics of the tone are like those of the harp. In addition to this device, an ordinary small cone is provided and there is a most original 'tweeter' in the form of a large, very hard brass cymbal to which a small moving-coil element is rigidly fixed. For monophonic music, the Martenot opens up endless possibilities, limited largely only by the skill of the player.

There have, of course, been earlier designs for continuous tone generators, notably the Theremin[8], but these have proved difficult to control or unstable in operation. The most versatile instrument for serious music is the Trautonium, widely used in Germany[9].

Fig. 5. Martenot loudspeaker

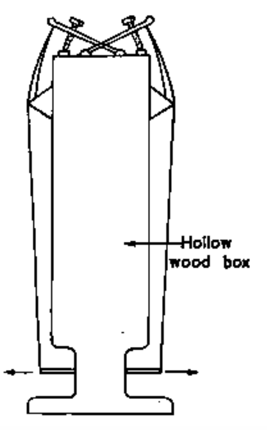
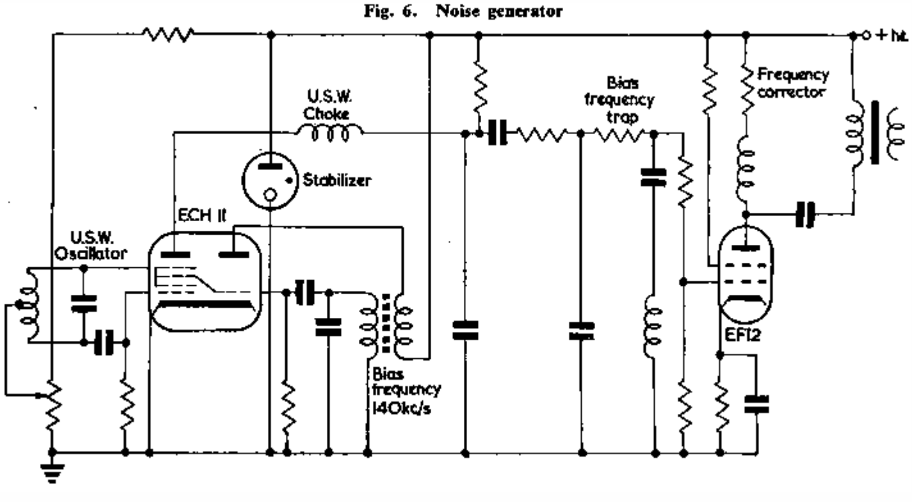


Fig. 6. Noise generator



Tape recording enters into the presentation (for performance) of most music these days. One ingenious application employing a tape permits of rhythms of any speed or complexity being imposed on music. If the output from a recorded tape is applied to one pair of terminals of a suitable ring modulator, and a series of high frequency pulses are applied to the other terminals, then the modulator can be blocked against the musical signal whenever the tape is free from impulses. Thus a regular rhythmic pattern can be imparted to steady tones. It must not, of course, be possible to reproduce the pulse frequency.

It is surprising—at least to the author—that noise generators have not received greater attention in this country. Both in Germany and the U.S.A. there has been quite widespread use of these adjuncts to fidelity. When one compares the sound of an electronic flute or an organ diapason with its physical counterpart, the immediate reaction is that there is no wind noise. This exceedingly vital ingredient of such sounds is only to be found in the Gregorian organ in this country. Many circuits can be used to generate white noise, from gas tubes to hard valves. One successful arrangement is shown in Fig. 6.

In the present search for sound sources which can be used for the composition of 'electronic music', stress is naturally laid on those properties which are either not possible, or very difficult to attain in conventional instruments. These are, infinite control of attack; continuous pitch spectrum; and musical percussions. An instrument which can be played with a continuous pitch spectrum is a multivibrator. As shown[10] on page 419, a large pulley is attached to the control spindle of an unsymmetrical multivibrator. If the potentiometer is logarithmic, then the cord can be stretched over a scale so that linear movement produces equally spaced semitone intervals and any subdivision thereof. Instruments of this kind are also useful for research into musical fundamentals.

More recently, the acoustics research department of the Bell Telephone Laboratories has been investigating the use of digital computers as a means for creating both new and conventional music. Since sound waves can be produced by means of a digital-to-analogue convertor operating on a sequence of numbers generated by a general purpose digital computer, it is possible to programme the computer to produce any sound. Conventional computers have several disadvantages, notably the difficulty in processing enough data for music of any length; and the irregular spacing of information on the tape. There are solutions to these and other problems, and if they are overcome, then digital computers could be invaluable for musical research.

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A Desk-Sized Computer

A new transistorized desk-size electronic digital computer, called SIRIUS, designed to fill the gap in the small general purpose computer field in Europe, is being manufactured by Ferranti Ltd.

It will cost £15 000 and the first production models will be available this autumn and the manufacturing schedule will enable delivery to be made within three months of ordering.

SIRIUS is believed to be the smallest and most economically priced computer yet made in Europe. It weighs only 5cwt, measures approximately 7ft × 3ft 6in × 4ft high, including a standard office desk, and has a power consumption of only 3·8A at 250V.

The new computer is designed to provide a small and relatively inexpensive machine for those sections of industry, commerce and research that need a computer but do not have the volume of work to keep a large machine economically occupied. Its chief application is expected to be statistical analysis in industry, commerce and laboratory where the rapid turn-around of statistical calculations can appreciably speed-up the continuity and flow of production work, but it can also perform many other types of jobs.

It is also intended for technical colleges for general educational purposes and training of electronic engineers, technicians, and physicists. Moreover, large commercial or government organizations that are geographically scattered may require many small machines rather than large central computing centres. Other suitable applications are for consulting engineers, partial process control of industrial plant, general statistical analysis work in new fields such as medical and forestry centres and zoological laboratories.

It may also provide a useful ancillary machine for a large computing installation so that additional experimentation outside the major stream of the organization's activities can be computed without interrupting the main machine, or compute statistics for the large machine which otherwise it may have to do itself. This is an important consideration where large installations are at present forced to work on secondary tasks before they can be used for their main functions.

The computer is constructed of neuron packages, magneto striction delay lines and is fully transistorized. It works in decimals, each decimal digit being represented by four binary digits. Checking is done inside the machine by automatic detection of forbidden combinations. Input and output is on paper tape and a set of 100 buttons, similar to a desk calculator, may be used to operate the machine or put information directly into the main store. Very few buttons need be used for developed programmes. The basic input/output speeds are the same as for the company's larger all-purpose Pegasus.

For many statistical problems, the time taken by any computer is dictated by the time taken on output of results, and where the particular application could be done on the smaller machine, it would be appreciably cheaper to do do.

The Company's London Computer Centre has installed the first SIRIUS and it will be offered for hire at about £15 an hour.

Design Factors for Industrial Cold-Cathode Timers

By J. F. Young*, A.M.I.E.E., A.M.Brit.I.R.E.

The accuracy of timers is determined by the stability of the components used and by the variations of circuit voltages. An optimum design exists if a pre-ignition current is required by the cold-cathode trigger tube. A simpler timer can only be used to control heating loads accurately over a limited range of supply voltage, but by adding a step function to the timing wave, the range can be greatly increased.

(Voir page 441 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 446)

ELECTRONIC time delay relays are now widely used in industrial equipment. Timers using valves having heated cathodes have the disadvantage that they are not available for use instantly when the supply is switched on, since time must be allowed for the heater to warm up the cathode. Recent developments in the construction of reliable cold-cathode trigger tubes having accurately determined triggering characteristics have made such devices attractive for use in timers, in particular when the above mentioned heating delay of the hot cathode valve is inconvenient. The main source of inaccuracy in a timer using a modern cold-cathode trigger tube is supply voltage variations. When the delay obtained must be maintained as constant as possible regardless of supply voltage variations it is necessary to use cold-cathode reference tubes to stabilize the supply. In some cases, however, the variation of delay with supply voltage can be made use of to obtain desirable characteristics.

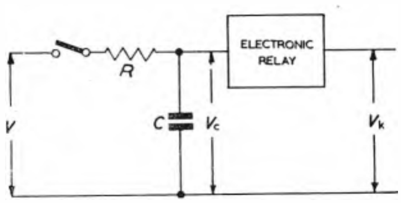

Fig. 1. Basic timer arrangement

The basic circuit of a typical time delay relay is shown in Fig. 1. This uses the exponential form of the voltage across a capacitor when charged through a resistor from a constant voltage supply. When the voltage across the capacitor reaches a certain value, the electronic circuits operate a relay which terminates the timing interval. The requirement for operation of the relay is, therefore:

$$V_c = V(1 - e^{-(t/CR)}) = V_k \quad \ldots\ldots\ldots\ldots (1)$$

where the voltages are as indicated on Fig. 1, and t is the time from the instant of application to the CR circuit of the voltage V. From the above expression the value of t can be found, since it follows that:

$$e^{-(t/CR)} = \frac{V - V_k}{V} \quad \ldots\ldots\ldots\ldots (2)$$

and hence, at the instant of operation of the relay,

$$t = CR \ln \frac{V}{V - V_k} \quad \ldots\ldots\ldots\ldots (3)$$

In order to prevent variation of the operating time, all quantities on the right-hand side should be maintained at constant values.

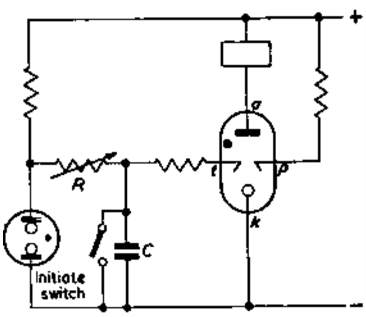

Fig. 2. Constant time relay with stabilizer

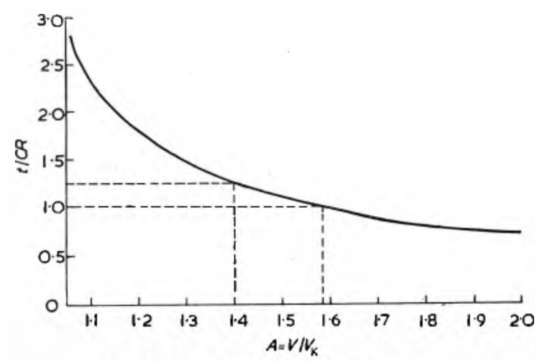

Fig. 3. Variation of delay with voltage

If a cold-cathode trigger tube is used in the timer, then V_k is the trigger electrode to cathode firing voltage. This is very stable in modern tubes. The use of good quality capacitors and resistors will limit variations of the CR value. The only variable quantity on the right-hand side of equation (3) is, therefore, the supply voltage V. This voltage is often derived from a cold-cathode stabilizer tube as shown in Fig. 2. Very good accuracies are obtainable for medium times of operation with this arrangement. The

* *The General Electric Co. Ltd., Witton.*

limiting factor for short operating times is the variable speed of closure of the electro-mechanical relay. For very long operating times the voltage across the capacitor is changing very slowly at the instant of firing and the small uncertainties of the voltages V and V_k become important. The rate of change of t with variations of V_k is $CR/(V-V_k)$ so that it is directly proportional to the time required, if V and V_k are nominally fixed. Similarly the rate of change of t with variations of V is:

$$-\frac{CRV_k}{V(V-V_k)}$$

which also increases in value as the time required increases. The relationship between t/CR and V/V_k is shown in Fig. 3. From this curve the effect of variations of either voltage on the timed interval can be determined if it is

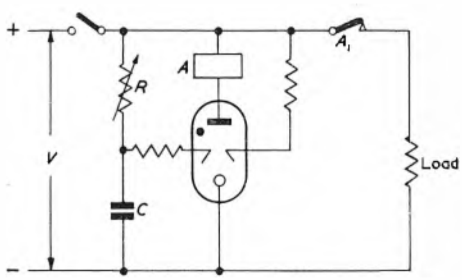

Fig. 4. Relay controlling heating load

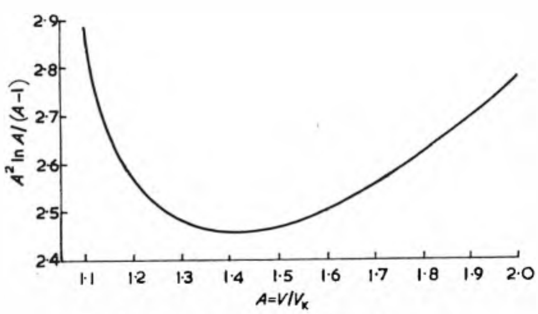

Fig. 5. Variation of heat energy with voltage

assumed that CR is unvarying. L. H. Light[1] has discussed the percentage errors of timing resulting from variations of V and of V_k in timers of this type. He derives the ratio of percentage change of time to percentage change of either V or V_k as:

$$\frac{V_k}{(V-V_k)\ln\dfrac{V}{V-V_k}}$$

and gives a curve of this factor against V/V_k. This curve shows a steep rise at values of V_k greater than about 0·8V, corresponding to operating times greater than about 1·5RC.

The trigger electrode of a cold-cathode tube requires a certain pre-ignition current in order to strike the discharge. The charging resistor must therefore be carrying a greater current than the required pre-ignition value at the instant of striking. Now at this instant the voltage across the series resistor R in Fig. 1 is $V - V_k$. The absolute maximum permissible value of R is therefore:

$$R = \frac{V-V_k}{I}$$

where I is the required pre-ignition current. Substituting this value of R into equation (3):

$$t = \frac{C(V-V_k)}{I}\ln\frac{V}{V-V_k}$$

$$= -\frac{CVX}{I}\ln X$$

where $X = 1 - (V_k/V)$

hence $dt/dX = -(CV/I)(1 + \ln X)$

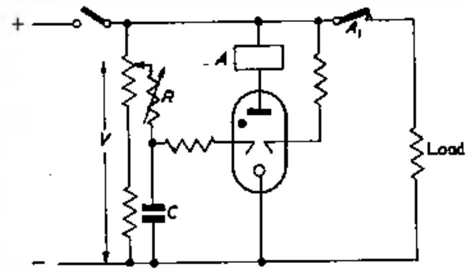

Fig. 6. Adjustment for capacitor tolerance

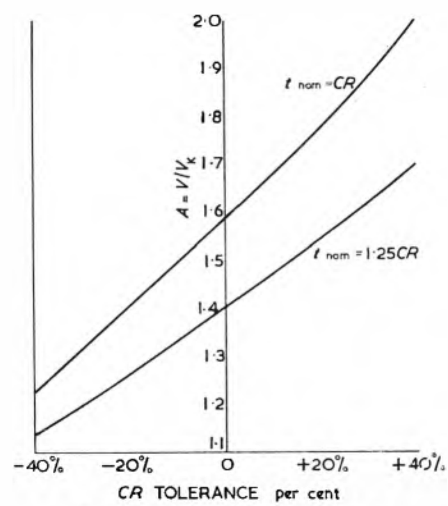

Fig. 7. Effect of tolerance adjustment

The maximum time of operation for a given capacitor size, supply voltage and required pre-ignition current is therefor obtained when:

$$\ln X = -1$$

$$X = 0{\cdot}368$$

$$(V_k/V) = 0{\cdot}632$$

The corresponding maximum time is

$$t = \frac{CVX}{I}$$

$$= \frac{0{\cdot}368\,VC}{I}$$

$$= CR$$

It is usual to design such timers so that the delay equals the CR value because this simplifies the selection of values of capacitor and resistor which will give the required delay. The above analysis shows that where a pre-ignition current is required, a design making the delay equal to the CR product has the additional advantage that it gives the longest possible delay for a given supply voltage and capacitor value.

The additional expense of the cold-cathode stabilizer tube shown in Fig. 2 is often not justified. This is particularly so when the timing device is used to control heating loads, such as welders or ovens. In such cases it can often be assumed that the load has a constant resistance R which is connected to a supply voltage V_s for a time of t seconds. The power in the load is then V_s^2/R and, if it can be assumed that the voltage V_s is constant over the interval t, the energy appearing as heat is V_s^2t/R joules. In order to maintain the heating effect constant from one timing cycle to another, it is necessary to maintain constant the quantity V_s^2t, rather than the time t. The heating effect will then be automatically compensated for supply voltage variations.

If the voltage V in Fig. 1 is derived from the load supply voltage V_s as shown in simplified form in Fig. 4, then from equation (3):

$$V^2t = CRV^2 \ln \frac{V}{V - V_k} \quad \ldots\ldots\ldots\ldots \quad (4)$$

If it can be assumed that V_k and CR are constant, then:

$$V^2t = KA^2 \ln \frac{A}{A - 1} \quad \ldots\ldots\ldots\ldots \quad (5)$$

where $A = (V/V_k)$ and $K = CRV_k^2$ is a constant. The value of $A^2 \ln A/(A - 1)$ is plotted against A in Fig. 5.

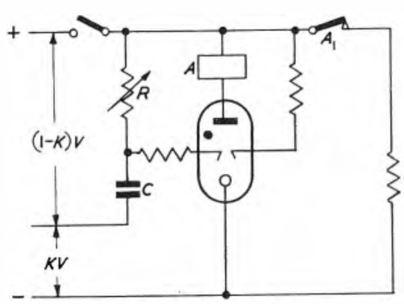

Fig. 8. Improved constant energy relay

From this it can be seen that, over a certain range of A, variations of V^2t with A are minimized. The minimum value of V^2t is obtained when $A = 1{\cdot}4$, and the time of operation is then $1{\cdot}25\ CR$. This is the best point at which to design the circuit if both increases and decreases of supply voltage are to be encountered. It was shown above that a design in which the delay $t = CR$, gives favourable operating conditions. This operating point corresponds to $A = 1{\cdot}58$. It can be seen from Fig. 5 that this is a satisfactory design for use where a decrease rather than an increase of supply voltage from the nominal value is normally encountered.

In a practical design, consideration must be given not only to the variation of supply voltage and of trigger voltage, but also to the tolerances on the values of resistors and capacitors used in the timing circuit. The effect of variations of the capacitor and of the resistor on the timed interval can be found from Fig. 3 for any given values of V and V_k. One common way of setting up a timer to give the correct nominal interval when wide tolerance components are used is shown in simplified form in Fig. 6. The voltage V is adjusted to give the correct nominal operating time. The effect of this on the value of A (i.e. of V/V_k) is shown in Fig. 7 for various tolerances of the CR product. The actual operating points on Fig. 3 or Fig. 5 can be found from this curve. It is seen from this that excessive component tolerances can make it necessary to have the operating point well away from the point at which variations of supply voltage have little effect on the V^2t product. If possible, therefore, close tolerance components should be used in the timing circuit.

If equation (1) is restated in a general form as:

$$V_c = V\,f(t) = V_k \quad \ldots\ldots\ldots\ldots \quad (6)$$

it can be seen that, in order to obtain a true constant energy characteristic $V^2t = K$, it is necessary to have:

$$f(t)\ (K/t)^{\frac{1}{2}} = V_k$$

or

$$f(t) = B(t)^{\frac{1}{2}} \quad \ldots\ldots\ldots\ldots \quad (7)$$

where B is a constant. With a simple CR circuit this characteristic is obtained over a small range as has been shown above. An approximation over a wider range can be

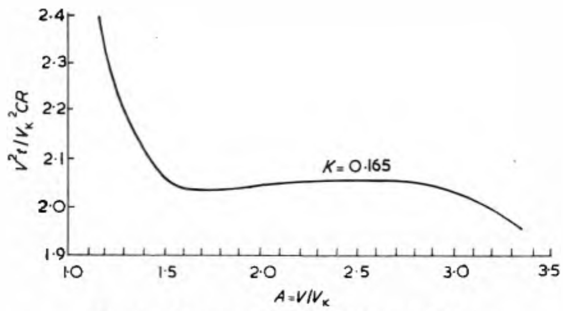

Fig. 9. Variation of heat energy with voltage

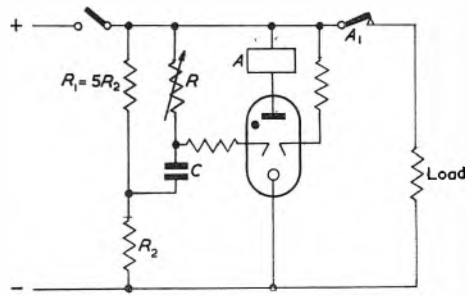

Fig. 10. Resistive divider in constant energy relay

obtained by the addition to the exponential charging wave of a steady voltage proportional to the supply voltage. If this is done the requirement for operation of the relay becomes:

$$V_c = V\,[K + (1 - K)\,(1 - e^{-(t/T)})] = V_k \quad \ldots\ldots \quad (8)$$

where K is as shown on Fig. 8. The time of operation is then:

$$t = CR \ln \frac{A\,(1 - K)}{A - 1} \quad \ldots\ldots\ldots\ldots \quad (9)$$

where $A = (V/V_k)$ as before.

Equation (9) can be written as:

$$t = CR \ln (1 - K) + CR \ln \frac{A}{A - 1} \quad \ldots\ldots \quad (10)$$

Hence:

$$V^2t \propto A^2 \ln (1 - K) + A^2 \ln \frac{A}{A - 1} \quad \ldots\ldots \quad (11)$$

The second term will be seen to be that plotted in Fig. 5, and to this the quantity $A^2 \ln (1 - K)$ is added. Since K is constant, positive and fractional the added quantity therefore compensates to some extent for the rising part of the characteristic of Fig. 5 as A increases. The best compensation is obtained when K is about 0·165, the resulting characteristic of V^2t shown in Fig. 9 being reasonably flat over a wide range of A. From this curve it can be seen that a two to one range of supply voltage can be tolerated with less than $\pm 1{\cdot}5$ per cent variation of energy V^2t.

