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Commentary

SEPTEMBER promises to be a month of great interest to the engineer and all concerned with the radio and electronic industry.

The industry's exhibition season reaches its climax with the opening of the Radio Show at Earls Court on 2 September, while the Society of British Aircraft Constructors Flying Display opens on 7 September and the annual meeting of the British Association for the Advancement of Science is being held at Liverpool from 2 to 9 September.

At the Radio Show the main emphasis will be, as at the immediately preceding shows, on television, and this year the general public will be given an even greater opportunity than previously of seeing, at first hand, the workings of the BBC. In previous years the control room has produced a closed circuit programme to enable the demonstration of television receivers to take place throughout the hours of the exhibition every day. This year, for the first time, the BBC is arranging to stage, twice daily, small programmes which may be transmitted from their national television network—an undertaking that will require a high standard of technical competence. Another innovation will be the use of a cinema size television screen in the television studio so that the audience will be able to see not only all that happens on the studio floor but also the picture as finally transmitted.

The more serious aspect of the industry will also be well represented by a number of exhibits showing the use of radio and radar as a navigational aid and the use of electronic equipment as a means for the controlling of various processes in industry and commerce.

One exhibit which will doubtless cause widespread interest is in the form of an outsize toy, but it nevertheless demonstrates very ably the use of radio as a control medium. It is a working model of a Churchill bridge-laying tank in which the radio control enables it to carry out any of its normal movements, including the sounding of a Klaxon horn!

Other notable exhibits will be a scale model of the 250ft radio telescope which is being built at Jodrell Bank and a full size, 25ft, guided missile as used for research purposes by the Royal Aircraft Establishment.

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At the SBAC flying display at Farnborough from 7 to 13 September a full range of airborne and ground station communication and navigational equipment will be exhibited and, in addition, a phase of the electronic industry that has expanded greatly during the post-war years will be well demonstrated. This is the measurement and, perhaps more important, the instantaneous and continuous recording of physical quantities such as strain and torque. The aircraft industry being itself comparatively young has adopted modern electronic methods of measurement and analysis more readily than the older established branches of engineering, and judged from the advances in aeroplane design and construction in recent years, there can be no doubt as to the soundness of this policy.

* * *

When visiting the wide variety of exhibitions held during the course of the year it is gratifying for all connected with the science of electronics to see the ever increasing number of industries in which electronic equipment in various forms is playing a vital part. These exhibitions must, however, impose a considerable strain on the manufacturers of electronic equipment, and it may be that there is a danger of expending too great a proportion of the industry's resources in this way.

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The third event is of a more academic nature although, as has been proved in the past, it can have great practical benefits. This is the annual meeting of the British Association for the Advancement of Science and it is, possibly, of greater interest than usual to electronic engineers since the President is Sir Edward V. Appleton, G.B.E., K.C.B., F.R.S., one of the pioneers of the theory of ionospheric propagation.

There is one session in section G, engineering, which should be of outstanding interest. It includes two papers, "The utilization of electronic computors in engineering practice" and "Some aspects of automatic computing in aircraft engineering." Although new developments, such as the transistor, will doubtless affect the physical concept of these machines, their theory and the necessary circuit techniques are now fairly well understood, and what is now required is a full investigation into the uses to which both the digital and analogue types of computor can be put. The large digital computors, such as those at Manchester and the N.P.L. may have limited use, but it is certain that more specialized and, perhaps, smaller machines will, in the fairly near future, find wide application in various fields of industrial and commercial activity and any new thoughts on this subject should prove of value.

Transistor Circuits and Applications†

By G. C. Sziklai*

Transistors have characteristics not found in vacuum tubes. Some of these characteristics are classified as symmetrical properties, the first kind of symmetry being found in the complementary characteristics of n-p-n and p-n-p transistors, the second in the interchangibility of emitter and collector in single units. Novel circuits are described which by the use of these properties permit circuit simplification and considerable reduction in the number of components.