A circuit giving this result is shown in Fig. 10. A similar arrangement has been described[2] for use with a timer intended to maintain constant V^5t. One object of the use of this circuit was to inject some ripple from the supply voltage in addition to the more smoothed exponential timing wave. This makes it possible to use a larger value of charging resistor, since the added ripple causes the capacitor to discharge into the trigger electrode and so supply

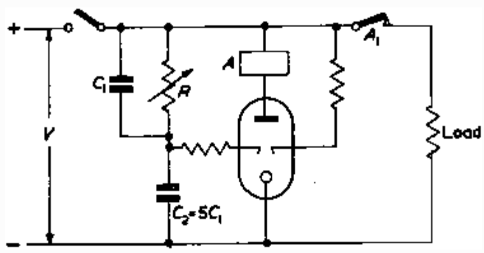

Fig. 11. Capacitive divider in constant energy relay

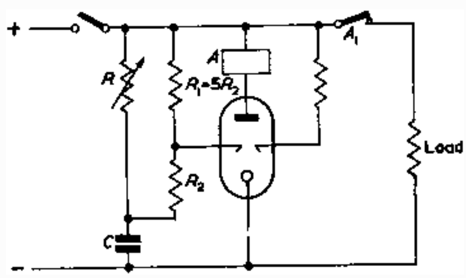

Fig. 12. Alternative constant energy relay

the pre-ignition current. The main disadvantage of this arrangement is that the resistors R_1 and R_2 must always be much smaller than R if the value of K is not to change as R is adjusted to set the energy. There is therefore an appreciable power loss in resistors R_1 and R_2. This can be avoided by the use of one of the circuits given in Figs. 11 and 12. If the circuit of Fig. 11 is used, the value of K is dependent on the values of the two capacitors so that if components having normal tolerances are used K can vary over a wide range. As an example, if C_2 is 20 per cent high, K will be 30 per cent below the nominal value. This will completely upset the correct operation of the circuit.

The arrangement of Fig. 12, which only requires one more resistor than the basic circuit of Fig. 4, is to be preferred. The value of K can easily be adjusted if a potentiometer is used for R_1 and R_2, or alternatively high stability resistors can be used for these units. Adjustment of resistor R to set the required energy level has no effect on the value of K. However, with this circuit the energy level is not linearly dependent on R, and if a calibrated scale is required it must be non-linear. The calibration of a scale with units of time is not very informative with this kind of device, since the scale can only be correct at one value of supply voltage. If a potentiometer across the supply voltage is adjusted in order to set up a time scale correctly at nominal voltage and so compensate for capacitor tolerance, the supply voltage range over which operation is satisfactory will be changed. This change is not great with the circuits described here, but the use of a potentiometer across the supply involves an additional power loss. In any timer circuit the maximum internal resistance of such a potentiometer should be much lower than the minimum value of the charging resistor. If this is not so the potentiometer will have a larger effect on shorter times than on longer ones, whereas to compensate for capacitor tolerance over the whole range of the control a constant multiplier is required. An alternative to the use of a potentiometer is the use of a movable scale which is set up on test to allow for the tolerance of the capacitor. The use of capacitors with a closer tolerance than the usual ± 20 per cent will simplify both the circuit and the initial setting up with any of the arrangements described.

A number of alternative circuits capable of giving constant V^2t characteristics will now be considered. They are all based on the principles derived above, but are not so generally useful as the schemes given previously. The charging curve obtained with those circuits is an initial step of amplitude KV followed by an exponential wave of amplitude $(1 - K)V$, as can be seen from equation (8). The

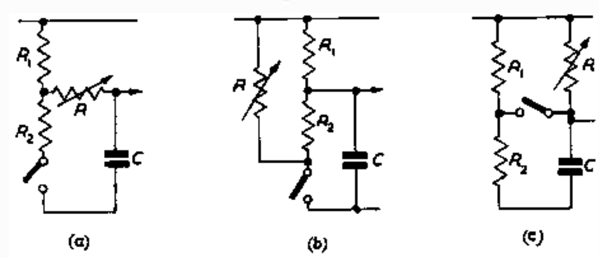

Fig. 13. Circuits with initially charged capacitor

same characteristics can be obtained if the capacitor in a simple RC circuit is initially charged to a voltage KV. This has been mentioned by Hercock and Neale[2]. The arrangements of Fig. 13 give simple methods of obtaining this result, in each case the switch shown being opened to initiate the timing cycle. All three circuits involve an extra drain across the supply and have a long resetting time because of the necessity of recharging the capacitor to the correct level. If a supply transformer and rectifier are used to supply the voltage V, it is possible to obtain voltage KV from a tapping on the transformer with an auxiliary rectifier and smoothing capacitor. A similar method can be used with the circuit of Fig. 10 to avoid the power loss due to resistors R_1 and R_2. If a separate winding is added to the supply transformer, a voltage KV can be added in series with the cathode or the trigger electrode of the tube to obtain constant V^2t characteristics. Such methods are of use in certain applications, though the circuits mentioned previously are more generally useful.

Using the principles given, it is possible to produce timers which give a constant energy within a few per cent over the complete range of permissible trigger tube anode voltage variations. These timers are well suited to applications where a heating load is controlled, such as welding or oven controls.

Acknowledgments

The author wishes to thank the General Electric Company Ltd for permission to publish this article.

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LETTERS TO THE EDITOR

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

Triode Transfer Resistance

DEAR SIR,—It has been suggested[1] that, for small values of current, an exponential relationship exists between applied voltage and resultant current in a thermionic diode. An extension of this relationship allows a novel approach to the non-linear transfer characteristic of a triode in terms of a concept of transfer resistance. In a recent paper[2] it was shown that a common emitter transistor amplifier may be considered as a transfer resistor defined as:

$$R_t = |V_{be}/I_c| \quad V_{ce} = k$$

and that this value R_t was an exponential function of V_{be}. The same concept may be applied to thermionic valves such that:

$$R_t = |V_t/I_a| \quad V_a = k$$

where V_t is an equivalent voltage, always positive in sign and defined as the algebraic sum of the negative grid bias, $-V_b$, and a constant positive voltage, V_k, which is greater in magnitude than the cut-off voltage.

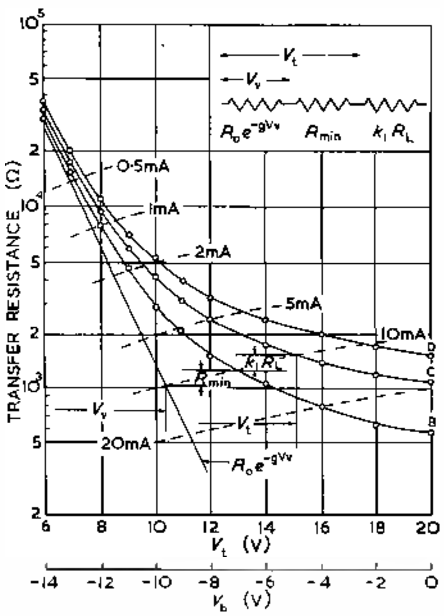

Fig. 1. Transfer resistance of 6J5
Anode supply −250V V_k = 20V

A typical plot of R_t versus V_t for a 6J5 valve is given in Fig. 1. The curve AB represents the static condition of an anode supply voltage of 250V with no anode load resistor. This resistance characteristic is similar to that obtained[3] for certain types of varistors and may be analysed in terms of the equivalent circuit shown in Fig. 1. The dotted lines in this figure represent lines of constant anode current and, as shown, allow the transfer resistance to be written as:

$$R_t = R_o\,e^{-gV_v} + R_{min} \quad \ldots\ldots\ldots (1)$$

The equivalent circuit corresponding to this condition is shown in Fig. 2(a). The addition of a resistive anode load modifies the transfer resistance as illustrated in the curves AC and AD corresponding to anode loads of 5kΩ and 10kΩ respectively. The addition of an anode load, R_L, results in an addition of a further resistor k_1R_L to the transfer resistance and:

$$R_t = R_o\,e^{-gV_v} + R_{min} + k_1R_L \quad \ldots\ldots (2)$$

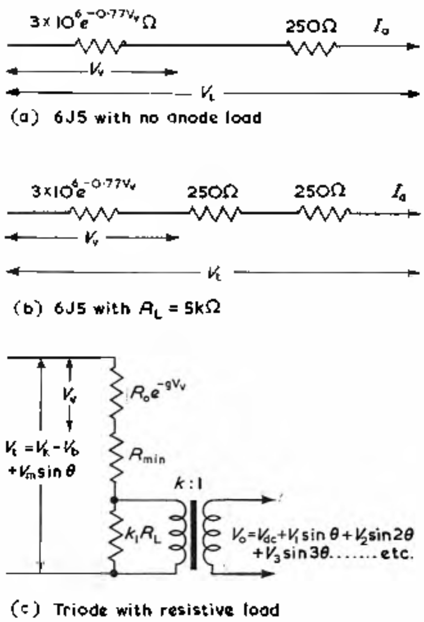

Fig. 2. Equivalent circuits

The equivalent circuit for this condition is shown in Fig. 2(b).

An overall equivalent circuit for the stage may now be postulated as shown in Fig. 2(c). The instantaneous input voltage V_t is shown as the sum of the bias voltage, V_b, the constant added voltage V_k and a sinusoidal input $V_m \sin\theta$. The instantaneous output voltage V_o is developed across the secondary winding of an ideal d.c. transformer of step up ratio k_1:1. V_o consists of a d.c. component V_{dc} plus the amplified component of the input signal, $V_1 \sin\theta$, and various harmonics, $V_2 \sin 2\theta$, $V_3 \sin 3\theta$, etc.

The voltage relations in Fig. 3(c) can be written as:

$$V_t = V_v \left[1 + \frac{R_{min} + k_1R_L}{k_1R_L} + \frac{R_{min} + k_1R_L}{R_o} e^{gV_v}\right] \quad \ldots\ldots\ldots\ldots (3)$$

It has been shown[3] that in circuits of this nature an approximate explicit relationship between V_t and V_v can be established in the form:

$$V_t = (1/k_3)\sinh(V_v/k_2)$$

where k_2 and k_3 are constants which can be determined analytically. This allows the output voltage to be written as:

$$V_o = \frac{R_L}{k_1R_L + R_{min}}[V_t - k_2 \sinh^{-1} k_3V_t] \quad \ldots\ldots\ldots (4)$$

Since the inverse hyperbolic function is readily differentiated a Taylors expansion of equation (4) can be achieved, thus allowing the output voltage to be written in terms of a Fourier series.

It will be noted that the value of k_1, 0·05, illustrated in Fig. 1 corresponds to the inverse of the accepted value of μ, 20, for this particular valve. In addition the output impedance of the stage, as seen from the secondary of the transformer will appear as R_L/k_1 in parallel with the incremental values of R_{min} in series with $R_o\,e^{-gV_v}$. Thus the incremental value of these internal resistances corresponds to the conventional incremental anode resistance.

Yours faithfully,

G. W. HOLBROOK and T. B. MAHOOD.
Royal Military College of Canada,
Kingston, Ontario.

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Stabilized Power Supplies

DEAR SIR,—The interest in and enthusiasm for electronically stabilized power supplies seems to be increasing, but we feel that some of this enthusiasm is misdirected and that more thought should be given to the use of such devices. In our view it seems that the designers of these gadgets have to some extent lost touch with reality, and regard them as an end in themselves and as some sort of signal source, rather than the necessary nuisance that they are. In our experience d.c. supplies for valve electronic equipment falls roughly into four classes:

(1) Variable and fixed portable supplies for bench use.
(2) Supplies intended to permanently power instruments and smaller collections of apparatus.
(3) Large supplies for complete systems.
(4) Precision reference supplies.

Users of type (1) supplies are perhaps better catered for than the others and we shall say nothing about this class.

Type (2) equipment is normally of the series valve −d.c. amplifier type and should be assessed on minimum nuisance value in the complete equipment. Over elaboration should be sternly guarded against and in our experience, for the majority of cases, a single pentode type amplifier is adequate. The performance, if properly designed, is only slightly inferior to its more complicated long-tailed pair counterpart.

Type (3) is normally capable of delivering several amperes and here the main electrical requirements are low output impedance and reasonably low hum. The electronic stabilizer is not normally used here, but rather the higher efficiency magnetic amplifier or constant voltage transformer types.

Type (4) is very special and complexity and cost must be accepted.

From the above it appears that most users of stabilized power supplies will be interested in groups (2) and (3). It is unfortunate, however, that many manufacturers of 'off the shelf' power units restrict themselves to type (2) supplies but devote much effort to rather useless improvements of their electrical performance. For instance a decrease in output impedance from 0·5 to 0·1Ω for a small supply is something that has little interest for most people, but if it is achieved by using more components, the decrease in reliability will give more cause for concern. We feel that manufacturers and designers should stop regarding stabilized power supplies as some sort of highly interesting d.c. amplifier, and ask themselves the following questions:

(1) Is this the simplest and most reliable device with respectable performance we can make?

Do not always take a customer's specifications seriously. If he quotes 0·1Ω output impedance, 0·1 per cent stability, 1 000 : 1 stabilization ratio etc., he has probably merely cribbed that from somebody's catalogue.

(2) Is it in a convenient and usable form?

The little hot cubes that seem to be so popular are difficult to house and some of them really reach surprising temperatures in an enclosure, and for rack mounted equipment many manufacturers still cling to the old horizontal inverted-tray type of chassis although it will produce the most annoying hot spots in vertical enclosed racks.

(3) Has it some real protection against overloads and faults?

This seems to be a problem most manufacturers solve with a fuse link and a lot of faith. In our experience only thermal type cut-outs are of any use at all in input circuits.

To sum up, in our view many of the criteria used to evaluate and sell commercial power supplies have little relevance to its user. The most important single quality is reliability which is easiest to achieve in a simple device. Series valve stabilizers are inherently somewhat clumsy and other solutions are often worth searching for.

Yours faithfully,
A. VAUGHAN, B. SAGNELL,
and J. SHARP
CERN-PS Electronic Workshop
Laboratory, Geneva, Switzerland.

A Regenerative Modulator Frequency Divider Using Transistors

DEAR SIR,—The writer, on first reading Mr. Butler's article (ELECTRONIC ENGINEERING, February, 1959) was mislead into thinking that the circuit would not produce an exact sub-multiple of the input frequency. This arose from the description of block B, Fig. 2, as a mixer-amplifier*. This appears incorrect and it seems worth stressing that the general method described only works if a frequency multiplier is involved. If block B is a mixer it will be seen that a small change in the phase-shift/frequency characteristics of the regenerative loop could cause a small change, δ, in the output frequency (see Fig. A).

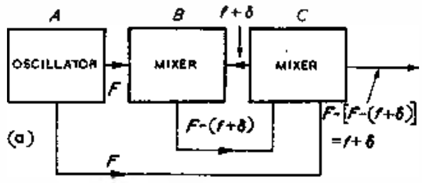


Fig. A. The effect of Block B as a mixer

If, on the other hand, block B is a frequency multiplier, Fig. B will apply.

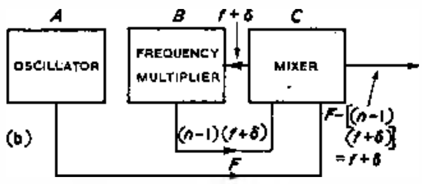


Fig. B. The effect of block B as a frequency multiplier

Since, by definition, $F = nf$ one can re-write the expression for the output frequency

$$nf - (n - 1)(f + \delta) = f + \delta$$

which is only satisfied when $\delta = 0$. Thus the only error in f will be caused by an error in the multiplication factor n. n can be well stabilized provided that the tuned circuits in blocks B and C are sufficiently sharp and stable.

If some connexion is made between blocks A and B, and mixing can take place in B, two modes of oscillation are possible, one by multiplication, the other by mixing. However, the systems will only oscillate in the mode giving adequate loop-gain. The mixing mode can give rise to the error δ and should therefore be avoided.

On looking at Mr. Butler's circuit, Fig. 5, it appears that the '90kc/s mixer-amplifier' is, in fact, a multiplier. T_2 is tuned to 90kc/s and its impedance at 100kc/s signal should appear at X_3 collector. The mixing action is likely to be small since collector current is largely independent of collector-emitter voltage over most of the collector voltage swing.

Yours faithfully,
S. J. M. WHITFIELD,
Epsom.

* *on p. 72, last sentence and subsequently on p. 73, describing Fig. 5.*

The Author Replies:

DEAR SIR,—Mr. Whitfield has evidently misunderstood my description of the action of unit B in Fig. 2. This he describes as a frequency multiplier and states that it is incorrectly described as a mixer-amplifier. Yet in the last paragraph at the foot of page 72 it is quite clearly stated that "the 10kc/s output from C is fed to B where its ninth harmonic is selected to give the desired 90kc/s output." Surely this makes it plain that frequency multiplication is involved? However, when high division ratios are required, a high order of frequency multiplication is called for in unit B and such high harmonics are of very low level. Extremely high Q circuits would be needed to isolate the desired harmonic, although straight frequency multiplication alone, is adequate when division ratios between 2 and 5 are required.

It is for this reason that a mixing process is also introduced and Unit B is described as a mixer-amplifier. If incorrectly applied this procedure can certainly result in an output which is not a true sub-harmonic of the input frequency. When the transistors are properly biased and the relative signal levels correctly adjusted there is a self-synchronizing action which locks the output frequency on an exact integral sub-multiple of the drive frequency.

If the additional cost and complexity is tolerable, a better circuit arrangement is to use successive stages of division by 2 and 5 rather than to attempt single-stage division by a factor of ten. In this case the pure frequency multiplying action can be exploited in Unit B, but this simply does not work at high division ratios, even when using abnormally high Q tuned circuits. If the 100kc/s input to B is disconnected the action stops.

Finally, it must be understood that regenerative modulator dividers are only conditionally stable at large division factors. This is true regardless of the particular circuit arrangement that may be used. In this respect they are no worse than other dividers, with the exception of those which are derived from binary counters and these are of extreme complexity.

Yours faithfully,
F. BUTLER,
Cheltenham.

Correction

In the 'Letter to the Editor' from D. T. Jovanovic on page 301 of the May issue, the following corrections are necessary. In the caption of Fig. 2, $C = 0{\cdot}03\mu F$, $C_{1,2,3,4} = 0{\cdot}01\mu F$ and R_L = infinite. In the caption of Fig. 4, $C = 0{\cdot}03\mu F$.

BOOK REVIEWS

Calculateurs Analogiques Répétitifs

By R. Tomovic. 186 pp. 94 figs. Demy 8vo Masson et Cie, Paris. 1958. Price 3 000 fr.

BEFORE attempting to deal with the subject of repetitive analogue computers the author gives a clear account of the symbols and notations to be used. By this means his diagrams and formulae become understandable even to readers not well versed in analogue computer techniques. This is followed by an analysis of the relationships between the machine elements and the mathematical equations to be solved, first for a general case and then for a particular second order differential equation. Excellent references (in particular papers by Raymond and by Gutenmayer) enable the reader to acquire more detailed knowledge if required.

The difficult question of accuracy is neatly dealt with in the small space allocated and a timely warning regarding the need for stability criteria (requiring a knowledge of servomechanism techniques which, as he states in his preface, the author rightly assumes the reader to have). The choice of scales for the various quantities completes the purely theoretical section occupying slightly less than one-third of the book.

The main portion of the practical section is divided into linear and non-linear operations. The former deals with methods of addition and integration together with a study of the errors which are introduced by the components or elements used. The non-linear section describes methods for obtaining specific mathematical expressions (logarithms, hyperbolic, parabolic and trigonometrical expressions) as well as special function generators. Included also are short sections on the choice of the repetition frequency, methods of measurement, setting of initial conditions and power supply requirements.

A number of applications and brief descriptions of existing equipment form the last sixth of the book together with some pointers to the extension of the use of analogue computers to the solution of integral equations and the possibilities of higher repetition speeds. Seventy-four references are appended.

The list of existing equipment could have been omitted as it can so soon be out of date, but this is a minor criticism. The whole approach is along clearly defined logical lines and the author is to be congratulated on producing a book not only useful to the analogue computer expert wishing to enter the repetitive machine field, but also as a good introduction to the techniques of analogue computing in general. Admittedly the reader interested in the latter would find a number of gaps in his knowledge and would need to amplify his reading but this is in no way a criticism of the work, the author having deliberately limited himself.

To the reviewer's knowledge no similar book exists (although, of course, a large proportion of the techniques discussed occur in non-repetitive machines) and it is to be hoped that English editions of these monographs of the Association Internationale pour le Calcul Analogique, of which this is the first, will not be long delayed.

G. ASHDOWN

Introduction to Symbolic Logic and its Applications

By R. Carnap. 234 pp. 5 figs. Demy 8vo. Dover Publications Inc., New York. Constable & Co. London. 1959. Price 15s.

THIS book originally appeared in German in 1954 and has now been published in English. Developing the subject from elementary concepts and simple exercises, the author carries the student through the construction and analysis of relatively complex logical languages. The latter part of the book gives a detailed treatment of the applications of symbolic logic to the clarification and axiomatization of various theories in mathematics, physics and biology. The logic of relations is given extensive treatment.

Principles of Transistor Circuits

By S. W. Amos. 167 pp. 105 figs. Demy 8vo. Iliffe & Sons Limited. 1959 Price 21s.

THIS volume has been specially written by a member of the Engineering Training Department of the BBC to assist those who require an introduction to the design of transistorized equipment, and it is intended for professional designers, students and amateur constructors. Assuming no previous knowledge of the subject, it devotes the first two chapters to explaining the physical processes occurring in transistors. Subsequently, the main emphasis is on the application of these principles to the practical problems of design, and the bulk of the book is concerned with the determination of such quantities as input resistance, stage gain, optimum load, power output, values of coupling capacitors and transformer winding inductances.

A large number of worked examples are included, the mathematics of which are confined to simple algebraic manipulation. These show the order of the practical magnitude of the various quantities. Although the design of amplifiers and receivers is given the greatest prominence, some details are also included of photosensitive devices, transistor relaxation oscillators, and the newer types of transistors which may extend the frequency range over which semiconducting devices can operate satisfactorily.

Introduction to the Design of Servomechanisms

By J. L. Bower and P. M. Schultheiss. 510 pp. 50 figs. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1958. Price 104s.

THE authors state that the book deals with topics covered in the servomechanisms course at Yale University, the earlier chapters are intended for the final year student and the later chapters for the post-graduate course. For the undergraduate engineering student the book has too much detail; the methods used for calculating the stability etc. of control systems are not clearly stated, instead many illustrative examples are given from which the general method has to be deduced. For the post-graduate student, however, the book may be recommended especially if studied simultaneously with the well-known text 'Theory of Servomechanisms' by James, Nichols and Phillips, to which frequent reference is made. The fundamental mathematical methods presented in this earlier text are here developed in much detail.

The first chapter gives examples of control systems used in many branches of engineering; this is followed by a good discussion of the mathematical methods used in the later chapters. The next chapters discuss, in some detail, the methods of Hurwitz, Nyquist and Bode as used for determining the stability of control systems. The Bode method is used throughout the book, the presentation here is unusual as the Bode diagram ordinates are not plotted in decibels but directly as the logarithm of the vector magnitudes. The authors then discuss, as suggested post-graduate topics, performance criteria, minor loops and the root-locus technique. The discussion of performance criteria is excellent, it includes much detail about M-curves, and the relation between the frequency-domain concepts of pass-band and phase-shift and their time-domain equivalents. Another unusual feature is the development of the error-series to determine the errors present in control systems a given time after the input has been subjected to a known time function. A study of the performance of regulators when load disturbances are present is also included, the concept of 'compliance' as a figure of merit is introduced here. The chapter on the root-locus technique fails, not for lack of detail, but because of confusion between the methods used for construction of the locus and the application of the method to system synthesis. The chapters describing statistical methods and non-linearities in servomechanisms, are a good introduction to these topics which now form the basis of many post-graduate studies. The book discusses mainly the design of control systems, hence, as would be expected, only a brief description is given in an appendix of the many devices which are now used in control systems.

The book is notable for the large number of problems which are included, they use as examples control systems encountered in all branches of engineering. It is well produced and there are only a few minor errors. For the reader who requires a text intermediate between an elementary treatment and a full mathematical analysis of control systems the book may be recommended.

J. V. PARRY

Technology of Printed Circuits

By P. Eisler. 405 pp. 151 figs. Demy 8vo. Heywood & Co. Ltd. 1959. Price 60s.

IT is perhaps appropriate that this book should be written by Dr. Paul Eisler, who developed the first etched foil chassis for radio equipment in 1942. The book covers the story of printed wiring and printed circuits in considerable detail. The early chapters give the author's point of view of the start of printed wiring in this country, although this might not be agreed by others. The remaining chapters deal with the etched foil system in great detail and there are several supporting chapters contributed by experts in other fields such as that on microwave printed circuits. There is a very useful chapter on printed circuit trouble shooting contributed by the Admiral Corporation of the United States of America, which should be of value to radio and television servicemen in this country. The book is a little lengthy and too much time is spent on details such the regeneration of ferric chloride which might well have been included as an appendix.

The book covers the design and layout of printed wiring and gives sufficient information to enable a newcomer to appreciate some of the problems involved. A chapter on laboratory routine contributed by G. Parker is particularly useful here. Automatic assembly systems and automatic component insertion machines are described but, strangely, little detailed information on dip soldering techniques is given. There are many systems of dip soldering, some better than others, and it would be expected that these might have been dealt with in the same detail as that in the rest of the book. The chapter on printed components, while useful, illustrates transformer coils, but does not mention the associated problem of heat dissipation. Much work is also being done both in this country and in the United States of America on film components, particularly evaporation techniques for nichrome and chromium resistors and on dielectrics evaporation for capacitors, which might have been included in this chapter.

The book does, however, give a great deal of detailed information on printed wiring and printed circuits and will make a useful contribution to the literature. The bibliography (contributed by Mrs. K. Bourton) is probably one of the most comprehensive produced in this country. The book should certainly be useful to all those concerned with designing, using or servicing printed wiring equipments.

G. W. A. DUMMER

British Instruments Directory and Buyer's Guide

600 pp. Demy 4to. United Science Press Ltd. 1959 Price 42s.

THIS comprehensive directory and buyer's guide has been published in co-operation with the Scientific Instrument Manufacturers' Association (SIMA) and covers scientific and industrial instrumentation for research and production in the following main fields:

Measurement and control of temperature, pressure, humidity, flow, weight, texture, dimensions: inspection and gauge testing, manual and automatic: data processing, computers: optics, ophthalmics, navigation, meteorology, microscopy, astronomy, surveying and levelling: physics, mechanics, pneumatics, hydraulics and electrics: analysis: laboratory ware: atomics, electronics and nucleonics.

The twelve associations who have combined to produce this directory containing over 2 500 instruments and components featured in the classified list are:

SIMA—Scientific Instrument Manufacturers' Association: BEAMA—The British Electrical and Allied Manufacturers' Association: BIMCAM—British Industrial Measuring and Control Apparatus Manufacturers' Association: BLWA—British Laboratory Ware Association Limited: BLSGMA—British Lampblown Scientific Glassware Manufacturers' Association Limited: BPMA—British Photographic Manufacturers' Association: BPGMA—The British Pressure Gauge Manufacturers' Association: BSIRA—British Scientific Research Association: DOMMDA—The Drawing Office Material Manufacturers' and Dealers' Association: EEA—The Electronic Engineering Association: GTMA—The Gauge and Tool Makers' Association: RECMF—Radio and Electronic Component Manufacturers' Federation.

The chief Sections in the directory are: (1) Associations allied to the instrument industry. (2) British Standards Specifications. (3) Consultants, engineers for instrumentation schemes. (4) Manufacturers of prototypes and small batches. (5) Instruments and components with manufacturers. (6) Glossary in French, German, Spanish and English. (7) Addresses of manufacturers and their overseas agents.

Elliptic Integrals

By H. Hancock. 101 pp. 20 figs. Demy 8vo. Dover Publications Inc., New York. 1959. Price $1.25.

THIS book will be of help to those whose work involves differential equations containing cubics or quartics under the root sign, as in electrical theory, theory of the potential, resistance and elastica problems, and shows how these integrals may be changed into standard forms. In order to avoid the necessity of referring to another book, 5-place tables of these normal forms are included in Chapter VI, so that the integrals may be calculated to any degree of accuracy required.

Guide Technique de l'Electronique Professionelle

1,100 pp. 2 volumes. Demy 4to. 2nd Edition. Publicité et Editions Techniques, Paris. 1959. Price 5·500F.

THE second edition of the Buyers' Guide and Directory of the French electronic industry is now published in four languages (English, French, German and Italian) and is aimed specifically at the European Common Market.

Enlarged to 1 100 pages the directory is divided into sections as follows:

(1) A short account of the French electronic industry and its achievements.

(2) Names and addresses of advertisers and their range of products.

(3) A two-language, technical catalogue section. Information of the manufacturers' products in this section is given in the language of the country of the manufacturer (principally French) and then in English which is now accepted as the lingua-franca in the field of electronics.

(4) Directory of manufacturers and importing firms and their range of products.

(5) Cumulative four-language glossary of components and general products and directory.

(6) Structure of the French Electronic Industry and its professional and trade associations.

(7) Professional Directory of Government Departments and Scientific Research Organizations.

Synthesis of Linear Communication Networks Vol. I and II

By W. Cauer. 866 pp. 481 figs. Medium 8vo. McGraw-Hill Publishing Co. Ltd. New York, London. 1959. Price 151s.

THE publication of this English translation of the second edition of Wilhelm Cauer's celebrated work is an event of considerable importance for English-speaking workers in the field of electric network theory. Much of the contents of the book is already known to those unacquainted with the German language through the writings of E. A. Guillemin, but this translation will provide ready access to the classic treatise on the theory of communication networks.

Professor Cauer was born in 1900. He studied electrical engineering at the Technische Hochschule Berlin and later studied mathematics and physics at the Universities of Bonn and Berlin. In 1926 he obtained his doctor's degree from the Technische Hochschule Berlin with the thesis 'The Realization of Impedances with Prescribed Frequency Dependence'; it was this thesis which determined his future career. He spent a short time at the Massachusetts Institute of Technology where he suggested to Dr. Otto Brune the research which led to a basis for a theory of network synthesis. Eventually Cauer became head of the Mix and Genest laboratories in Berlin and lectured network theory in the Technische Hochschule Berlin. In tragic circumstances, he died in 1945.

At the time Cauer began to concern himself with the theory of electrical networks, the subject of analysis had reached an advanced state of development and attention was turning to the new problem of synthesis. Cauer recognized three aspects of the problem: physical realizability (the determination of the classes of functions of the complex frequency variable corresponding to possible networks), approximation (the determination of physically realizable functions which satisfactorily approximate the desired network behaviour), and equivalence (the determination of all networks electrically equivalent but not necessarily topologically equivalent to a given realization). He applied himself to this approach and, in the language of linear algebra, complex variable theory, and the combinatorial topology of linear graphs, established the modern theory of electric networks. In 1941 he published the essential results of his labours in the book *Theory of linear alternating current circuits.* With the help of manuscripts left by Cauer, Dr. Wilhelm Klein and Dr. Franz M. Pelz produced a revised edition which was published in 1954.

A painstaking effort has been made to bring the translation into line with current network theory literature but at the same time to preserve the character of the original book. The title has been changed to one more suited to its contents. Unlike the original, the translation is presented in two volumes bound together. Two of the original six appendices now follow Chapter 7 and the second volume begins with Chapter 8. The effect of this is that Volume I contains the principles of network analysis, two-port theory, physical realizability, reactance functions, and network design on the image-parameter basis. Volume II deals with design on the insertion-parameter basis, band separation networks, and equivalent reactance-networks. A notable feature of the translation is an additional appendix associated with Volume II giving a concise account of advances in the subject since the German edition was first published. Included in this appendix is a section by Dr. F. H. Effertz on the relation between the stability boundary surfaces of linear and non-linear servo-mechanisms and the realizability boundary surfaces of some classes of frequency characteristics of electrical networks.

According to the preface, the translation 'had no other sponsorship but the enthusiasm of the translators and those aiding them'. They are to be congratulated on a remarkable undertaking.

S. R. DEARDS

Modern Electronic Components

By G. W. A. Dummer. 467 pp. 232 figs. Demy 8vo. 1st Edn. Pitman & Sons. 1959. Price 55s.

WHEREAS the four previous books on components by the author dealt with resistors and capacitors, this new book contains not only a summary of the four previous books but is extended very considerably to include information on many other components not presented previously in book form.

Its 467 pages are divided up into 32 chapters and contain information on Ministry specifications and codings, plugs, sockets, sleeving, relays, switches, cables, batteries, inductive and magnetic materials. Other subjects dealt with include tropicalization, arctic rating, transistors, casting resins, reliability, effects of shock, faults and test methods to name but a few.

The book closes with a chapter on packaging and preservation.

Each chapter has a carefully selected bibliography to assist the reader with further relevant information and there is scarcely a page without an illustration of some kind, either sketch or photograph, to enhance the descriptions in the text.

There are many tables and lists of figures on information likely to be of assistance to electronic engineers. This is exemplified by the table on page 232 which shows voltage and temperature characteristics of typical connecting wires including various grades of p.v.c., polythene, and p.t.f.e.-covered wires together with dimensions and other remarks. There is also a comprehensive table on sleevings including silicone-bonded glass yarn, p.v.c. sleeves, oleo resinous enamel, silicone-bonded asbestos fibre and many others.

All the multiplicity of modern type plugs and sockets, both proprietary and Service patterns are dealt with, and pages of detailed drawings show the exact construction of each.

The chapter on batteries is interesting, dealing as it does with the silver-zinc cell, solar and nuclear batteries.

The chapter on miscellaneous components contains such items as tag panels, strips, stand-off insulators, rotary sealed couplings, etc.

There is also a chapter on future trends.

This useful work will undoubtedly find its way into most technical libraries and should be a 'must' for design and development engineers. Its prolific descriptions and illustrations also render it quite suitable for the student.

The author has obviously made a sensible selection from the mass of available data, gathering it together under one cover to present the book in its existing form. It fills a real need in the electronic industry.

R. H. MAPPLEBECK

High Fidelity Sound Reproduction

Edited by E. Molloy. 208 pp. 50 figs. Demy 8vo. 2nd Edition. Geo. Newnes Ltd. 1959. Price 20s.

THE aim of this book is to bring together authoritative and balanced information on the technique of 'hi-fi' in terms that will be easily understood by the engineer, both professional and amateur, and by the layman wishing to obtain a sound technical understanding of this subject. In this second edition, a chapter has been added on stereophonic disk records and other revisions made.

THE INTERNATIONAL TRANSISTOR EXHIBITION

A description, compiled from information supplied by the manufacturers, of a few of the exhibits at the International Transistor Exhibition held at Earls Court, London from 21 to 27 May.

(Voir page 437 pour la traduction en français: Deutsche Übersetzung Seite 442)

BELLING & LEE LTD

Great Cambridge Road, Enfield, Middlesex

GLASS SEALS

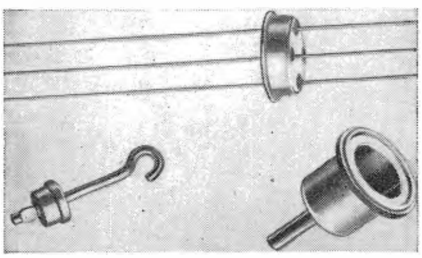
(Illustrated below)

By the use of special manufacturing techniques, Belling & Lee Ltd are now producing a range of compression and matched glass seals that ensure and maintain a reliable performance even under the most adverse conditions.

Among those exhibited were the L.1440—a single terminal glass seal, L.1441—a rectifier glass seal, L.1442—a class 30 transistor seal, and L.1459—a lead-through glass seal.

The L.1440 is primarily designed for individual leads to a power transistor, although it can be used in certain applications where a single insulated wire has to be hermetically sealed into equipment. The illustration shows: (top) the L.1442 transistor seal which conforms to an American Jetec 30 specification and is intended for use with low power transistors: (bottom left) the L.1440 single terminal glass seal primarily designed for individual leads to a power transistor and (bottom right) the L.1441 rectifier glass seal which conforms dimensionally to to the American standard for this type of component.

EE 11 751 *for further details*

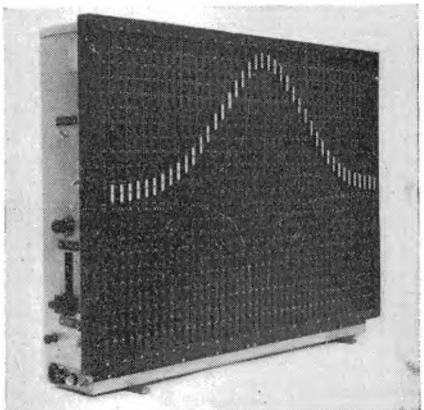


BRITISH BROADCASTING CORPORATION

Broadcasting House, London W.1

FOURIER TRANSFORM GENERATOR

(Illustrated above right)



This apparatus displays the cosine and sine transforms of any function which can be represented by not more than forty-one equispaced ordinates, of which the central ordinate must represent the function at the abscissa zero. The slider of each of forty-one potentiometers is set to the value of the function at the corresponding ordinate. Either the sine or the cosine transform is selected by means of a switch.

Each potentiometer is connected to one of twenty harmonically related sinusoidal push-pull voltages and the voltages on all the sliders are combined to form the output transform. Each frequency is connected to two potentiometers symmetrically disposed about the central potentiometer, the frequency increasing as the order of position from the centre. For the cosine transform the sinusoidal voltages are phased to reach a maximum amplitude simultaneously once per cycle of the fundamental, and the outputs of all potentiometer sliders added. For the sine transform the phasing is such that all voltages pass simultaneously through the zero in a positive direction, and the outputs of the sliders on the left-hand side are subtracted from those on the right.

The main items in the circuit are a master oscillator operating at 120 times the fundamental frequency of 500c/s, frequency dividing chains, and band-pass filter units. Of the twenty output frequencies, twelve are filtered directly from the output waveforms of dividers, and these waveforms are also intermodulated in suitable pairs to yield the other eight frequencies. The apparatus uses 159 germanium junction transistors, predominantly of an audio frequency type (OC71). For the 60kc/s oscillator, however, and for the output amplifier, which must not introduce phase distortion, an r.f. type (OC44) is used.

The oscillator is amplitude stabilized. Frequency dividers are of two types: Eccles-Jordan circuits for binary division, and a circuit due to Chaplin and Owens for division by three and by five.

Each band-pass filter is fed with either the output of a divider or that of a four-transistor balanced modulator. The secondary of the filter is balanced and drives a pair of grounded collector output stages which have as emitter loads the two halves of the centre-tapped output potentiometers.

The filters are so designed as to minimize relative phase errors due to oscillator instability and thermal variations of components. The resultant phase stability is such that, under normal conditions, 1 per cent accuracy is maintained for at least three months after calibration.

EE 11 752 *for further details*

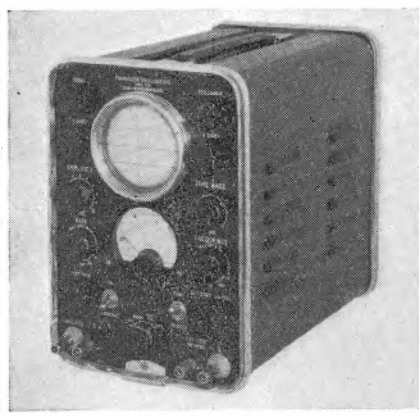
BURNDEPT LTD

West Street, Erith, Kent

STABILIZED POWER SUPPLY

The Burndept power supply unit type BE.253 provides a convenient power supply of good voltage regulation and low internal resistance for operating and developing equipment employing transistors. The output voltage is continuously variable, by coarse and fine controls, in the range from 1V to 30V: the current available being 1A. Both the output voltage and current are monitored on a 3½in meter. The output resistance is 0·05Ω and the noise and ripple less than 1mV peak-to-peak. The temperature coefficient of the unit is 6mV/°C and it may be used in ambient temperatures up to 50°C over the full voltage range and up to 50°C over a limited range.

EE 11 753 *for further details*



DAWE INSTRUMENTS LTD

99-101 Uxbridge Road, Ealing, London W.5

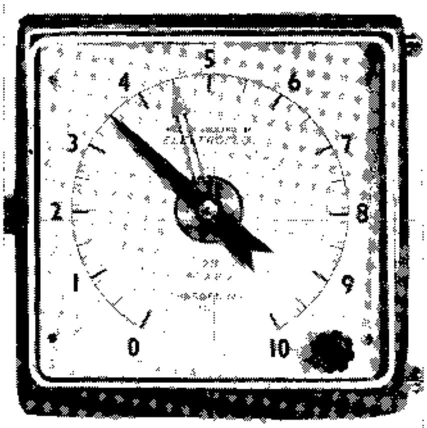
PORTABLE OSCILLOSCOPE

(Illustrated above)

This oscilloscope uses transistors throughout. In its present form it is intended primrily for the measurement and study of waveforms encountered in the vibration and acoustic fields. It may also be used as a general purpose oscilloscope up to frequencies of about 50kc/s. It is completely portable, the power supply being a small 6V accumulator contained in the cabinet. For bench use, the mains supply may be used to 'float-charge' this accumulator. The cathode-ray tube is a new very sensitive type; a 6V battery supplies the heater of this tube and also drives a d.c. convertor giving outputs of 1·3kV and 30V d.c. for the tube e.h.t. and transistor circuit supplies respectively. The time-base consists of a variable frequency multivibrator having a pulse, which after suitable amplification, is used to charge a capacitor.

This discharges through a constant current drain. The amplifier uses three stages and provides the drive to the tube from a transistor in the grounded base configuration.

EE 11 754 *for further details*



ELLIOTT-AUTOMATION LTD

34 Portland Place, London W.1

ELECTRONIC POTENTIOMETER

(Illustrated above)

Manufactured by Electroflo Meters, of the Elliott-Automation Group, the series 198 electronic pollentiometer is a continuous balance null type instrument designed to register electrical values of small potential derived from signal sources such as thermocouples, strain gauges, shaft position indicators, etc.

The input signal is applied to a modified Wheatstone bridge, across which is supplied a continuously standardized reference voltage. This produces an out of balance signal in the bridge, whose magnitude is directly related to the input error signal. The bridge output is applied to a transistor amplifier whose function is to drive a two phase 50c/s inductance motor in such a direction as to drive the moving member of a calibrated slide wire into a position of balance, whereby the bridge unbalance is nullified. Fixed to the slide wire moving member is a pointer which indicates on a 12·5in scale the value of the external error signal. These particular instruments accept d.c. input signals which are mechanically modulated by a 'chopper' relay; this approach being the well tried method of modulating small d.c. signals with minimum noise level interference.