IN the course of 1952 considerable progress has been made in the development of transistors and transistor circuits. A variety of different types of transistors were evolved, mainly because, as new applications were investigated, various new characteristics were required. Several different types of transistor developed in the RCA Laboratories are illustrated in Fig. 1, which shows them in comparison to a match stick. The first unit on the left is a point contact transistor, the second is a general utility junction transistor, and the other three are power transistors to handle dissipations up to and above 1 watt.

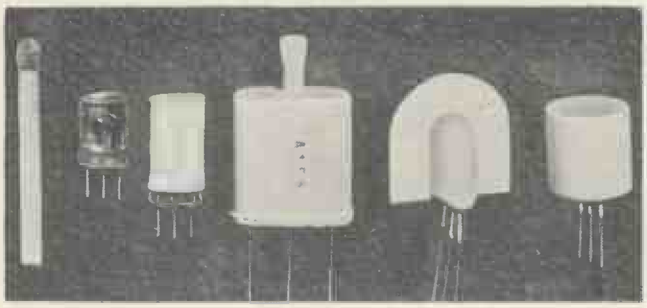

Fig. 1. Experimental transistors

With this array of transistors a number of transistorized devices were demonstrated to representatives of industry and the armed services in the United States during November 1952. In all cases the transistor permitted substantial reduction in size and weight, not merely because of its own small size, but mainly because of the reduction in the number of components which it permitted and the reduced power requirements.

The fact that the transistor could be applied to such a large variety of equipments at this stage of its development is to a large extent due to the studies of the transistor characteristics by a number of workers and their interpretations of the device as active networks with fairly well understood elements. The equivalent-*T* and -π network of a 2N34 type transistor are shown in Fig. 2. The elements shown are within the theoretical intrinsic transistor, but external resistances connected to the intrinsic terminals may degrade the performance of the transistor. The measurement of all terminal impedances may be easily performed by bridge methods such as shown in Fig. 3. and it has been found that a satisfactory balance indicated by an oscilloscope with square wave input to the bridge will provide impedance values acceptable for a wide frequency range[1,2].

A transistor circuit, approached from the point of view of small-signal analysis, is sometimes found to be an analogue

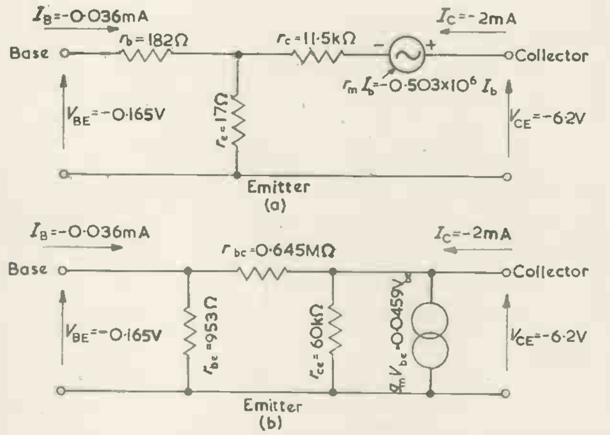

Fig. 2. *T* and π equivalent circuits of p-n-p junction transistor

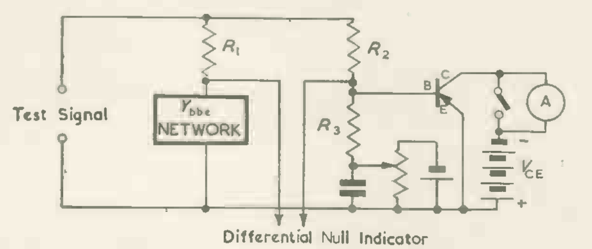

Fig. 3. Common-emitter input admittance bridge test set

of some vacuum-tube circuit. As a consequence a transistor circuit may be found for many vacuum-tube circuits by means of duality and analogue synthesis. However junction transistors have certain properties which are not available in vacuum tubes and which are unique in a number of applications[3].