In the transmission from servo motor to the measuring slide wire, additional control facilities are incorporated. In addition to front panel indication, three position control action together with a further re-transmitting slide wire and an alarm switch is fitted. Other forms of control include proportional and on-off, with high and low level alarms for all models.

The instrument is suitable for computer use, digital read-out and remote indication.

EE 11 755 *for further details*

A. & M. FELL LTD

1 Lambeth High Street, London S.E.1

PRECISION RESISTIVITY MEASUREMENT

(Illustrated below)

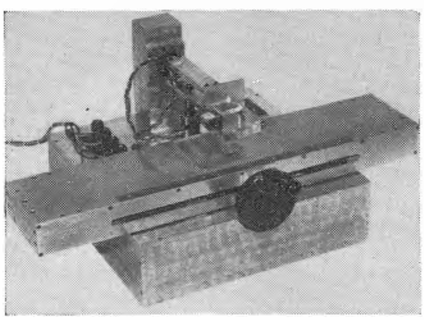
A number of resistivity test sets do not have sufficiently accurate location for the contact needles.

In this design the four needles are located in ruby guide holes, placed with as great a centre distance accuracy as the bearing holes in a watch. The needles are of tempered steel, the points being left dead hard. This design ensures a high degree of repeatability in the resistivity measurement.

The normal centre distance is $\frac{1}{2}\pi$cm, but other spacings are available to order, and the needles can be arranged either in a straight line or at the corners of a square, the straight lay-out being standard.

The needle pressure is 150 to 200 grammes, and the radius of curvature of the points is 0·1mm for germanium and 0·04 for silicon.

The head is mounted on a spring-hinged parallel motion, and a glass topped carriage with a travel of 30cm is provided for the crystal. A steel rule and index enables traverses along a crystal to be plotted.



EE 11 756 *for further details*

FERGUSON RADIO CORPORATION LTD

Great Cambridge Road, Enfield, Middlesex

TELEVISION WAVEFORM GENERATOR

(Illustrated above right)

This equipment is a portable mains or battery operated generator for use, in conjunction with monochrome television cameras and suitable video mixing units, in producing complete television picture waveforms.

It produces horizontal and vertical trigger pulses for the camera time-bases and composite (horizontal plus vertical) synchronizing and blanking waveforms for the video mixers. The use of transistors throughout this equipment has resulted in considerable reduction in size, weight, and power consumption as compared with an equivalent thermionic valve device.

The use of non-critical circuit arrangements and conservative transistor operating conditions jointly ensures good reliability and avoids the need for frequent maintenance.

The output waveforms comply fully with the BBC Specification TV42 for the 405 line system.

The equipment comprises six plug-in printed circuit boards carried in a metal framework contained in a metal case. The framework also carries a special printed

extension board which enables any one of the working boards to be withdrawn from the stack and operated under conditions such that easy access may be had to both sides for purposes of adjustment or maintenance.

The mains power unit plugs into the rear end of the frame and becomes an integral part thereof. This unit may be replaced with a battery pack where self-powered operation is required.

The master oscillator controlling the waveform generator may be locked:—

(a) To an internal quartz crystal oscillator.

(b) To a 50c/s signal derived from the mains power unit.

(c) To an exernal 50c/s reference signal.

The mains external lock is of a special long time-constant type ensuring signals suitable for operation with 'flywheel-sync' television receivers. With an equivalent time-constant of the order of 7sec, it will lock into any reference frequency from 47c/s to 52c/s within about 30sec of switching on. This feature enables the system to be brought into operation by remote switching of the power supply.

The waveform generator circuits use seventy-five 10Mc/s cut-off pnp transistors (Mullard OC44) and fourteen 6Mc/s cut-off npn transistors (Sylvania 2N94A).

The power unit contains eight pnp transistors plus one Zener diode, and provides a stabilized 12V output at up to 200mA. It can accommodate without adjustment mains supply variations from 260V to 150V.

EE 11 757 *for further details*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London W.C.2

TELECOMMUNICATIONS EQUIPMENT

Several examples were shown of transistor applications to units employed in multi-channel open wire and coaxial cable systems, including the following.

A *channel panel* which translates incoming voice frequencies to the required position in the group frequency band 60 to 180kc/s. In the reverse direction it provides demodulation to speech, and facilities are given for out of band signalling if required. A *buried repeater* for use with a 240 circuit small diameter coaxial cable system employing two negative feedback amplifier units, with the necessary equalizing networks inserted between. The amplifiers are designed to balance the cable loss, and the voltage across each is of the order of 3·5V for power feeding down the cable.

An *out of band signalling unit* has been developed to accept d.c. pulses of either -24V, -50V or E from the telephone exchange, and to convert them to pulses of 3825c/s from the channel panel, amplifies and rectifies it, and operates a transistor switch circuit which in turn operates a telegraph relay. The contacts are so arranged that the desired d.c. pulses may be transmitted to the exchange.

For use in coaxial amplitude modulated carrier systems is a *super group amplifier*. It is of negative feedback type with an overall feedback of approximately 25dB to give a forward gain of 25dB. Frequency response is better than $\pm$0·25dB and distortion characteristics are very good. A super group demodulating amplifier is available for amplifying the basic 12 circuit group (60 to 108kc/s) on multiplex amplitude modulated carrier systems. Negative feedback is employed to meet the required impedances and distortion characteristics. Current drain is 30mA at 20V.

Subscriber and junction units incorporating pulse modulators and demodulators as required in an experimental electronic exchange were also shown. The former consists of a 2-wire/4-wire termination at which speech signals (2-wire) are converted to and from amplitude modulated pulses (4-wire). Speech signals are applied via a hybrid to a simple semiconductor diode modulator to which is applied 0·5μsec pulses. The second unit is similar except that there is no hybrid and the working is 4-wire/4-wire.

EE 11 758 *for further details*

HATFIELD INSTRUMENTS LTD

Crawley Road, Horsham, Sussex

TWIN D.C. POWER UNIT

Although intended particularly for the transistor field, where two independently variable supply voltages are often essential, the power unit type LE.400 will serve as a useful addition to general laboratory equipment. The two outputs are identical and may be varied independently to give any voltage between 0 to 30V d.c. with a maximum current of 1A. Current metering for each output, automatic protection and smooth control of output voltages are featured in these units. Some further details are: source impedance less than 0·01Ω, stabilization ratio 200, ripple 1mV at full output. The mains inut transformer is tapped for 200 to 250V 50c/s mains, and a variation of $\pm$10 per cent from the nominal setting can be tolerated.

EE 11 759 *for further details*

GALVANOMETER AMPLIFIER

This amplifier features very low drift with a frequency range from 0 to 4kc/s and has an output linearity of better than 2 per cent. The input impedance is 2MΩ and at the maximum sensitivity setting of 250mA/V the amplifier will deliver $\pm$25mV into 75Ω. Subsequent to the initial fifteen minute warm-up period, drift measured at maximum sensitivity, does not exceed 3mV. Balancing and input attenuator controls are grouped on the front panel. The amplifier has an integral power supply, thus enabling several units to be operated together with complete independance. Since in practice these amplifiers may be used in a multiple arrangement, a standard 19in panel mounting assembly has been designed to accommodate eight amplifiers as plug-in units.

EE 11 760 *for further details*

KYNMORE ENGINEERING CO. LTD

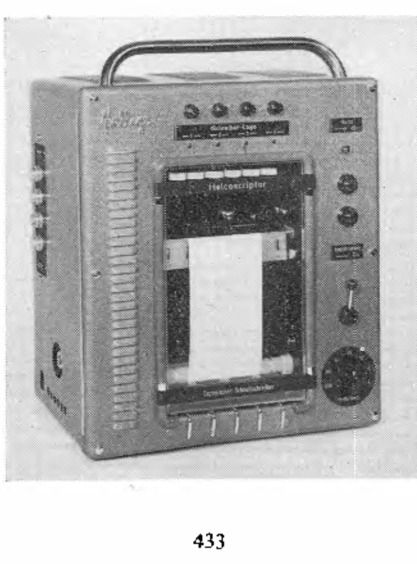
19 Buckingham Street, London W.C.2

PEN RECORDER

(Illustrated below)

The 'Helcoscriptor' manufactured by Fritz Hellige and Co of Germany and distributed by Kynmore Engineering is a high speed recorder using a heated stylus on plastic coated paper which provides an instantly visible recording in rectangular co-ordinates. It is available in two models: as a single channel recorder for the direct recording of one process; and as a multi-channel recorder for the direct recording of up to four processes.

Built-in two stage d.c. amplifiers are provided, the first stage acting as a differential amplifier with an input impedance of 1MΩ and a sensitivity of 10mm/V.



The frequency response is flat from 0 to 100c/s and 3dB down at 125c/s.

The single channel model provides two paper speeds and the four channel model seven speeds. According to the gearing fitted the paper speeds can lie between 0·5mm/sec and 300mm/sec; for extremely low speeds a special plug-in motor is available which will provide speeds down to 0·02mm/sec.

EE 11 761 *for further details*

LIVINGSTON LABORATORIES LTD

Retcar Street, London N.19

WAVE ANALYSER

The Hewlett-Packard model 302A wave analyser, which is distributed in the U.K. by Livingston Laboratories, will separate an input signal into its individual frequency components so that the fundamental, harmonics and intermodulation products may be separately measured and evaluated. It may also be used as a narrow band tuned voltmeter which will read absolute or relative levels. It uses transistors throughout and has a frequency range of 20c/s to 50kc/s.

A feature of this equipment which ensures accurate and reliable measurements is an automatic frequency control which maintains the difference frequency between the input signal and the local oscillator constant at 100kc/s even though the input signal drifts.

Another useful feature is a frequency restoring circuit so that accurate determinations of frequency at each waveform component can be made easily.

The frequency restorer output is a sinusoidal signal which is the same frequency as the measured component and this signal may be measured by an electronic counter for extreme accuracy. The amplitude of the restored frequency signal is determined by the level of the selected component and is available during normal and a.f.c. operation whenever the meter is indicating.

EE 11 762 *for further details*

JOSEPH LUCAS LTD

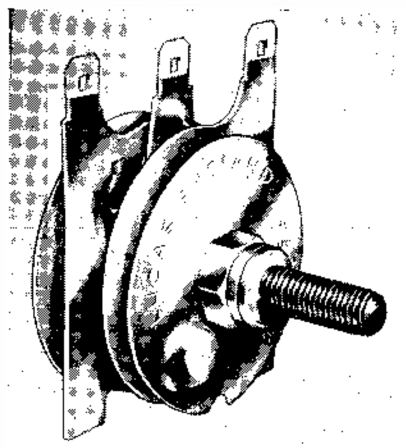
Great King Street, Birmingham 19

SEMICONDUCTOR RECTIFIERS

For several years the Lucas organization have been producing semiconductor devices for use within their own group; a number of these, including the following, are now being made generally available.

High Power Rectifiers

These rectifiers have forward current ratings of 35A and 100A when mounted on suitable heat sinks. Their good forward characteristics ensure low power loss even at these high currents. They are stud mounted devices and are manufactured with the stud of either polarity to assist in assembling them into 'bridge' configurations. They are available with peak inverse voltage ratings between 50 and 400V and they have a temperature range of -55°C to +100°C.



Silicon Rectifier Stacks

(Illustrated above)

These are compact assemblies in which the rectifying junctions are built as an integral part of the cooling fins. Their overall dimensions are only 1¼in diameter × 1⅞in, mounting is by means of a central bolt. Wires can be attached to the special tags by either soldering or quick assembly connectors.

These stacks have a forward current rating of 5A and p.i.v. ratings between 50 and 400V.

Low and Medium Power Rectifiers

These are available in a range of current ratings between 0·25 and 10A with p.i.v. ratings between 50 and 400V and are carefully constructed for use in extremes of temperature, humidity and mechanical stress. The low current types are wire-ended and the remainder stud mounting.

EE 11 763 *for further details*

MINISTRY OF SUPPLY

Three of the Research and Development Establishments of the Ministry of Supply presented exhibits showing the use of transistors, including the following.

HEARTBEAT DETECTOR

A pulse rate can be observed by monitoring the skin potentials associated with heart action. This is a practical method with passive or anaesthetized subjects, but when the subject engages in vigorous exercise, the heart signals are normally absecured by the interfering signals associated with muscular activity.

In the exhibit shown, the heart signals produce a clear output pulse even when the subject is active. Skin potential signals are picked up from two sites where the heart signals are similar and the muscle noise very different (e.g. from over opposing muscles, as with the inner and outer aspects of the forearm, using the skin over the heart as a common earth point). The two signals, after amplification, are applied to a gate circuit which gives an output pulse only on receipt of coincident signals, so that the muscle interference is ignored. The output signal is suitable for controlling a small portable transmitter if the subject must be free of leads, or for triggering an oscillator to give an audible output signal.

The original valve model was designed by the Clothing and Stores Experimental Establishment of the Ministry of Supply. The transistor version being demonstrated was developed by the Electrical Department of the Royal Aircraft Establishment, Farnborough, for use by C.S.E.E. in physiological research studies.

EE 11 764 *for further details*

MULLARD LTD

Mullard House, Torrington Place, London W.C.1

NEW TRANSISTORS

(Illustrated right)

The OC203, the latest addition to the Mullard range of silicon alloy junction transistors, is designed for general purpose low frequency applications at collector voltages of up to -60V d.c. or peak.

Features of the OC203 include low bottoming voltage of 130mV (average) at collector current of 7mA; low noise figure of 8dB (average); and low leakage current of 0·01μA (average at V_c = -10V and ambient temperature of 25°C).

Cut-off frequency is typically 1Mc/s and current gain 15. Maximum collector current is 50mA d.c. or peak, and maximum dissipation 100mW at 100°C. Junction temperature range is -50°C to +150°C.

Limited quantities are available now, and this type will shortly be in mass production.

The OC139 and OC140 are the first npn transistors to be introduced by Mullard. Both are germanium alloy junction types designed for high-speed computing circuits and other r.f. applications.

Features of the new npn transistors include high current ratings of 250mA and 200mA average; low bottoming voltage current of 220mV maximum at a collector current of 50mA; low leakage current of 3μA maximum (measured at 25°C). Maximum collector voltage is +20V d.c. or peak, and maximum junction temperature is 75°C for continuous operation, and 90°C for intermittent operation.

The OC139 has a minimum cut-off frequency of 3·5Mc/s and a minimum large signal current gain of 20 at 15mA collector current. For the OC140 these parameters are 4·5Mc/s and 50 respectively. The fall-off in gain at high currents is relatively small, the figures being respectively 14 and 35 for the OC139 and OC140 at a collector current of 20mA.

Both types have nearly symmetrical emitter and collector junctions and are controlled for reverse current gain.

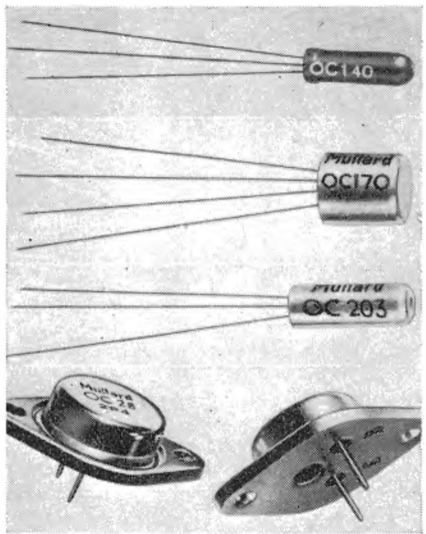
The OC28 and OC29 are industrial 6·0A power transistors designed to withstand a peak collector voltage of 60V while passing a current of 0·5A.

These new types are intended for high-voltage, high-current switching and control applications, the OC28 being particularly suitable, in addition, for d.c. convertors operating from fully charged 24V supplies.

They show marked improvement over Mullard types previously available for these purposes; apart from their much better high-voltage characteristics, they have substantially increased gain figures. At the peak collector current of 6·0A (for both types) the OC28 gives a minimum large signal current gain of 27, and the OC29 of 45.

Both types are designed for continuous operation at a junction temperature of up to 90°C (up to 100°C for intermittent use). They are made in the standard American power transistor construction which is now adopted for future Mullard power transistors.

The OC170 is the first of a new series of Mullard germanium r.f. transistors using alloy diffusion techniques, and is already in quantity production for applications in both industrial and domestic equipment.



The OC170 has an average cut-off frequency of 70Mc/s and provides a current gain of 80. Maximum collector dissipation is 60mW at an ambient temperature of 45°C and a maximum collector voltage -20V.

It will be followed shortly by a further, similar type designed for use as a mixer at frequencies reaching 200Mc/s and above.

Two new r.f. germanium alloy transistors designed specifically for digital computing and other high speed switching applications have been introduced.

These new types, OC41 and OC42, carry full computing specifications with closely controlled characteristics. High peak currents are permissible, and turn-on and turn-off times are specified. The OC41 and OC42 are intended to replace, in computing circuits, the currently-used OC44 and OC45, which were designed primarily for sine wave applications.

Both the new transistors are capable of passing a peak collector current of 125mA, and have a leakage current, in grounded emitter connexion, of 2μA (average) at an ambient temperature of 25°C. Collector bottoming voltage is 120mV (average) at peak collector current and the maximum junction temperature for continuous operation is 75°C.

EE 11 765 *for further details*

Short News Items

Pye Telecommunications Limited have recently received a contract for two Pye aircraft landing systems to be installed at Moscow airport. This is one of the first contracts to be placed in Britain by Russia following the signing of the Anglo-Russian Trade Agreement. The Pye system, which uses electronically stabilized beams, gives the pilot an instrument display in the cockpit and obviates the language difficulties encounted with talk-down systems. This is the same system recently used during demonstrations of automatic landing by the Royal Aircraft Establishment at Bedford.

E.M.I. Electronics Ltd have recently received an order from the Australian Government Aircraft Factory, Melbourne, for an EMIAC II two-module analogue computer. EMIAC II is a general-purpose analogue computer of great versatility which is suited to deal with the varying work pattern handled by the Australian establishment. It is designed to be built-up in module form to meet an expanding requirement. A large EMIAC II computer installation which when completed will be one of the most powerful of its type in Britain, is due to commence operation at the Sir W. G. Armstrong-Whitworth Aircraft Company Limited in July. This computer will be used as an additional aid to research into guided missiles and other problems associated with high-speed flight. The installation will be capable of studying performance to complete new guided weapons and aircraft systems while they are still on the drawing board.

The Radio Industry Council's Television Reception Policy Committee has been reconstituted to act for the Council in matters concerning television reception and to report to the R.I.C. Executive. The three constituent associations of the R.I.C. have nominated representatives as follows:

B.R.E.M.A.	G. Darnley-Smith, F. W. Perks, A. L. Sutherland, M. M. Macqueen, E. E. Rosen, G. W. Godfrey.
B.V.A.	J. W. Ridgeway, S. S. Eriks.
R.E.C.M.F.	P. D. Canning, E. M. Lee.

Mr. G. Darnley-Smith is chairman.

The Institute of Physics' Conference on "Some Aspects of Magnetism" will be held in Sheffield from 22-24 September, 1959. Details and application forms to attend can be obtained from the Secretary of the Institute of Physics, 47 Belgrave Square, London S.W.1. If accommodation in one of the University Halls is desired, early application is advisable.

The German Radio, Television and Phonograph Exhibition will be held from 14-23 August, 1959, in Frankfurt am Main. The purpose of the exhibition is to give all interested parties from Germany and abroad a comprehensive survey, showing the status of technical development and of the supplying potentiality of the manufacturers of apparatus and accessories. Some 200 firms have signified their intention to participate in this event so that the relevant industry will be well represented. The exhibition programme comprises radio and television sets together with transmitters, electro-acoustic apparatus, valves, phonographs and tape-recorders and their accessories and component parts as well as the various publications of the specialized publishers. The leading firms of the German industry for aerials will also show their latest models. A typical 'Aerial Street' in the outdoor space will enable the visitors to compare the various types and models..

Transitron Electronic Corporation of Wakefield, Mass., one of America's leading manufacturers of a wide range of semiconductors, announce that nine months ago they assigned 33 000 square feet of factory space and a large number of their research engineers under the direction of Mr. Nicholas De Wolf, to study the aspects of reliability for semiconductor devices. This is to enable the design engineer to be offered a semiconductor whose dependability will be the highest possible, thus reducing failure rates to the minimum, and to acquaint the engineer of this work by publishing new data sheets written especially for him. To achieve this purpose, several million dollars have been allocated and tens of thousands of semiconductor devices will be tested in quantities and conditions never before attempted.

Harwell Reactor School are holding their nineteenth Standard Course from 31 August till 18 December, 1959. Altogether, 951 students from 32 different countries have so far attended these courses, which began in September 1954 and are designed to train engineers in the techniques of reactor construction and operation, particularly in connexion with nuclear power stations. Half the places at each course are allocated to overseas students and a fee of £250 exclusive of accommodation is charged. A special course for Senior Technical Executives will be held from 21 September to 1 October, 1959. This is the ninth course of this kind and the fee is fifty guineas exclusive of accommodation. Application forms and details of both courses can be obtained from: The Principal, Reactor School, A.E.R.E., Harwell, Didcot, Berks.

Mullard Research Laboratories at Salford's, near Redhill, Surrey, are having two new laboratory blocks built on a site adjacent to their existing buildings. An extra 45 000 square feet of floor space will be provided by the new buildings, which will consist of a three and a four storey block. This will house the electronics, telecommunication, transistor applications and television laboratories which have been built up during recent years, parts of which have hitherto been operating in temporary buildings. It will also improve the facilities for the valve, semiconductor and materials research activities. The contract for the new buildings has been awarded to a local firm, James Longley & Co. Ltd, of Crawley.

The Brunel College of Technology will be holding an evening course of ten lectures on the Techniques of Non-Destructive Testing in their Physics Department at Woodlands Avenue, Acton, London W.3, every Wednesday from 7 till 8.30 p.m., commencing on Wednesday 7 October, 1959. Lectures will be given by well-known specialists in the subjects and by members of the college staff. These will be of interest to Inspectors and others engaged in the work, and also to graduates and holders of the H.N.C. in Engineering and Science subjects, who require an introduction to the methods employed in non-destructive testing. The fee for the course will be £1 and further details may be obtained from the Physics Department at the College.

Decca Radar Limited will be installing 14 Type D303 radars in seven double-ended municipal ferry boats operating between Brooklyn and Staten Island under a contract recently awarded by the New York City Department of Marine and Aviation. These ferries maintain a 24 hours' continuous service over this 1¼ mile stretch and operate directly across the main shipping lanes and through two anchorages. This order follows a contract already placed by the New York Fire Department for the Decca Type 214 River Radar, seven sets of which have already been supplied.

The 6th National Symposium on Reliability and Quality Control in Electronics will be held at the Statler Hilton Hotel, Washington D.C., from 11-13 January, 1960. Information regarding the submission of papers and attendance at the Symposium may be obtained from Mr. R. Brewer of the Research Laboratories, The General Electric Co. Ltd, Wembley, Middlesex.

The Institution of Electrical Engineers have announced that the result of the ballot to fill the vacancies which will occur on the Council on the 30 September next, is as follows:

President: Sir Willis Jackson, F.R.S.; vice-presidents: G. S. C. Lucas, O.B.E., O. W. Humphreys; C.B.E.; honorary treasurer: E. Leete; ordinary members of council—*Members*: Professor H. E. M. Barlow, C. O. Boyse, L. Drucquer, H. G. Nelson, J. R. Rylands, G. A. V. Sowter. *Associate member*: R. A. Hore.

The Department of Scientific and Industrial Research will shortly resume publication of its Technical Digests. These were last issued in 1957 and their object was to call attention to useful ideas appearing in technical literature. The Technical Digests will be published monthly as a small compendium of ideas, each presented on a separate sheet of paper. Annual subscription £3 3s. 0d. for sets of fifteen items per month, but a reduction will be offered for quantity supplies.

The Plessey Company Limited have announced that an agreement providing for the manufacture of Metallux Resistors in the United Kingdom has been concluded with Elettronica Metal Lux s.p.a. of Milan, Italy. Metallux metal film resistors, have been available through Plessey as sole U.K. agents during the past 18 months. Under the new agreement, in addition to sole manufacturing rights in Britain, Plessey will hold selling rights for the United Kingdom and all Commonwealth countries.

The 1959 Eastern Joint Computer Conference, sponsored by A.I.E.E., A.C.M., and I.R.E. will be held at the Statler Hilton Hotel, Boston, Massachusetts, from 1-3 December, 1959. Papers will be accepted on any phase of computing and persons wishing to present papers should submit by 15 August, four copies of a 100 word abstract and a 1 000 word summary. Present plans call for a single session conference, and papers will be limited to a presentation time of 20 minutes followed by a brief discussion period. At the discretion of the programme committee, papers of exceptional interest may be allowed a longer period of time for presentation, provided written request by the author is made at the time the abstract and summary are submitted. Abstracts should be suitable for inclusion in the programme of the conference. The chairman of the conference will be Mr. F. E. Heart, Lincoln Laboratory, Lexington, Mass., and Mr. H. W. Fuller, Laboratory for Electronics Inc., Boston, Mass., will direct the local arrangements. Exhibit management will be handled by John Leslie Whitlock Associates, Arlington, Virginia. Abstracts and summaries should be sent by 15 August, 1959, to: J. H. Felker, Chairman, E.J.C.C. Programme Committee, Bell Telephone Laboratories, Mountain Avenue, Murray Hill, New Jersey, Room 5C-101.

PUBLICATIONS RECEIVED

INSTRUMENT CONSTRUCTION is the English translation of the Russian monthly journal Priborostroenie, which is published in Moscow and is described as a scientific technical and production journal. Because of the time needed for translating and printing, each issue of Instrument Construction is published two to three months after the date printed on its cover, which is the date on the corresponding issue of Priborostroenie. This English translation is produced for the Department of Scientific and Industrial Research by the British Scientific Instrument Research Association and is published by Taylor and Francis Ltd, Red Lion Court, Fleet Street, London E.C.4. Single copies 15s. plus postage, annual subscription £6 post free. A special rate of £3 annually post free is available to University and Technical College libraries, for which application must be made direct to the publishers.

TECHNICAL PUBLICATIONS—THE CASE FOR A SPECIALIST SERVICE is issued by Technivision Limited and pin-points for discussion problems and consequent remedies personally encountered in the sphere of writing, illustrating, editing and publishing technical literature. This small treatise is written mainly for organizations who are on the fringe of the technical publications field and yet not quite sure of the correct method of approach. Technivision Ltd., Braywick House, Nr. Maidenhead, Berks.

FUNDAMENTALS OF NUCLEAR ENERGY AND POWER REACTORS by H. Jacobowitz is a book designed not only for the personnel of power utilities, but also for students of physics. After presenting the basic concepts in atomic and nuclear physics essential to understanding the operation of nuclear reactors, the book discusses the construction, principles of operation, cost and power output of specific plants. John F. Rider, Publisher, Inc., 116 West 14th Street, New York 11, U.S.A. Price $2·95.

RCA RECEIVING TUBE MANUAL, RC-19 covers basic valves theory and application information in the same easily understood style used in previous editions. It contains technical data for more than 625 receiving valves, including types for black-and-white and colour television, series-string applications, 12-V automobile radio receivers, and high-fidelity audio applications, and more than 95 picture tubes including colour types. Commercial Engineering, Radio Corporation of America, Electron Tube Division, Harrison, New Jersey. Price 75 cents.

FUNDAMENTALS OF RADIO TELEMETRY by M. Tepper is a book covering the field of radio telemetry, explaining its purpose and exploring its techniques. Special sections are devoted to missile and satellite telemetry data recording and processing. Specially prepared illustrations help to clarify important telemetry fundamentals. John F. Rider Publisher, Inc., 116 West 14th Street, New York 11, U.S.A. Price $2·95.

ALUMINIUM FOR INSULATED CABLES is published by Aluminium Union Limited and is the fourth in a series of electrical handbooks, giving a detailed account of cable manufacturing methods using aluminium. This 72-page booklet contains a survey of established designs of aluminium conductor and aluminium sheathed cables, and the techniques of jointing and insulation are fully treated. Aluminium Union Limited, The Adelphi, John Adam Street, London W.C.2.

GLOSSARY OF TERMS USED IN VACUUM TECHNOLOGY is issued by the Committee on Standards, American Vacuum Society, Inc. and published by Pergamon Press Ltd. This glossary is published as a reference work complete with an alphabetical index. Closely related terms have been grouped together for comparison. In many cases two or more definitions have been given for the same term, and synonyms have been listed after many definitions. Pergamon Press Ltd, 4 & 5 Fitzroy Square, London W.1. Price 10s.

THE EPOK STORY is a comprehensive booklet on resins and paint published by British Resin Products Limited. Produced as an illustrated guide to the nature and uses of Epok Surface Coating Resins, the booklet mainly discusses the uses of paint in fifteen important fields. A special section is devoted to the particular requirements of the electrical industry where great care is necessary in the choice of varnishes for insulation. For this reason special resins have been developed which impart flexibility and adhesion together with resistance to heat and damp. They also possess the required degree of resistance to acids, alkalis and oils and are widely used for the production of wire enamels and as insulation for coils and windings. Information Department, British Resin Products Limited, Devonshire House, Piccadilly, London W.1.

HOLDING AND LIFTING PERMANENT MAGNETS is a leaflet illustrating a range of power, button and pot type permanent magnets suitable for holding, lifting and positioning purposes, together with dimensional details of standard sizes available and their approximate weight lifting capacity. Publicity Department, Darwins Limited, Tinsley, Sheffield 9.

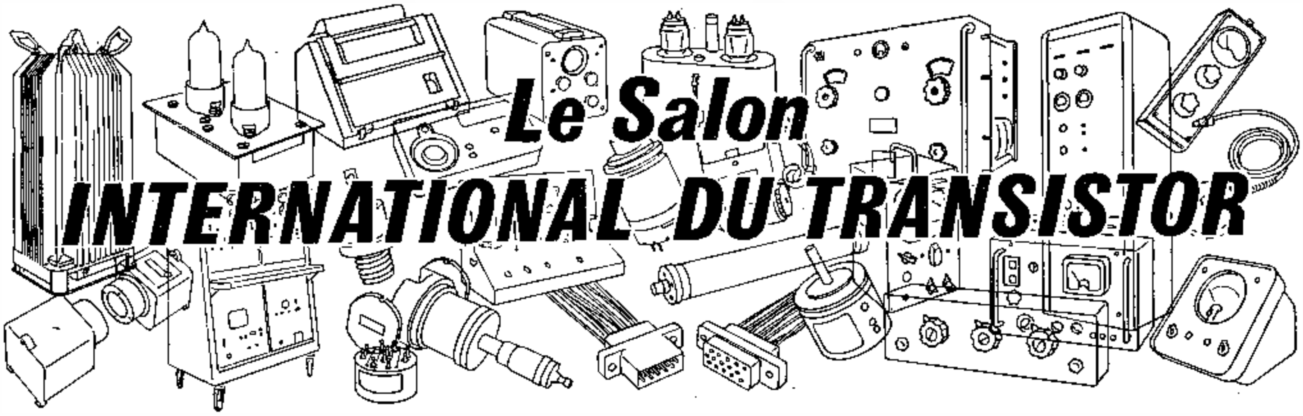
NEW DEVELOPMENTS IN TRAINING is edited by F. A. Heller of the Department of Management Studies and is the third booklet in a series on "New Developments". The five papers were first given as lectures to an audience of nearly 200 directors and senior executives at the Polytechnic, Regent Street and most of the contributions deal with industrial and commercial aspects of training and should be of value to business men and specialists who are concerned with communication and efficiency through the development of skills. Department of Management Studies, St. Katherine's House, 194 Albany Street, London N.W.1. Price 5s.

SYNTHETIC RESIN BONDED LAMINATED SHEET is the tenth in a series of reports on plastics in the tropics. This particular trial, arranged by a Ministry of Supply and British Plastics Federation Joint Sub-Committee, examines the effect of tropical exposure on the physical and electrical properties of various synthetic resin bonded laminated materials in sheet form. Published by Her Majesty's Stationery Office, London, Price 6s.

WIRELESS AND ELECTRICAL TRADER YEAR BOOK 1959 is published by the Trader Publishing Co. Ltd and includes condensed specifications of nearly 250 current commercial television receivers and information on valve and cathode ray tube connexions, with over 300 valve base diagrams. The Buyer's Guide is in two sections, one covering domestic electrical appliances, the other general radio and electrical goods. For ease of reference, the Year Book is divided into sections printed on distinctively coloured paper. Each section is separated by a card, with thumb index, giving details of contents. Trader Publishing Co. Ltd, Dorset House, Stamford Street, London, S.E.1. Price 12s. 6d. (postage 1s.).

THE PRACTICAL DICTIONARY OF ELECTRICITY AND ELECTRONICS by R. L. Oldfield defines those terms most used in electrical and electronic applications. A brief handbook section has been added after the dictionary containing formulas, tables, symbols and circuit diagrams. The handbook has been designed mainly for the student, since basic formulas and other fundamentals are included. The Technical Press Ltd, 1 Justice Walk, Chelsea, London S.W.3. Price 52s.

THE PROPERTIES AND CHARACTERISTICS OF ALUMINIUM CASTING ALLOYS is the title of a publication containing data for all the aluminium casting alloys to British Standard 1490. It is based on the original A.L.A.R. data sheets but is published in a more compact form and contains considerably more information. The Association of Light Alloy Refiners and Smelters Ltd, 3 Albemarle Street, London, W.1.



Le Salon INTERNATIONAL DU TRANSISTOR

Une description, basée sur des renseignements fournis par les fabricants, de certaines des réalisations qui ont été préséntees au Saloon International du Transistor à Earls Court, Londre, du 21 au 27 mai 1959.

Traduction des pages 431 à 434

BELLING & LEE LTD

Great Cambridge Road, Enfield, Middlesex

SCELLEMENTS EN VERRE

(*Illustration à la page 431*)

Appliquant une technique spéciale de fabrication, la Société Belling & Lee Ltd produit actuellement une série de scellements en verre comprimé, d'une sécurité d'emploi parfaite, même dans les conditions les plus défavorables.

Les modèles présentés au Salon comprenaient le L.1440: un scellement en verre à une seule borne, le L.1441: un scellement en verre pour redresseur, le L.1442: un scellement pour transistor, et le L.1459: un scellement en verre de traversée.

Quoique principalement prévu pour les conducteurs individuels du transistor de puissance, le L.1440 peut également être utilisé pour certaines applications où un seul fil isolé doit être hermétiquement scellé dans l'appareil.

On peut voir sur la gravure:

—(en haut) le scellement de transistor L.1442 qui est conforme à la spécification américaine Jetec 30 et a été prévu pour être utilisé avec des transistors de faible puissance,

—(en bas, à gauche) le scellement en verre à une seule borne, prévu pour les conducteurs d'un transistor de puissance,

—(en bas, à droite) le scellement en verre de redresseur, dont les dimensions se conforment aux normes américaines pour ce genre de pièce détachée.

EE 11 751 *pour plus amples renseignements*

BRITISH BROADCASTING CORPORATION

Broadcasting House, London W.1

GÉNÉRATEUR DE FORMES D'ONDES COMPLEXES DE FOURIER

(*Illustration à la page 431*)

Cet appareil présente les formes d'ondes complexes des sinus et cosinus de n'importe quelle fonction pouvant être représentée par pas plus de quarante et une ordonnées équidistantes, dont l'ordonnée centrale doit représenter la fonction au zéro des abscisses. Le coulisseau de chacun de quarante et un potentiomètres est réglé à la valeur de la fonction de l'ordonnée correspondante. La forme d'onde complexe des sinus ou des cosinus est choisie au moyen d'un commutateur.

Chaque potentiomètre est relié à l'une des vingt et une tensions push-pull sinusoïdales, harmoniquement connexes, et les tensions sur tous les coulisseaux sont combinées de façon à former la forme d'onde complexe de sortie. Chaque fréquence est reliée à deux potentiomètres, disposés symétriquement autour du potentiomètre central, la fréquence augmentant selon d'ordre de position à partir du centre. Pour la forme d'onde complexe des cosinus, les tensions sinusoïdales sont mises en phase de manière à ce qu'elles atteignent simultanément une amplitude maximum une fois par cycle de la fréquence fondamentale, les sorties de tous les coulisseaux de potentiomètres étant ensuite ajoutées. Pour la forme d'onde complexe des sinus, le cadrage est tel que toutes les tensions passent simultanément par le zéro dans une direction positive et que les sorties des coulisseaux du côté gauche sont soustraites de celles de droite.

Les principaux éléments du circuit sont un maître-oscillateur, fonctionnant à 120 fois la fréquence fondamentale de 500 Hz, des chaînes de division de fréquence et des fibres passe-bande. Douze des vingt fréquences de sortie sont filtrées directement par les formes d'ondes de sortie des diviseurs, et ces formes d'ondes sont également intermodulées en paires appropriées afin de produire les huit autres fréquences. L'appareil utilise 159 transistors de jonction au germanium, en majeure partie du type à fréquence acoustique (OC71). Dans le cas de l'oscillateur de 60 kHz, cependant, ainsi que pour l'amplificateur de sortie, un transistor haute fréquence (OC44) est utilisé.

L'amplitude de l'oscillateur est stabilisée. Deux types de diviseurs de fréquence sont prévus: les circuits Eccles-Jordan pour division binaire et le circuit Chaplin-Owens pour la division par trois et par cinq.

EE 11 752 *pour plus amples renseignements*

BURNDEPT LTD

West Street, Erith, Kent

ALIMENTATION STABILISÉE

Le Bloc d'Alimentation BURNDEPT Type BE 253 constitue un élément fort utile, grâce à une excellente régulation de tension et une basse résistance interne, pour le fonctionnement de matériels utilisant des transistors. La tension de sortie est à variation continue, par réglages approximatifs et précis, dans une gamme allant de 1 V à 30 V. Le courant disponible est de 1 A. Tant la tension de sortie que le courant sont contrôlés par un appareil de mesure de 89 mm. La résistance de sortie est de 0,05 ohms, le bruit et les ondulations étant inférieurs à 1 mV de crête à crête. Le coefficient de température du bloc est de 6 mV/°C et il peut être employé dans des températures ambiantes allant jusqu'à 40°C, sur toute la gamme de tensions, et jusqu'à 50°C, sur une gamme limitée.

EE 11 753 *pour plus amples renseignements*

DAWE INSTRUMENTS LTD

99-101 Uxbridge Road, Ealing, London, W.5

OSCILLOSCOPE PORTATIF

(*Illustration à la page 431*)

Cet oscilloscope est entièrement transistorisé. Dans sa forme actuelle, il est principalement prévu pour la mesure et l'étude des formes d'ondes rencontrées dans les champs acoustiques et vibratoires. Il peut également être employé comme oscilloscope universel pour des fréquences d'environ 50 kHz. Il est portatif, étant alimenté par un petit accumulateur logé dans le boîtier. Pour l'utilisation au banc, l'alimentation secteur pourra servir de charge d'entretien à l'accumulateur. Le tube à rayons cathodiques est d'un nouveau type extra sensible. Une batterie de 6 volts alimente le réchauffeur du tube et commande aussi un transformateur de courant continu donnant des sorties de 1,3 kV et de 30 V c.c. au tube très haute tension et à l'alimentation du circuit à transistors respectivement. La base de temps consiste en un multivrateur à fréquence variable dont l'impulsion est utilisée, après amplification, pour charger un condensateur.

Ce dernier se décharge à travers un drain de courant constant. L'amplificateur emploie trois étages et actionne le tube à partir d'un transistor dans la configuration de la base à la masse.

EE 11 754 *pour plus amples renseignements*

ELLIOTT-AUTOMATION LTD

34 Portland Place, London W.1

POTENTIOMÈTRE ÉLECTRONIQUE

(*Illustration à la page 432*)

Le Potentiomètre électronique Série 198, fabriqué par Electroflo Meters, du Groupe Elliott-Automation, est un instrument basé sur la méthode du zéro et à équilibre permanent, conçu pour l'enregistrement de valeurs électriques d'un potentiel réduit, émanant de sources de signaux telles que les thermocouples, extensiomètres, indicateurs de positions d'axes, etc.

Le signal d'entrée est appliqué à un pont de Wheatstone modifié, parcouru par une tension de référence à étalonnage continu. Ceci produit dans le pont un signal déséquilibré dont l'ampleur est en fonction directe du signal d'erreur d'entrée. La sortie du pont est appliquée à un amplificateur à transistors dont le rôle est d'actionner un moteur d'induction diphasé de 50 Hz, de manière à ce que ce dernier entraine le membre mobile d'un fil à contact glissant étalonné dans une position équilibrée, ce qui annule le déséquilibre du pont. Une aiguille fixée au membre mobile du fil à contact glissant indique, sur une échelle de 317,5 mm, la valeur du signal d'erreur extérieur.

Cet instrument peut recevoir des signaux d'entrée en continu qui sont modulés mécaniquement par un relais écrêteur, selon la méthode bien éprouvée de modulation de petits signaux de courant continu, avec un minimum d'interférences de niveau de bruit.

Un dispositif de contrôle supplémentaire est incorporé à la transmission du servomoteur au fil à contact glissant de mesure. En plus des indications données sur le panneau avant, l'instrument est muni d'une commande à trois positions, d'un autre fil à contact glissant et d'un commutateur de relais avertisseur. Il y a également des commandes proportionnelles et de marche-arrêt, avec avertisseurs de haut et de bas niveau pour tous les modèles.

Le Potentiomètre électronique est particulièrement indiqué pour les calculatrices analogiques et arithmétiques, ainsi que pour les téléindicateurs.

EE 11 755 *pour plus amples renseignements*

A. & M. FELL LTD

1 Lambeth High Street, London S.E.1

MESURE PRÉCISE DE RÉSISTIVITÉ

(*Illustration à la page 432*)

Certains appareils d'essai de résistivité ne permettent pas d'y loger les aiguilles de contact avec suffisamment de précision.

Dans l'appareil que produit la Société A. & M. Fell Ltd, les quatre aiguilles sont logées dans des trous de guidage en rubis, exécutés avec le même degré de précision dans la distance entre centres que pour les trous de palier d'une montre. Les aiguilles sont en acier trempé, les pointes étant laissées absolument dures. Ce système permet de répéter la mesure de résistivité avec un haut degré de fréquence.