Complementary Symmetry

One unique property of the junction transistors is that there are two basic kinds: n-p-n and p-n-p units. One is the symmetrical counterpart of the other, as a hypothetical positron tube would form the counterpart of the conventional electron tube. This is illustrated in the typical

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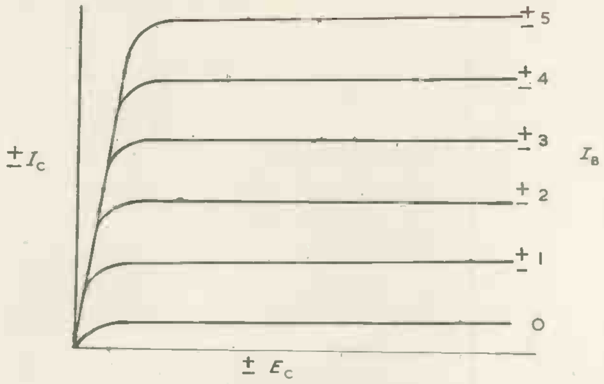


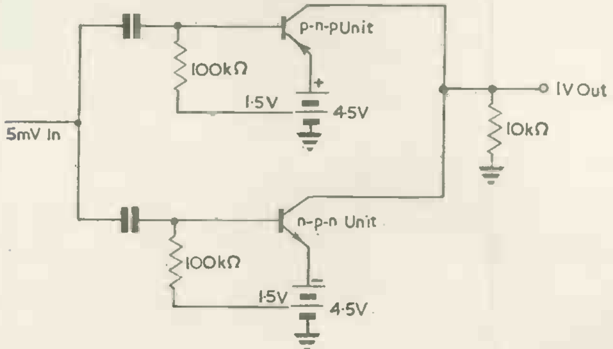
Fig. 4. Characteristic of n-p-n or p-n-p transistor

static characteristic curves of Fig. 4, where the upper signs apply to an n-p-n transistor, while the lower signs correspond to its p-n-p counterpart. The abscissa is the potential drop between the emitter and the collector (E_c), and the ordinate is the collector current (I_c) for a family of base currents (I_b).

This property can be used in a number of ways. As the base current is changed in the same direction in both units, the emitter-collector current flow will increase in one and decrease in the other. A pair of these units fed from the same signal will, therefore, provide a single-ended push-pull output. Fig. 5 shows such a single-ended push-pull circuit which operates without a transformer or phase invertor. The constants are based on an experimental set-up using developmental forms of what are now the RCA 2N34 p-n-p transistor, and its n-p-n counterpart the 2N35.

One of the features of this circuit is the position of the power supply, which is connected between the common element of the transistor and ground. As the base current is increased (it is made more negative for the p-n-p unit or more positive for the n-p-n unit) the resistance between the emitter and the collector is reduced and the potential at the output circuit will approach that of the respective power supplies. Thus if the input swings positive, the resistance of the n-p-n unit is reduced, and the top of the load becomes negative since, at the same time, the emitter-collector resistance of the p-n-p unit is increased. Similarly a negative-input swing reduces the resistance of the p-n-p unit and increases the resistance of the n-p-n unit and increases the resistance of the n-p-n unit thus making the potential of the load impedance positive. The amplifier, as shown, operates in class-A and provides a gain of 46db. The waveforms given in Fig. 6 show the output of the individual units with one unit removed and with both units inserted. With both units inserted, the applied voltage is

Fig. 5. Push-pull transistor amplifier with complementary symmetry



9V instead of 4·5V; thus the amplifier operates in class-A mode, and, since one unit acts as the load for the other, the external load impedance may be removed and a high voltage gain can be obtained for a number of special applications.