La distance normale entres centres est de $1/2\pi$ cm, mais d'autres espacements peuvent être exécutés sur commande, et les aiguilles peuvent être disposées soit en ligne droite soit dans les coins d'un carré, la disposition en ligne droite étant la plus courante.

La pression de l'aiguille est de 150 à 200 grammes, et le rayon de courbure des pointes est de 0,1 mm pour le germanium et de 0,04 mm pour le silicium.

La tête est montée sur un mouvement parallèle à rabattement à ressort et un chariot à sommet de verre, ayant une course de 30 cm, est prévu pour le cristal. Une règle d'acier et un index permettent de tracer des transversales le long d'un cristal.

EE 11 756 *pour plus amples renseignements*

FERGUSON RADIO CORPORATION LTD

Great Cambridge Road, Enfield, Middlesex

GÉNÉRATEUR DE FORMES D'ONDES DE TÉLÉVISION

(*Illustration à la page 432*)

Cet équipement portatif, fonctionnant sur pile ou secteur, est prévu pour l'utilisation avec des appareils de prises de vues monochromes de télévision, ainsi qu'avec des mélangeurs vidéo appropriés, afin de produire des formes d'ondes complètes d'images de télévision.

Il produit des impulsions de déclenchement verticales et horizontales pour les bases de temps de la caméra ainsi que des formes d'ondes composées (horizontales et verticales) de synchronisation et de suppression pour les mélangeurs vidéo. L'emploi exclusif de transistors a considérablement contribué à réduire les dimensions, le poids et la consommation électrique de l'équipement, par rapport à ceux d'un appareil équivalent à lampes électroniques.

Un dispositif à circuits non-critiques, allié à des conditions de fonctionnement éprouvées en ce qui concerne les transistors, assurent la parfaite sécurité de marche de l'équipement et évitent un entretien fréquent.

Les formes d'ondes de sortie se conforment pleinement aux Spécifications TV42 de la B.B.C. pour le système à 405 lignes.

L'équipement comprend six plaquettes de circuits imprimés interchangeables, fixées sur un cadre métallique et enfermées dans un boîtier en métal. Le cadre porte également une plaquette rallonge spéciale qui permet de retirer de l'assemblage n'importe laquelle des plaquettes de service et de l'utiliser dans des conditions permettant un accès facile aux deux côtés pour les besoins de réglage et d'entretien.

Le bloc d'alimentation secteur peut être uni par une prise à l'extrémité arrière du cadre avec lequel il fera corps. Ce bloc peut être remplacé par un élément à pile lorsque le fonctionnement à alimentation propre est requis.

Le maître-oscillateur contrôlant le générateur de formes d'ondes peut être asservi à:

(a) un oscillateur à cristal de quartz interne,

(b) un signal de 50 Hz émanant du bloc d'alimentation principal,

(c) un signal de référence extérieur de 50 Hz.

La fermeture externe du secteur est d'un modèle spécial à longue constante de temps, assurant des signaux convenant au fonctionnement avec des télérécepteurs à "volant de synchronisation". Avec une constante de temps équivalente, de l'ordre de 7 secondes, elle pourra être asservie à n'importe quelle fréquence de référence de 47 Hz à 52 Hz, une trentaine de secondes environ après la mise en circuit. Cette caractéristique permet de faire fonctionner le système par la mise en circuit à distance de l'alimentation.

EE 11 757 *pour plus amples renseignements*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London W.C.2

EQUIPEMENT DE TÉLÉCOMMUNICATIONS

Cette Société a présenté de nombreux exemples d'applications de transistors à des éléments utilisés dans des systèmes à fils multivoies découverts et à conducteurs coaxiaux. Ces applications comprennent:

Une panneau de direction qui dirige les fréquences sonores reçues vers la position voulue dans la bande du groupe de fréquences de 60 à 108 kHz. Dans la direction inverse, il opère la démodulation de la parole et permet également des signaux hors de bande, si nécessaire.

Un répétiteur "enterré", destiné à l'emploi dans un système à conducteurs coaxiaux de diamètre réduit, à 240 circuits, utilisant deux amplificateurs à réaction négative, avec les réseaux nécessaires de compensation insérés entre eux. Les amplificateurs sont destinés à compenser la perte de câblage et la tension parcourant chacun d'eux est de l'ordre de 3,5 volts pour l'alimentation le long du câble.

Un élément de signalisation hors de bande a été développé pour recevoir des impulsions de courant continu de −24. −50 volts ou "E" d'un central téléphonique, qu'il transforme en impulsions de 3825 Hz, les "injectant" ensuite au panneau de direction. Le récepteur reçoit 3825 Hz du panneau de direction, les amplifie et les redresse, et actionne un circuit de commutation à transistors qui commande à son tour un relais télégraphique. Les contacts sont disposés de façon à ce que les impulsions requises de courant continu puissent être transmises au central.

Un suramplificateur de groupe destiné aux systèmes porteurs de fréquences modulées en amplitude coaxiale. Ce suramplificateur est du type à réaction négative, avec une réaction totale de 250 décibels à peu près, afin de produire un gain avant de 25 décibels. La réponse de fréquence est supérieure à ±0,25 décibels et les caractéristiques de distorsion sont excellentes. Un suramplificateur démodulateur de groupe peut être fourni pour amplifier le groupe basique de 12 circuits (60 à 108 Hz) sur des systèmes porteurs de fréquences modulées en amplitude. La réaction négative est utilisée pour répondre aux impédances et aux caractéristiques de distorsion. La consommation de courant est de 30 mA à 20 volts.

EE 11 758 *pour plus amples renseignements*

HATFIELD INSTRUMENTS LTD

Crawley Road, Horsham, Sussex

BLOC D'ALIMENTATION C.C. DOUBLE

Quoique principalement prévu pour le champ à transistors, où deux tensions d'alimentation à variation indépendante sont souvent essentielles, le Bloc d'alimentation Type LE 400 constitue un auxiliaire précieux de l'équipement général de laboratoire. Les deux sorties sont identiques et peuvent être variées indépendamment afin de fournir n'importe quelle tension voulue entre 0 et 30 volts C.C. avec un courant maximum de 1 Ampère. Ces blocs permettent de mesurer le courant de chaque sortie et ils assurent la protection automatique et le contrôle régulier des tensions de sortie. Ils se caractérisent également par une impédance de source de moins de 0,10 ohms, un rapport de stabilisation de 200, des ondulations de 1 mV, à pleine sortie. Le transformateur d'entrée du secteur peut recevoir une tension de 200 à 250 volts 50 Hz, et une variation de ±10% du réglage nominal peut être tolérée.

EE 11 759 *pour plus amples renseignements*

AMPLIFICATEUR DE GALVANOMÈTRE

Cet amplificateur se distingue par sa dérive fort réduite sur une gamme de fréquences de 0 à 4 kHz et par une linéarité de sortie supérieure à 2%. L'impédance d'entrée est de 2 MOhms et au réglage de sensibilité maximum de 250 mA/V, l'amplificateur fournira ±25 mA/V dans 75 ohms. Après le temps initial de chauffage, d'une durée de 15 minutes, la dérive, mesurée à sa sensibilité maximum, ne dépasse pas 3 mV. Les commandes d'équilibrage et de l'atténuateur d'entrée sont groupées sur le panneau avant. L'amplificateur a une alimentation intégrale, permettant ainsi de faire fonctionner plusieurs éléments simultanément de manière entièrement indépendante.

EE 11 760 *pour plus amples renseignements*

KYNMORE ENGINEERING CO. LTD

19 Buckingham Street, London, W.C.2

ENREGISTREUR À PLUME

(Illustration à la page 433)

Le "HELCOSCRIPTOR", fabriqué par Fritz Hellige & Cie., Allemagne, et distribué dans le Royaume-Uni par la Société Kynmore Engineering, est un enregistreur à action rapide utilisant une plume chauffée sur du papier recouvert de matière plastique et donnant ainsi un enregistrement en coordonnées rectangulaires instantanément visible. Il est livré en deux modèles, soit comme enregistreur à voie unique pour l'enregistrement direct d'un processus, soit comme enregistreur multivoies pour l'enregistrement direct de trois à quatre processus.

Des amplificateurs a courant continu, à deux étages, sont incorporés à l'enregistreur, le premier étage faisant fonction d'amplificateur différentiel avec une impédance d'entrée de 1 MΩ et une sensibilité de 10 mm/V.

La réponse de fréquence est horizontale de 0 à 100 Hz et de 3 dB jusqu'à 125 Hz.

Le modèle à voie unique fournit deux vitesses de papier et celui à quatre voies en fournit sept. Les vitesses du papier peuvent varier, selon l'engrenage fixé, entre 0,5 mm/sec. et 200 mm/sec. Pour les vitesses extrêmement réduites, il existe un moter interchangeable spécial donnant des vitesses aussi réduites que 0,02 mm/sec.

EE 11 761 *pour plus amples renseignements*

LIVINGSTON LABORATORIES LTD

Retcar Street, London, N.19

ANALYSEUR D'ONDES

L'Analyseur d'ondes Hewlett-Packard, Modèle 302A, dont la distribution dans le Royaume-Uni est assurée par les Laboratoires Livingston, peut diviser un signal d'entrée dans ses composantes de fréquence individuelles, de manière que la composante fondamentale, les harmoniques et les produits d'intermodulation puissent être mesurés et évalués séparément. On peut également l'utiliser comme voltmètre accordé à bande étroite indiquant les niveaux absolus ou relatifs. Il est entièrement muni de transistors et sa gamme de fréquences est de 20 Hz à 50 kHz.

Des mesures précises et sûres sont assurées grâce à une commande de fréquences automatique qui maintient la différence de fréquence entre le signal d'entrée et l'oscillateur local constante à 100 kHz, même s'il y a dérive du signal d'entrée.

L'Analyseur est également pourvu d'un circuit de rétablissement de fréquence, un appoint des plus utiles car il permet d'effectuer facilement des mesures précises de fréquence pour chaque composante de forme d'ondes.

EE 11 762 *pour plus amples renseignements*

JOSEPH LUCAS LTD

Great King Street, Birmingham 19

REDRESSEURS SEMI-CONDUCTEURS

(Illustration à la page 434)

Depuis de nombreuses années, la Société Lucas produit des matériels semi-conducteurs pour son propre usage. Certaines de ces fabrications, dont les suivantes, sont maintenant offertes sur le marché.

Redresseurs de grande puissance: Ces redresseurs ont des régimes de courant avant de 35 A et de 100 A lorsqu'ils sont montés dans des cuves thermiques appropriées. Leurs bonnes caractéristiques avant assurent une perte réduite de puissance, même à ces courants élevés. Ce sont des redresseurs goujonnés pouvant être assemblés dans des configurations en pont. Ils sont livrables pour des tensions nominales inversées de crête allant de 50 à 400 volts et leur gamme de températures va de −55°C à +100°C.

Assemblages redresseurs au silicium: Ce sont des assemblages compacts dans lesquels les jonctions redresseuses font corps avec les ailettes de refroidissement. Leurs dimensions hors-tout sont de 38 mm de diamètre × 47,6 mm de long. Le montage est effectué par un boulon central. On peut fixer des fils aux cosses spéciales soit par soudure soit par des connecteurs à assemblage rapide. Ces assemblages ont des caractéristiques de courant avant de 5 A et des caractéristiques de tension anodique inverse de crête allant de 50 à 400 volts.

Redresseurs de basse et moyenne puissance: Ces redresseurs sont fournis dans une gamme de caractéristiques courantes allant de 0,25 à 10 A pour des tensions anodiques inverses de crête entre 50 et 400 volts et ils sont fabriqués avec le plus grand soin pour l'emploi dans des conditions extrêmes de température, d'humidité et d'effort mécanique. Les modèles pour courants faibles sont à fils, le reste étant à montage goujonné.

EE 11 763 *pour plus amples renseignements*

MINISTRY OF SUPPLY

Trois des services de recherches et de développement du Ministère des Approvisionnements ont présenté des fabrications montrant l'utilisation du transistor, dont, entre autres, les suivantes:

DÉTECTEUR DE PULSATIONS

On peut observer le rythme des pulsations du coeur en contrôlant les potentiels de la peau se rapportant à l'action du coeur. C'est une méthode pratique dans le cas de sujets passifs ou anesthésiés, mais lorsque le sujet se livre à des exercices vigoureux, les signaux du coeur sont normalement assourdis par les signaux se rapportant à l'activité musculaire.

Dans l'instrument présenté, les signaux du coeur produisent des impulsions nettes de sortie, même lorsque le sujet est actif. Les signaux de potentiel de peau sont captés de deux points où les signaux du coeur se ressemblent et où les signaux musculaires sont très différents (comme, par exemple, de muscles se faisant face, tels que ceux de la face intérieure et extérieure de l'avant-bras, en employant la peau au dessus du coeur comme point de masse commun). Les deux signaux sont appliqués, après amplification, à un circuit à porte qui ne produit une impulsion de sortie que lorsqu'il reçoit des signaux coïncidents, de sorte qu'il n'y a pas lieu de tenir compte du brouillage musculaire. Le signal de sortie permet de contrôler un petit émetteur portatif dans le cas où le sujet ne doit pas porter des fils, ou de déclencher un oscillateur afin de produire un signal de sortie audible.

Le modèle original à tubes a été conçu par les services du Ministère des Approvisionnements, tandis que la version à transistors présentée au Salon a été mise au point par les services électriques du Royal Aircraft Establishment de Farnborough, pour les recherches physiologiques.

EE 11 764 *pour plus amples renseignements*

MULLARD LTD

Mullard House, Torrington Place, London W.C.1

NOUVEAUX TRANSISTORS

Le Transistor OC203, qui constitue le modèle le plus récent de la série des transistors à jonction en alliage de silicium, a été prévu pour des applications universelles à basse fréquence, a des tensions collectrices allant jusqu'à −60 V c.c. ou de crête.

Le OC203 se caractérise par une basse tension de saturation de 130 mV (en moyenne) à un courant collecteur de 7 mA, un bas facteur de bruit, d'environ 8 décibels, ainsi qu'un faible courant de perte, de 0,01 μA (moyenne à V_c = −10 V dans une température ambiante de 25°C).

La fréquence de coupure est de 1 MHz et le gain en courant de 15. Le courant collecteur maximum est de 50 mA c.c. ou de crête, la dissipation maximum étant de 100 mW à 100°C. La gamme de températures de jonction va de −50°C à +150°C.

Des quantités limitées de ce modèle sont livrables actuellement, mais il sera bientôt fabriqué en grande série.

Les modèles OC139 et OC140 sont les premiers transistors à jonction NPN introduits par Mullard. Tous deux sont du type à alliage de germanium, étudiés pour les circuits de calculateurs à action rapide ainsi que pour d'autres applications à haute fréquence.

Les nouveaux transistors NPN ont des caractéristiques élevées de courant, d'une moyenne de 200 mA à 250 mA, une basse tension de saturation de 220 mV pour un courant collecteur de 50 mA, un faible courant de perte, de 30 μA au maximum (mesuré à 25°C). La tension collectrice maximum est de +20 V c.c. ou de crête. La température maximum de la jonction est de 75°C, pour le fonctionnement continu, et de 90°C pour le fonctionnement intermittent.

Le modèle OC139 a une fréquence de coupure minimum de 3,5 MHz et un gain en courant minimum à grand signal de 20, pour un courant collecteur de 15 mA. Dans le cas du OC140, ces paramètres sont de 4,5 MHz et de 50 respectivement. La chute de gain aux courants élevés est relativement faible, les chiffres étant respectivement de 14 et 35 pour le OC139 et le OC140 à un courant collecteur de 20 mA.

Les deux modèles ont des jonctions collectrices et émettrices presque symétriques et sont à gain en courant inverse contrôlé.

Les modèles OC28 et OC29 sont des transistors de puissance industriels de 6,0 A, calculés pour supporter une tension collectrice de crête de 60 volts, en transmettant un courant de 0,5 A.

Ces nouveaux types sont prévus pour les applications de contrôle à haute tension et de commutation de courant élevé, le OC28 étant particulièrement indiqué, en outre, pour les transformateurs de courant continu émanant de sources d'alimentation de 24 volts à pleine charge. Ils constituent une amélioration marquée des fabrications précédemment produites à cet usage par cette Société. En plus de leurs caractéristiques de haute tension nettement meilleures, ils ont des chiffres de gain sensiblement supérieurs. Pour un courant collecteur de crête de 6,0 A (pour les deux modèles), le OC28 donne un gain en courant à grand signal minimum de 27, et le OC29 de 45.

Les deux modèles sont prévus pour le fonctionnement continu à une température de jonction allant jusqu'à 90°C (et jusqu'à 100°C pour l'usage intermittent). Ils sont faits selon la technique standard américaine de fabrication de transistors de puissance, également adoptée pour les futurs transistors de puissance Mullard.

Le OC170 est le premier d'une nouvelle gamme de transistors haute fréquence au germanium appliquant la méthode de diffusion d'alliage et il est déjà produit en grande série pour les appareils industriels et ménagers.

Le modèle OC170 a une fréquence de coupure moyenne de 70 MHz et assure un gain en courant de 80. La dissipation maximum admissible au collecteur est de 60 mW dans une température ambiante de 45°C et la tension maximum collecteur est −20V.

Ce modèle sera suivi prochainement par un transistor similaire, prévu pour être utilisé comme mélangeur à des fréquences allant jusqu'à 100 MHz avec un facteur de bruit réduit, et, par la suite, par d'autres modèles encore, dont les fréquences de coupure moyenne atteindront 200 MHz et davantage.

Deux nouveaux transistors haute fréquence en alliage de germanium, spécialement prévus pour les calculateurs arithmétiques ainsi que pour les autres application à commutation rapide, ont été présentés pour la première fois au Salon. Ces nouveaux transistors, types OC41 et OC42, se conforment entièrement aux spécifications des calculatrices et leurs caractéristiques sont rigoureusement contrôlées. Ils permettent des courants de crête élevés, et les temps de coupure et de branchement sont spécifiés. Ils sont destinés à remplacer dans les circuits calculateurs les types OC44 et OC45, actuellement utilisés, et qui furent conçus principalement pour les applications à ondes sinusoïdales.

Les deux nouveaux transistors peuvent transmettre un courant collecteur de crête de 125 mA et leur courant de perte est de 2 μA en moyenne, avec branchement d'émetteur à la masse dans une température ambiante de 25°C. La tension de saturation collecteur est de 120 mV (moyenne) pour un courant collecteur de crête et la température de jonction maximum pour le fonctionnement continu est de 75°C.

EE 11 765 *pour plus amples renseignements*

Résumés des Principaux Articles

Un appareil transformateur d'images pour l'observation de décharges gazeuses auto-éclairées par K. G. Beauchamp et A. J. Tyrrell

Résumé de l'article aux pages 384 à 389

Il s'agit ici d'un équipement de prise de vues à action rapide, utilisant un tube transformateur d'images modulées par impulsions. Les auteurs indiquent des méthodes pour le déclenchement périodique du tube, à des potentiels pouvant atteindre 20 kV, et ils examinent les avantages qu'il y aurait à faire dévier l'image électronique à travers le bulbe du tube.

Les dimensions et la complexité d'un transformateur d'images peuvent être réduites considérablement par l'utilisation d'un aimant permanent pour la concentration du tube. L'appareil dont il est question comprend effectivement un élément de concentration de ce genre.

Cavités d'accord à gamme étendue pour Klystrons réflex interchangeables par R. Mather et J. Sharpe

Résumé de l'article aux pages 390 à 393

Les cavités externes dont traite cet article ont une gamme d'accord de plus de 30% dans la bande de fréquences de 5 kMHz à 12 kMHz et elles sont prévues pour l'emploi avec les klystrons réflex interchangeables. Leurs dimensions sont fort réduites et elles présenteraient certains avantages par rapport aux modèles décrits précédemment. Des résultats d'essais sont également donnés par les auteurs ainsi que des suggestions d'applications.

La résistance à déviation totale: un nouveau concept par R. H. W. Burkett

Résumé de l'article aux pages 393 à 397

La demande croissante d'appareils électroniques d'un fonctionnement sûr pendant de longues périodes de temps a rendu nécessaire l'examen du comportement de résistances chargées pour plusieurs années. Certains types de résistances ont des caractéristiques contrôlées et prévisibles, de sorte qu'il est possible de garantir que ces résistances maintiendront leur tolérance pendant plusieurs années d'usage continu. On peut donc fournir une résistance à déviation totale garantie, comportant un tolérance sélective et prévoyant tous les changements pouvant amener une variation de la résistivité nominale de cette résistance. L'auteur montre que des excursions maxima de 5, 7 et 10% sont réalisables et que l'on peut réduire ces chiffres à des excursions totales de 1, 2 ou 3% pour des besoins spéciaux. Il démontre, en s'appuyant sur certaines données, qu'il est possible de fabriquer des résistances répondant à ces spécifications et suggère, en conclusion, une méthode de codage de couleurs pour ces résistances à déviation totale.

Un calculateur analogique pour prédire la propagation acoustique dans l'atmosphère par F. A. Key et W. G. P. Lamb

Résumé de l'article aux pages 398 à 402

Les auteurs décrivent un calculateur analogique étudié pour permettre de prédire les conditions de propagation acoustique dans l'atmosphère. Ce calculateur enregistre les données concernant les variations de température de l'air et de la vitesse du vent, en fonction de l'altitude, et présente sur un écran cathodique les oscillogrammes indiquant la propagation acoustique dans la direction voulue.

Circuits de transistors pour calculateur arithmétique de 1 MHz par I. Krajewski

Résumé de l'article aux pages 403 à 407

L'auteur examine les conditions requises pour les amplificateurs utilisés pour le rétablissement de formes d'impulsions. Il décrit ensuite des oscillateurs d'arrêt ainsi que des amplificateurs à réaction dotés de transistors et pouvant reformer et régler à nouveau les impulsions d'un calculateur arithmétique ayant un taux numéral de 1 mHz. Il conclut par quelques applications.

Méthodes approximatives pour calculer le comportement de noyaux magnétiques à boucle carrée dans les circuits par D. A. H. Brown

Résumé de l'article aux pages 408 à 411

La durée de commutation dans un noyau magnétique à boucle carrée est en proportion inverse de l'excédent de force magnétomotrice d'entraînement par rapport à la force magnétomotrice coercive. Cette donnée connue est alliée à l'équation de force magnétomotrice pour calculer la durée de commutation et le courant secondaire lorsque le noyau a un bobinage secondaire chargé. L'auteur de l'article démontre que la durée de commutation est la somme de la durée de commutation du circuit ouvert, auquel a été appliqué le courant primaire, et du rapport entre le changement de flux et la tension appliquée. Il indique ensuite deux applications de cette théorie, dont l'une aux noyaux interconnectés et l'autre aux cas où l'entraînement ne provient pas d'une source de courant pur.

Inverseurs de transistors et éléments à filtres de redresseurs par F. Butler

Résumé de l'article aux pages 412 à 418

Après avoir rappelé sommairement la théorie des inverseurs de transistors, l'auteur décrit en détail deux modèles pratiques d'inverseurs fonctionnant sur une alimentation de 12 volts. L'un de ceux-ci consiste en un circuit push-pull symétrique comportant quelques perfectionnements pour améliorer la forme d'onde de sortie, tandis que l'autre utilise quatre transistors dans un branchement en pont, pouvant être adapté pour produire une très grande puissance de sortie.

Quelques instruments inusités de musique électronique par Alan Douglas

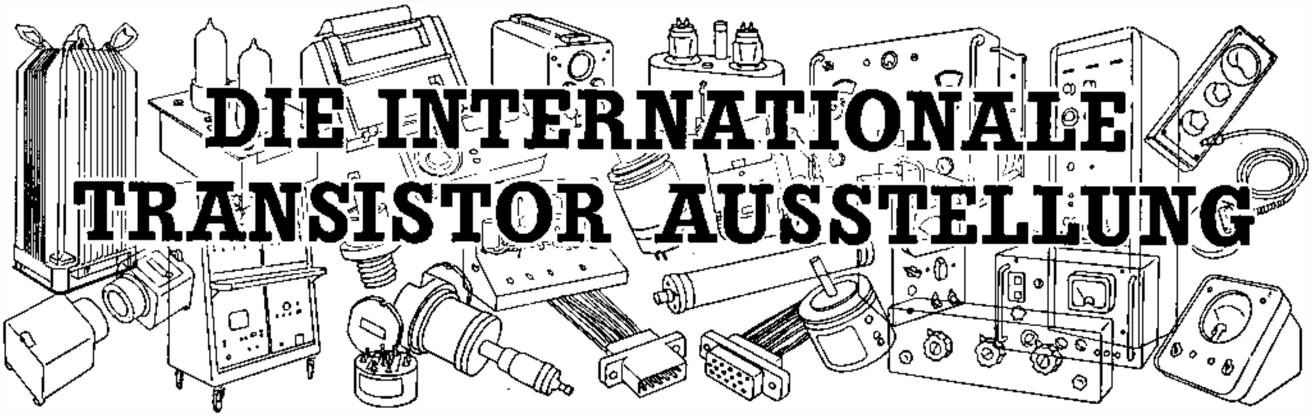
Résumé de l'article aux pages 419 à 421

Les instruments à clavier électronique classique sont fort connus aujourd'hui. Par contre, les instruments pouvant produire ou modifier des sons musicaux par d'autres moyens continuent à occuper l'attention des chercheurs. Cet article décrit certains instruments inusités de ce genre.

Facteurs déterminants dans la fabrication des minuteries industrielles à cathode froide par J. F. Young

Résumé de l'article aux pages 422 à 425

La précision d'une minuterie est déterminée par la stabilité des pièces qui la composent et par les variations de la tension de circuit. On peut dire que des conditions de fabrication optimum ont été remplies lorsqu'un courant de préallumage est nécessaire au tube de déclenchement à cathode froide. Une simple minuterie ne peut être employée que pour le contrôle précis de charges de chauffe dans une gamme limitée de tensions d'alimentation, mais cette gamme peut être sensiblement étendue en ajoutant une fonction à plots à l'onde de minuterie.



DIE INTERNATIONALE TRANSISTOR AUSSTELLUNG

Beschreibung einiger auf der Internationalen Transistor Ausstellung in Earls Court London vom 21.-27.Mai gezeigten Gegenstände, die auf Grund der von Herstellern gemachten Angaben zusammengestellt wurde. *(Ubersetzung der Seiten 431 bis 434)*

BELLING & LEE LTD

Great Cambridge Road, Enfield, Middlesex

GLAS-DURCHFÜHRUNGEN

(Abbildung Seite 431)

Mit Hilfe besonderer Fertigungsmethoden stellt Belling and Lee Ltd jetzt eine Reihe von Druckglas- und angepassten Glas-Durchführungen her, die selbst unter den schwierigsten Bedingungen eine zuverlässige Leistung garantieren und aufrecht erhalten.

Ausgestellt wurden unter anderem eine Durchführung für Einzelanschlüsse L1440, ein Gleichrichter-Sockel L1441, ein Transistor-Sockel für Klasse 30-L1442 und eine Glas-Durchführung L1459.

Typ L 1440 wurde hauptsächlich für die Einzelanschlüsse eines Leistungstransistors entwickelt, findet jedoch auch in gewissen Fällen Verwendung, in denen es sich um die Durchführung einzelner isolierter Drähte in luftdichten Geräten handelt.

Die Abbildung zeigt oben den L 1442 Transistor-Sockel, der den amerikanischen Normen Jetec 30 entspricht und für Transistoren geringer Leistung eingesetzt wird; unten links die Durchführung L 1440, die hauptsächlich für die einzelnen Kontaktstifte von Leistungstransistoren entwickelt wurde; und unten rechts der Sockel L 1441, der den amerikanischen Normen für Gleichrichter-Sockel entspricht.

EE 11751 *für weitere Einzelheiten*

BRITISH BROADCASTING CORPORATION

Broadcasting House, London W.1

FOURIER-TRANSFORMATIONS-GENERATOR

(Abbildung Seite 431)

Dieses Gerät zeigt die Sinus- und Kosinus Transformation jeder Funktion an, die durch nicht mehr als 41 gleichentfernte Ordinaten versinnbildlicht werden kann und deren Mittelordinate die Funktion für die Null-Abszisse darstellen muss. Für jedes der 41 Potentiometer wird der Gleitschieber auf den Wert der Funktion für die entsprechende Ordinate eingestellt. Die Sinus- oder Kosinus Transformationen weiden wahlweise durch einen Schalter eingestellt.

Jedes Potentiometer wird mit einer der zwanzig Gegentakt-Spannungen verbunden, die miteinander in harmonischer Beziehung stehens und die Schieberspannungen werden zusammengefasst, um die Transformation zu bilden. Jede Frequenz wird an zwei Potentiometer angelegt, die dem Mittelpotentiometer gegenüber symmetrisch liegen, wobei die Frequenz mit der Entfernung von der Mitte ansteigt. Für die Kosinus Transformation sind die sinusförmigen Spannungen im Gleichtakt und erreichen für jede Periode der Grundfrequenz gleichzeitig einmal einen Amplitudenhöchstwert. Die Ausgangswerte aller Potentiometerschieber werden addiert. Für die Sinus Transformation gehen alle Spannungen in positiver Richtung durch den Nullpunkt. Die Ausgangswerte der linken Schieber werden von denen der rechten Schieber abgezogen.

Die hauptsächlichen Schaltkreis-Bausteine sind ein Steuergenerator, dessen Frequenz 12Ox höher liegt als die Grundfrequenz von 500 Hz, Frequenzteiler und Bandpassfilter. Zwölf der zwanzig Ausgangsfrequenzen kommen direkt durch Filter von den Ausgangswellenformen der Frequenzteiler. Geeignete Paare dieser Wellenformen werden gegenseitig moduliert und ergeben die restlichen acht Frequenzen. Das Gerät ist mit 159 Germanium-Flächentransistoren bestückt, überwiegend NF Transistoren OC 71. Die HF Type OC 44 wird für den 60 kHz Oszillator und für den Ausgangsverstärker eingesetzt, der keine Phasenverzerrung hervorrufen darf.

Der Oszillator ist amplitudenstabilisiert. Zwei verschiedene Frequenzteiler-Schaltungen werden verwandt: für binare Teilung die Eccles-Jordan Schaltung und für Teilung durch drei und fünf die Chaplin-Owens Schaltungen.

Jedes Bandpassfilter liegt entweder an einem Frequenzteilerausgang oder an einem mit vier Transistoren bestückten, symmetrischen Modulator. Die Sekundärseite des Filters ist symmetrisch und steuert zwei Kollektor-Ausgangsstufen, deren Emitterbelastung durch die beiden Hälften des in der Mitte angezapften Ausgangspotentiometers gebildet wird.

Die Filter sind so konstruiert, dass sie den relativen Phasenfehler verringern, der durch Oszillator-Instabilität und thermische Schwankungen in Bauelementen hervorgerufen wird. Die erzielte Phasenstabilität ist unter normalen Bedingungen ausreichend, um 1% Genauigkeit für mindestens drei Monate nach der Eichung zu garantieren.

EE 11752 *für weitere Einzelheiten*

BURNDEPT LTD

West Street, Erith, Kent

STABILISIERTES NETZGERÄT

Das Burndept Netzgerät BE 253 ist eine für Betrieb und Entwicklung von Transistor-Geräten geeignete Spannungsquelle mit guter Regelung und geringem Innenwiderstand. Die Ausgangsspannung ist mit Grob- und Feineinstellung über den Bereich von 1 V bis 30 V bei einer Stromabgabe von 1 A kontinuierlich regelbar. Ausgangsspannung und -strom werden durch ein 89 mm Messgerät überwacht. Der Ausgangswiderstand ist 0,05Ω; der doppelte Spitzenwert für Geräusch und Restwelligkeit ist 6mV/°C. Bei einer Umgebungstemperatur von 40°C kann das Gerät über den vollen, bei 50°C über einen begrenzten Spannungsbereich benutzt werden.

EE 11753 *für weitere Einzelheiten*

DAWE INSTRUMENTS LTD

99-101 Uxbridge Road, Ealing, London W.5

TRAGBARER OSZILLOGRAF

(Abbildung Seite 431)

Dieser Oszillograf ist durchgehend mit Transistoren bestückt. In der vorliegenden Form ist er hauptsächlich zum Messen und zur Untersuchung von Wellenformen gedacht, die im Vibrations- und NF-Gebiet auftreten. Er kann aber auch als Universal-Oszillograf für Frequenzen bis ungefähr 50 kHz eingesetzt werden. Er ist tragbar und wird von einem eingebauten 6 V Akku gespeist. Wenn der Oszillograf am Messplatz gebraucht wird, kann er vom Netz während des Gebrauches aufgeladen werden. Eine neue, sehr empfindliche Kathodenstrahlröhre wird benutzt; eine 6 V Batterie liefert die Heizspannung für die Röhre und speist auch einen Gleichstromwandler mit einer Gleichstromabgabe von 1,3 kV und 30 V für die Röhrenhochspannung bezw. die Transistorspannungen. Für das Ablenkgerät liefert ein Multivibrator einen Impuls variabler Frequenz, der nach geeigneter Verstärkung einen Kondensator aufladet.

Der Kondensator wird mit gleichbleibender Stromstärke entladen. Ein Transistor in Basisschaltung steuert die Röhre über einen dreistufigen Verstärker.

EE 11 754 *für weitere Einzelheiten*

ELLIOTT-AUTOMATION LTD

34 Portland Place, London W.1

ELEKTRONISCHES POTENTIOMETER

(Abbildung Seite 432)

Ein elektronisches Potentiometer mit kontinuierlicher Nullpunktabstimmung wird unter Typennummer 198 von Electroflo Meters der Elliott-Automation Gruppe zur Grössenanzeige kleiner Spannungsunterschiede hergestellt, die z.B. in Thermoelementen, Dehnungsmessern, Drehwinkelanzeigern u.s.w. auftreten.

Das Eingangssignal wird an eine abgewandelte Wheatstone Brücke angelegt, die von einer kontinuierlich stabilisierten Bezugsspannung gespeist wird. In der Brücke wird dadurch ein Unbalanzsignal hervorgerufen, dessen Grösse zur Grösse des Eingangsfehlersignales in direkter Beziehung steht. Die Brückenausgangsspannung geht über einen Transistorverstärker zu einem zweiphasigen 50 Hz Induktionsmotor, der einen Schieber in einer solchen Richtung über einen geeichten Schleifdraht bewegt, dass der Brückennullpunkt wieder hergestellt und die Unbalanz ausgeglichen wird. Mit dem beweglichen Schleifdrahtschieber ist ein Zeiger verbunden, der über eine 317,5mm lange Skala läuft, auf der er die Grösse des angelegten Fehlersignales anzeigt. Das Gerät wurde für Gleichstrom-Eingangssignale entwickelt, die mit Hilfe eines Zerhackerrelais mechanisch moduliert sind, da diese Methode sich zur Modulation kleiner Gleichstromsignale gut bewährt hat und die geringsten Störgeräuschpegel hervorruft.

Zusätzliche Reglermöglichkeiten sind in der Übertragung zwischen Stellmotor und Schleifdrahtschieber eingebaut. Ausser der Anzeige auf der Frontplatte ist eine dreistellige Reglerbetätigung, sowie ein zusätzlicher Schleifdraht für Signalweitergabe und ein Warnschalter vorgesehen. Andere Steuermethoden sehen Proportional- und „Ein- und Aus" Steuerung vor. Alle Modelle haben Warnanlagen für Minimum- und Maximumwerte.

Das Gerät kann für Rechner, Digitalanzeige und Fernanzeige eingesetzt werden.

EE 11 755 *für weitere Einzelheiten*

A. & M. FELL LTD

1 Lambeth High Street, London S.E.1

FEINMESSUNG DES SPEZIFISCHEN WIDERSTANDES

(Abbildung Seite 432)

In einigen Geräten zur Messung des spezifischen Widerstandes haben die Kontaktspitzen nicht genügend Führung.

In diesem Gerät bewegen sich vier Spitzen in Rubinführungsbuchsen, deren Mittellinienabstand mit derselben Genauigkeit eingehalten wird wie der für Uhrenlager. Die Kontaktspitzen bestehen aus gehärtetem und angelassenem Stahl mit glashartem Radius. Durch diese Konstruktion können Messungen des spezifischen Widerstandes mit sehr guter Übereinstimmung wiederholt werden.

Der Mittellinienabstand der Normalausführung ist $1/2\pi$ cm, und die Kontakte sind in einer Reihe angeordnet. Andere Mittellinienabstände können auf Bestellung geliefert werden. Eine Kontaktanordnung, in der die Spitzen die Ecken eines Quadrates bilden, ist auch lieferbar.

Der Spitzendruck liegt zwischen 150 und 200 Gramm. Der Radius der Spitze für Germanium ist 0,1 mm und für Silizium 0,04 mm.

Der Messkopf sitzt auf einem mit Federscharnier versehenen Parallelgang, und ein Schlitten mit Glasoberfläche und 30 cm Durchgang ist für das Kristall vorgesehen. Ein Stahllineal mit Anzeiger ermöglicht es, Durchgänge graphisch aufzutragen.

EE 11 756 *für weitere Einzelheiten*

FERGUSON RADIO CORPORATION LTD

Great Cambridge Road, Enfield, Middlesex

FERNSEH-IMPULSGEBER

(Abbildung Seite 432)

Das Gerät ist als tragbarer Geber mit Netzanschluss oder Batteriebetrieb gestaltet und ergibt bei Einsatz mit Schwarz-Weiss Fernsehkameras und geeigneten Video - Mischverstärkern vollständige Fernseh-Bildwellenformen.

Der Geber liefert die für den Ablenkteil der Kamera erforderlichen horizontalen und vertikalen Auslösungsimpulse, sowie die komplexen (horizontalen und vertikalen) Synchron- und Austastimpulse für das Video-Mischgerät. Durch ausschliessliche Verwendung von Transistoren wurden wesentliche Ersparnisse in Abmessungen, Gewicht und Leistungsaufnahme gegenüber gleichartigen Röhrengeräten erzielt.

Die Anwendung nichtkritischer Schaltungen und eine konservative Wahl der Betriebsdaten tragen zu hoher Betriebssicherheit und Vermeidung häufiger Wartung bei.

Die Ausgangsimpulse entsprechen der BBC Norm TV 42 für 405 Zeilen.

Der Geber enthält sechs gedruckte Schaltplatten, die in ein Metallgestell innerhalb eines Stahlblechgehäuses eingesteckt werden. Eine im Gestell vorgesehene besondere Anschluss-Schaltplatte macht es möglich, jede Betriebsschaltplatte herauszuziehen und notwendige Nachstellungen oder erforderliche Wartung unter Betriebsbedingungen doch von beiden Seiten leicht zugänglich, vorzunehmen.

Das Netzgerät wird von der Rückseite in das Gestell eingeschoben und dadurch in das Gerät eingebaut. Es kann jedoch gegen Batterien ausgewechselt werden, wenn Batteriebetrieb wünschenswert erscheint.

Der den Impulsgeber kontrollierende Steueroszillator kann folgendermassen gekoppelt werden:

(a) Mit einem eingebauten Quartzoszillator.

(b) Mit einem 50 Hz Signal vom Netz.

(c) Mit einem 50 Hz Fremdsignal.

Die Fremdverkopplung des Netzes hat eine besonders lange Zeitkonstante, und die abgegebenen Signale sind für Betrieb mit „Schwungrad-Synchron" Fernsehempfängern geeignet. Mit einer Zeitkonstanten von ungefähr 7 sec wird sich das Gerät innerhalb 30 sec nach Einschalten mit jeder Bezugsfrequenz zwischen 47 Hz und 52 Hz verriegeln. Durch diese Eigenart kann das Gerät durch Fernschaltung des Netzanschlusses eingeschaltet werden.

Für die Schaltkreise des Impulsgebers werden 75 pnp Transistoren mit lo MHz Grenzfrequenz (OC 44) und 14 npn Transistoren mit 6 MHz Grenzfrequenz (Sylvania 2N94A) benötigt.

Acht pnp Transistoren und eine Zener Diode bestücken das Netzgerät, das eine stabilisierte Spannung von 12 V bis zu 200 mA abgibt. Der Netzteil kann sich ohne Nachstellen Netzschwankungen von 260 V bis 150 V anpassen.

EE 11 757 *für weitere Einzelheiten*

THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London W.C.2

GERÄTE FÜR DAS FERNMELDEWESEN

Die folgenden befanden sich unter den ausgestellten transistorbestückten Geräten für Mehrkanalbetrieb mit Freileitungen und Koaxialkabeln:

Eine Kanal-Schalttafel, die ankommende Sprachfrequenzen in den geeigneten Kanal im Gruppenfrequenzband 60 - 180 kHz umsetzt. In umgekehrter Richtung wird das Signal entmoduliert und die Sprache weitergeleitet. Signalübertragung ausserhalb des Bandes ist möglich.

Ein Erdleistungsverstärker für Einsatz mit einem Koaxialkabel für 240 Schaltkreise arbeitet mit zwie gegengekuppelten Verstärkern, zwischen die eine Entzerrungskette geschaltet ist. Die Verstärker sind so konstruiert, dass sie die Kabelverluste ausgleichen. Die am Verstärker liegende Spannung zur Weiterführung der Leistung ist ungefähr 3,5 V.

Ein ausserhalb des Bandes arbeitendes Gerät spricht auf vom Fernsprechamt empfangene Gleichstromimpulse von -24V, -50V oder E an und wandelt sie in Impulse von 3825 Hz um, verstärkt sie und leitet sie an eine normale Kanal-Schalttafel weiter, um einen Transistorschalter zu betätigen, der seinerseits ein telegraphisches Relais steuert. Die Kontakte sind so angeordnet, dass die gewünschten Gleichstromimpulse an das Fernsprechamt weitergeleitet werden.

Ein Übergruppen-Verstärker für koaxiale, amplitudenmodulierte Trägerfrequenzsysteme wurde gezeigt. Ungefähr 25 dB negative Gegenkupplung wird benötigt, um eine Gesamtverstärkung von 25 dB zu erreichen und dabei den Frequenzgang innerhalb ± 0,25 dB zu halten. Der Klirrfaktor ist sehr gut. Ein Übergruppen-Entmodulationsver stärker ist für die Verstärkung der 12 Grundschaltungsgruppen (60-108 kHz) eines amplitudenmodulierten Mehrfach-Trägerfrequenznetzes lieferbar. Negative Gegenkupplung ist erforderlich, um die gewünschten Scheinwiderstand- und Klirrfaktor-Kennlinien zu erzielen. Der Stromverbrauch ist 30 mA bei 20 V.

Für Versuche mit einem elektronischen Fernsprechamt werden Teilnehmeranschlüsse und Schaltstellen mit Impulsmodulatoren und -entmodulatoren benötigt. Die ersteren bestehen aus Zweidraht/Vierdrahtgabeln, in denen Sprache (Zweidraht) in amplitudenmodulierte Impulse (Vierdraht) umgewandelt wird, oder umgekehrt. Sprachsignale werden über einen Zwitter an einen einfachen Halbleiter-Modulator angelegt, dem 0,5µs Impulse zugeleitet werden. Das zweite Gerät ist ähnlich, hat aber keinen Zwitter und arbeitet nach dem Vierdraht/Vierdraht Prinzip.

EE 11 758 *für weitere Einzelheiten*

HATFIELD INSTRUMENTS LTD

Crawley Road, Horsham, Sussex

ZWILLINGS-NETZGERÄT

Das Netzgerät LE 400 ist hauptsächlich für Anwendung auf dem Transistorgebiet bestimmt, wo oft zwei unabhängig von einander einstellbare Spannungen benötigt werden, wird aber auch als nützliches Gerät für allgemeine Arbeiten im Labor dienen. Beide Ausgänge sind identisch und können unabhängig von einander bei 1A Höchststrom kontinuierlich über den Gleichstrombereich von 0 bis 30 V eingestellt werden. Stromanzeige für jeden Ausgang, automatischer Überlastungsschutz und glatte Einregelung sind Merkmale dieses Gerätes. Ein Scheinwiderstand von weniger als 0,01Ω, ein Regelfaktor von 200 und 1mV Restwelligkeit bei Höchstleistung sind weitere Kennzeichen. Der Eingangsumspanner ist für Netzspannungen zwischen 200 und 250V, 50Hz angezapft. Abweichungen von ± 10% vom Sollwert der Netzspannung sind zulässig.

EE 11 759 *für weitere Einzelheiten*

GALVANOMETER-VERSTÄRKER

Dieser Verstärker hat sehr geringe Drift, einen Frequenzbereich von 0 - 4 kHz und eine Ausgangsgeradlinigkeit besse als 2%. Der Eingangsscheinwiderstand ist 2mΩ und der Verstärker liefert ± 25mA an 75Ω, wenn der höchste Empfindlichkeitsbereich (250mA/V) eingestellt ist. Nach einer Anlaufzeit von 15 Minuten ist die Drift bei Höchstempfindlichkeit unter 3mV. Bedienungsknöpfe für Ausgleich und Eingangsdämpfung befinden sich auf der Frontplatte. Der Verstärker hat ein eingebautes Netzgerät, so dass mehrere Geräte vollständig unabhängig von einander zusammen arbeiten können. Da die Verstärker üblicherweise in Mehrfachanordungen eingesezt werden, wurde eine Anlage mit genormter 48,26 cm (19 Zoll) Frontplatte entwickelt, die acht als Einschub-Bausteine gestaltete Verstärker aufnehmen kann.

EE 11 760 *für weitere Einzelheiten*

KYNMORE ENGINEERING CO. LTD

19 Buckingham Street, London W.C.2

SCHNELLSCHREIBER

(Abbildung Seite 433)

Der in Deutschland von Fritz Hellige und Co hergestellte „Helcoscriptor" wird durch Kynmore Engineering vertrieben. Es ist ein Schnellschreiber, der mit einem angewärmten Schreibstift auf kunststoffbeschichtetem Papier eine sichtbare Aufzeichnung in rechtwinkligen Koordinaten herstellt. Zwei Modelle sind lieferbar: ein Einkanalschreiber zur direkten Aufzeichnung eines Vorganges und ein Mehrkanalmodell zur Aufzeichnung bis zu vier Vorgängen.

Die erste Stufe eines zweistufigen Gleichstromverstärkers arbeitet als Differenzialverstärker mit einem Eingangsscheinwiderstand von 1MΩ und 10 mm/V Empfindlichkeit.

Der Frequenzgang ist flach von 0 bis 100Hz und -3dB bei 125Hz.

Zwei Papiergeschwindigkeiten sind für das Einkanalmodell vorgesehen und sieben für das Vierkanalmodell. Papiergeschwindigkeiten sind von den eingebauten Getrieben abhängig und können zwischen 0,5 mm/s und 200 mm/s liegen. Besonders geringe Geschwindigkeiten können mit Hilfe eines Einschubmotors erreicht werden und können bis zu 0,02 mm/s heruntergehen.

EE 11 761 *für weitere Einzelheiten*

LIVINGSTON LABORATORIES LTD

Retcar Street, London N.19

WELLENFORMANALYSATOR

Der in England durch Livingston Laboratories vertriebene Hewlett-Packard Wellenformanalysator 302 A zerlegt ein Eingangssignal in die einzelnen Frequenzkomponenten, so dass Grundfrequenz, Harmonische und Intermodulationsprodukte getrennt gemessen und bewertet werden können. Er kann auch als engbandabgestimmtes Voltmeter eingesetzt werden, das absolute oder relative Pegel messen kann. Das Gerät ist durchgehend mit Transistoren bestückt und hat einen Frequenzbereich von 20Hz bis 50kHz. Es hat eine automatische Frequenzregelung, die die Frequenzdifferenz zwischen dem Eingangssignal und dem eingebauten Oszillator bei 100kHz konstant hält, selbst wenn das Eingangssignal driftet. Dadurch werden genaue und zuverlässige Messungen gewährleistet.

Nützlich ist auch die Frequenzwiedergabeschaltung, die genaue Frequenzbestimmungen für jede Wellenformkomponente leicht macht.

Die Frequenzwiedergabe liefert ein sinusförmiges Ausgangssignal der Frequenz der gemessenen Komponente, und dieses Signal kann für grösste Genauigkeit durch einen elektronischen Zähler gemessen werden. Die Amplitude des wiedergegebenen Frequenzsignals hängt vom Pegel der gewählten Komponente ab und wird abgegeben, wenn das Messgerät anzeigt, wobei es keine Rolle spielt, ob das Gerät mit oder ohne automatische Frequenzregelung arbeitet.

EE 11 762 *für weitere Einzelheiten*

JOSEPH LUCAS LTD

Great King Street, Birmingham 19

HALBLEITER-GLEICHRICHTER

Schon seit einigen Jahren hat Lucas Halbleiterelemente für den eigenen Verbrauch hergestellt. Die folgenden und andere Elemente werden jetzt für den Vertrieb freigegeben.

Halbleiter-Gleichrichter

Der Durchlasstrom beträgt für die Gleichrichter bei guter Wärmeabfuhr 35 und 100 mA. Ihre guten Durchlasskenndaten rufen trotz dieser hohen Stromstärken nur kleine Leistungsverluste hervor. Die Elemente sind für Stehbolzenmontage ausgebildet. Um den Zusammenbau von Brückenschaltungen zu erleichtern, können die Gleichrichter mit Stehbolzen jeder Polarität geliefert werden. Sie werden für Spitzensperrspannungen zwischen 50 und 400 V und Temperaturbereiche von −55°C und +100°C hergestellt.

Silizium-Gleichrichter-Bausteine

(Abbildung Seite 434)

Die gleichrichtenden Flächen sind als Teile der Kühlrippen ausgebildet, aus denen die Bausteine zusammengestellt sind. Sie haben nur 38 mm Durchmesser und sind 47,6 mm lang. Der Einbau wird mittels eines Stehbolzens vorgenommen. Anschlussdrähte werden entweder an die vorgesehenen Fahnen gelötet oder geklemmt.

Diese Bausteine werden für 5A Durchlasstrom und Spitzensperrspannungen zwischen 50 und 400V geliefert.

Gleichrichter für kleinere und mittlere Leistungen

Diese Elemente werden für Strombereiche von 0,25 bis 10A mit Spitzensperrspannungen zwischen 50 und 400V geliefert. Ihre sorgfältige Konstruktion erlaubt Verwendung bei äusserst grossen Temperaturunterschieden, Feuchtigkeit und mechanischer Belastung. Typen für geringe Stromabgabe haben Drahtenden, die anderen Stehbolzen.

EE 11 763 *für weitere Einzelheiten*

MINISTRY OF SUPPLY

Drei Forschungsanstalten des Ministeriums für Waffenbeschaffung stellten transistorbestückte Geräte aus. Darunter befand sich der

HERZSCHLAGANZEIGER

Der Puls kann durch Überwachung der mit der Herztätigkeit zusammenhängenden Hautspannungen beobachtet werden. Diese Methode ist für passive oder betäubte Patienten praktisch anwendbar. Wenn jedoch der Patient kräftige Bewegungen macht, werden die Herzsignale normalerweise durch Störsignale, die mit der Muskeltätigkeit zusammenhängen, unkenntlich gemacht.

In dem ausgestellten Gerät rufen die Herzsignale eines aktiven Patienten klare Ausgangsimpulse hervor. Hautspannungssignale werden dem Gerät von zwei Flächen zugeleitet, die dieselben Herzsignale, jedoch verschiedene Muskelgeräusche abgeben (z.B. über zwei gegeneinander arbeitenden Muskeln an der Innen-und Aussenseite des Vorarms, wobei die Haut über dem Herzen als gemeinsame Erde benutzt wird). Nach Verstärkung werden beide Signale einer Torschaltung zugeleitet, die nur bei Eingabe übereinstimmender Signale einen Ausgangsimpuls abgibt, so dass die Muskelstörungen nicht beachtet werden. Das Ausgangssignal kann dazu benutzt werden, einen tragbaren Sender zu triggern, falls der Patient nicht durch Zuleitungen behindert werden soll, oder es kann einen Oszillator auslösen, der ein Tonsignal abgibt. Das ursprüngliche Röhrenmodell wurde von der Versuchsstelle für Kleidung und Vorräte des Ministeriums entwickelt und die gezeigte Transistorausführung von der elektrischen Abteilung der Königlichen Luftwaffen Anstalt in Farnborough für physiologische Versuchsarbeiten entwickelt.

EE 11764 *für weitere Einzelheiten*

MULLARD LTD

Mullard House, Torrington Place, London W.C.1

NEUE TRANSISTOREN

(Abbildung Seite 434)

Der OC 203 ist ein neuer legierter Silizium-Flächentransistor in der Mullard-Reihe, der für allgemeine NF Verwendung mit Kollektor-Spitzenspannungen von −60V bestimmt ist.

Merkmale des OC 203 sind die geringe Restspannung mit durchschnittlich 130 mV bei 7 mA Kollektorstrom, ein bei 8 dB liegender Rauschfaktor und ein niedriger Reststrom mit einem Mittelwert von 0,01 μA bei U_c = -10V und 25°C Umgebungstemperatur.

Eine typische Grenzfrequenz ist 1 MHz und die Stromverstärkung 15. Der Kollektorspitzenstrom ist 50 mA und die höchste Verlustleistung 100 mW bei 100°C. Der Kristalltemperaturbereich liegt zwischen −50°C und +150°C.

Geringe Mengen dieser Type sind jetzt lieferbar, Massenfertigung wird jedoch in Kürze beginnen.

OC 139 und OC 140 sind die ersten von Mullard eingeführten npn Transistoren. Beide sind Germanium-Flächentransistoren für Rechner mit hoher Schaltgeschwindigkeit oder andere HF Zwecke.

Die neuen npn Transistoren sind für hohe Spitzenströme von durchschnittlich 250 mA bemessen, haben eine Höchstrestspannung von 220 mV bei 50 mA Kollektorstrom und einen Höchstreststrom von 3 μA bei 250°C. Kollektorspitzenspannung ist + 20 V, und die Höchstkristalltemperatur ist 75°C für Dauerbetrieb und 95°C für periodischen Betrieb.

Die Grenzfrequenz für den OC 139 liegt über 3,5 MHz, und die Stromverstärkung ist 20 für grosse Signale bei einem Kollektorstrom von 15mA. Für den OC 140 sind diese Parameter 4,5 MHz und 50. Der Verstärkungsabfall für hohe Stromstärken ist verhältnismässig gering und entsprechende Werte sind 14 bzw. 35 für den OC 139 und OC 140 bei 20 mA Kollektorstrom.

Beide Typen haben nahezu symmetrische Emitter- und Kollektorflächen und werden auf Rückwirkungsstromverstärkung kontrolliert.

OC 28 und OC 29 sind 6,0 A Leistungstransistoren für die Industrie, die bei einem Strom von 0,5 A eine Kollektorspitzenspannung von 60 V vertragen.

Diese neuen Typen sind Schalt- und Reglertransistoren für hohe Spannung und hohe Ströme. Darüber hinaus ist der OC 28 besonders für Gleichspannungswandler geeignet, die von einer voll geladenen 24 V Stromquelle gespeist werden.

Sie weisen gegenüber den bisher von Mullard gelieferten Typen wesentliche Vorteile auf und haben nicht nur eine bessere Leistung bei hohen Spannungen, sondern auch eine bedeutend höhere Verstärkung. Für 6,0A Kollektorspitzenstrom ist die OC 28 Stromverstärkung für grosse Signale mindestens 27 und 45 für den OC 29.

Beide Typen sind für Dauerbetrieb bis zu 90°C Kristalltemperatur (100°C für periodischen Betrieb) bemessen. Die Ausführung der Typen entspricht der amerikanischen Norm für Leistungstransistoren, die für alle zukünftigen Mullard Leistungstransistoren Anwendung finden wird.

Der OC 170 wird als erster einer neuen Mullard Germanium-HF-Transistorenreihe nach Diffusionslegierungsmethoden in Massenproduktion für industrielle und Haushaltsgeräte hergestellt.

Der Mittelwert der Grenzfrequenz für den OC 170 ist 70 MHz bei einer Stromverstärkung von 80. Bei einer Umgebungstemperatur von 45°C und −20V Kollektorspitzenspannung ist die Kollektor-Höchstverlustleistung 60 mW.

Eine weitere, ähnliche Type mit geringem Rauschfaktor wird für den Einsatz als Mixer bis zu 100 MHz folgen. Danach sind andere Transistoren mit einer Grenzfrequenz bis zu 200 MHz und höher vorgesehen.

Zwei neue Germanium- HF-Legiertransistoren wurden besonders für Digitalrechner und Schnellschaltzwecke eingeführt.

Die neuen Typen OC 41 und OC 42 erfüllen alle für Rechner aufgestellten Bedingungen, und die Einhaltung ihrer Kenndaten wird in der Herstellung genau überwacht. Hohe Spitzenströme sind zulässig; Ein- und Abschaltzeiten sind angegeben. OC 41 und OC 42 sind Nachfolger der im Gebrauch befindlichen OC 44 und OC 45 für Rechnerschaltungen, da die älteren Typen für Sinuswellenformen entwickelt wurden.

Beide neuen Transistoren können Kollektorspitzenströme von 125 mA aufnehmen und haben in Emitterschaltung einen Reststrom von 2 μA (Mittelwert) bei 25°C Umgebungstemperatur. Bei Kollektorspitzenstrom ist die Kollektorspannung 120 mV (Mittelwert). Die Kristallhöchsttemperatur für Dauerbetrieb ist 75°C.

EE 11765 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Bildwandlergerät zur Beobachtung selbstleuchtender gasförmiger Entladungen von K. G. Beauchamp und A. J. Tyrell

Zusammenfassung des Beitrages auf Seite 384-389

Der Beitrag beschreibt die Verwendung einer impulsgesteuerten Bildwandlerröhre in einem Zeitdehner. Einzelheiten über die Methoden zur Austastung dieser Röhren bis zu 20kV werden gegeben und die Vorteile der Ablenkung des elektronischen Bildes über den Röhrenschirm behandelt.

Die Verwendung eines Dauermagneten zum Fokussieren der Röhre ermöglicht es, das Bildwandlergerät unkomplizierter zu gestalten und die Grösse wesentlich zu reduzieren. Das beschriebene Gerät enthält solch eine Fokussieranordnung.

Breitband Resonanzhohlräume für Einsteck-Reflexklystrons von R. Mather und J. Sharpe

Zusammenfassung des Beitrages auf Seite 390-393

Der Beitrag beschreibt eine Reihe aussenliegender Hohlräume, die im Einsatz mit Einsteck-Klystrons für Band X einen Abstimmbereich von über 30% im Frequenzband 5 GHz bis 12 GHz haben. Die Hohlräume haben verhältnismässig geringe Abmessungen und scheinen gegenüber früher beschriebenen Typen Vorteile aufzuweisen. Prüfresultate werden angeführt und Vorschläge für Verwendung gemacht.

Widerstände mit „Gesamtabweichung" — Ein neuer Begriff von R. H. W. Burkett

Zusammenfassung des Beitrages auf Seite 393-397

Der wachsende Bedarf für elektronische Geräte, die für lange Zeit betriebssicher arbeiten, hat zu einer Überprüfung des Verhaltens von Widerständen geführt, deren Belastung sich über mehrere Jahre erstreckt. Gewisse Schichtwiderstände haben kontrollierte und voraussagbare Eigenschaften, so dass die Toleranzeinhaltung dieser Widerstände für mehrjährigen Dauerbetrieb garantiert werden kann. Solche Widerstände könnten mit einer garantierten „Gesamtabweichung" (total excursion) geliefert werden, die die Auswahltoleranz und alle Änderungen umfasst, die eine Abweichung vom Widerstandnennwert hervorrufen können. Es wird gezeigt, dass Höchstabweichungen von 5%, 7% und 10% praktisch möglich sind, und dass für Sonderzwecke der Grundbegriff auf Gesamtabweichungen von 1%, 2% und 3% erweitert werden kann. Gemachte Angaben zeigen, dass die Entwicklung von Widerständen möglich ist, die diesen Bedingungen entsprechen. Vorschläge zur Farbkennzeichnung von Widerständen mit „Gesamtabweichung" werden gemacht.

Ein Analogrechner zur Voraussage der Schallfortpflanzung in der Atmosphäre von F. A. Key und W. G. F. Lamb

Zusammenfassung des Beitrages auf Seite 398 402

Der beschriebene Analogrechner wurde entwickelt, um Bedingungen für die Schallfortpflanzung in der Atmosphäre voraussagen zu können. Daten für Lufttemperatur- und Windrichtungsschwankungen werden in den Rechner eingegeben, und die Schallfortpflanzung in jeder Richtung wird durch Strahlenmuster auf dem Bildschirm einer Kathodenstrahlröhre dargestellt.

Transistor-Schaltungen für einen 1 MHz Digitalrechner von I. Krajewski

Zusammenfassung des Beitrages auf Seite 403-407

Erfordernisse für Pulsrückbildungsverstärker werden besprochen. Sperrschwinger und Rückkopplungsverstärker mit Transistorenbestückung werden beschrieben, die für Rückbildung und Korrektur der Laufzeit des Impulses in einem Digitalrechner mit einer Digitgeschwindigkeit von 1 MHz geeignet sind. Anwendungsbeispiele werden angeführt.

Annäherungsmethode für die Berechnung des Verhaltens magnetischer Kerne mit rechteckiger Hystereseschleife in Schaltungen von D. A. H. Brown

Zusammenfassung des Beitrages auf Seite 408-411

In magnetischem Material mit rechteckiger Hystereseschleife ist die Schaltzeit dem Überschuss der treibenden MMK über die koerzitive MMK umgekehrt proportional. Dieses wohlbekannte Ergebnis wird mit der MMK-Gleichung verbunden, um die Schaltzeit und, im Falle einer belasteten Sekundärwicklung, den Sekundärstrom zu kalkulieren. Es wird gezeigt, dass die Schaltzeit die Summe 1. der Schaltzeit im offenen Kreis mit angelegter primärer Steuerung, und 2. des Verhältnisses der Flussänderung zur angelegten Spannung ist. Zwei Anwendungsbeispiele für die Theorie werden besprochen: 1. für zusammengeschaltete Kerne und 2. für Bedingungen, unter denen die Steuerung nicht von einer reinen Stromquelle kommt.

Transistor-Zerhacker und Gleichrichter-Siebketten von F. Butler

Zusammenfassung des Beitrages auf Seite 412-418

Nach einer Einführung in die Theorie der Transistor-Zerhacker werden zwei praktische Konstruktionen beschrieben, die mit 12 V gespeist werden. Eine derselben beruht auf einer symmetrischen Gegentaktschaltung mit eingebauten Verfeinerungen zur Verbesserung der Ausgangswellenform. Die andere beruht auf der Verwendung von vier Transistoren in Brückenschaltung und ist für sehr hohe Ausgangsleistungen verwendbar.

Ungewöhnliche Apparate für elektronische Musik von Alan Douglas

Zusammenfassung des Beitrages auf Seite 419-421

Die üblichen elektronischen Tastaturgeräte sind nunmehr allgemein bekannt; aber Geräte, die musikalische Töne auf andere Weise hervorrufen oder ändern, beschäftigen Forscher immer noch. In diesem Beitrag werden einige ungewöhnliche Apparate dieser Art beschrieben.

Konstruktionsfaktoren für industrielle Kaltkathoden-Zeitmesser von J. F. Young

Zusammenfassung des Beitrages auf Seite 422-425

Die Zeitmessergenauigkeit hängt von der Stabilität der verwandten Bauelemente und den Spannungsschwankungen im Schaltkreis ab. Eine optimale Anordnung liegt vor, wenn für die Kaltkathoden-Kippröhre ein Strom vor der Zündung erforderlich ist. Ein einfacher Zeitmesser kann Heizungsbelastungen nur über einen begrenzten Bereich der Netzschwankungen regeln. Wenn jedoch der zeitregulierenden Wellenform eine Stufenfunktion zuaddiert wird, kann der Bereich wesentlich vergrössert werden.