Another application of p-n-p and n-p-n transistors in combination is shown in Fig. 7, in which the two complementary units form a direct-coupled amplifier. The circuit shows only two stages, but the chain can be extended by additional cascaded stages using the same power supply. Again the power supply is between the common electrode and ground. Although the circuit provides a slightly lower voltage or current gain than the collector-to-base current gain of the transistor itself, and a considerably lower gain than would be obtainable with matching circuits, it pro-

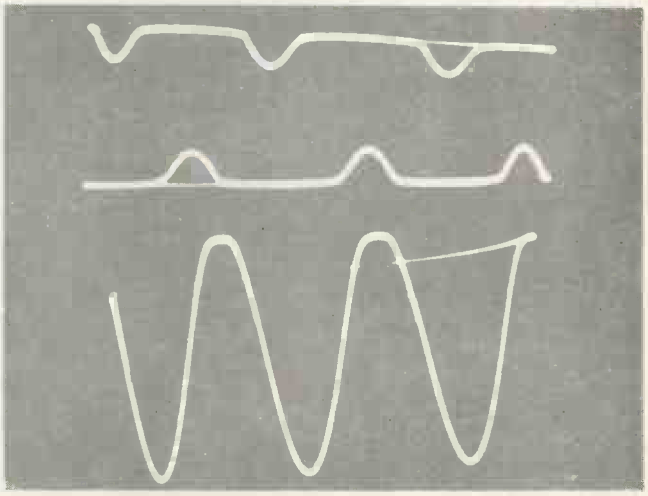


Fig. 6. Waveforms of push-pull amplifier
Top curve: With n-p-n unit only in circuit.
Middle curve: With p-n-p unit only in circuit.
Bottom curve: With both units in circuit.

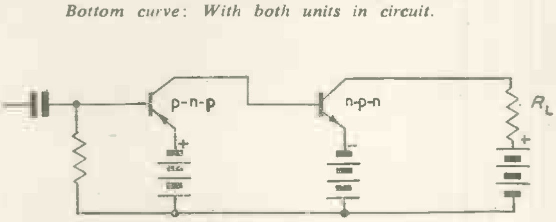


Fig. 7. Direct-coupled transistor amplifier

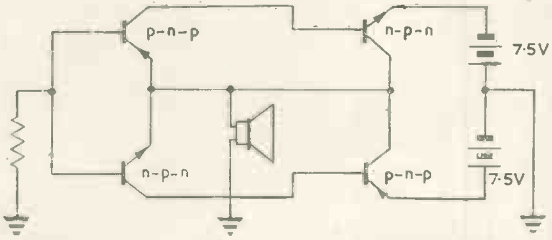


Fig. 8. Class-B push-pull transistor amplifier with complementary symmetry

vides a simple D.C. amplifier with a minimum number of components. A voltage gain in the order of 25 per stage was obtained.

The complementary symmetry of transistors finds an interesting application in the cascading of push-pull amplifier stages. This principle is applied in the two-stage direct-coupled class-B amplifier shown in Fig. 8. This circuit draws negligible current until a signal is applied. Unlike conventional class-B amplifiers, however, it does not require either an input or output transformer. As may be observed from its circuit diagram, it does not contain any parts other than the transistors themselves when operating from a high-point resistive source directly into a 16-ohm

Fig. 9. A half-watt class-B push-pull transistor amplifier

loudspeaker voice coil. The low output impedance and the stable operation is made possible by the over-all feedback which extends down to D.C. Incidentally this amplifier is a very stable D.C. amplifier, and it is unique in that it is a zero-centre D.C. amplifier.

The amplifier was built in one form with two 2N34 and two 2N35 transistors (specially selected for characteristics) in a small clear plastic case with a four-pin base, as shown in Fig. 9. The four connexions correspond to the input, output, and the two battery terminals. The 2N34 and 2N35 transistors are normally considered for applications not exceeding 50mW. In this case, however, a maximum R.M.S. power output of half a watt was obtained on a short-time basis. Since for speech or music, the average power is 10db lower than the maximum requirement, and since the amplifier has an overall efficiency (power output ÷ total power input for both the driver and the output stage) of 50 per cent at practically all levels, the unit can be used at the maximum R.M.S. rating as an audio amplifier. The power gain of the amplifier is approximately 28db. Due to feedback, the voltage gain is slightly less than unity. The total distortion at the 0·5 watt level is approximately 2 per cent.

Single-Unit Symmetry

Transistors display another type of symmetrical property

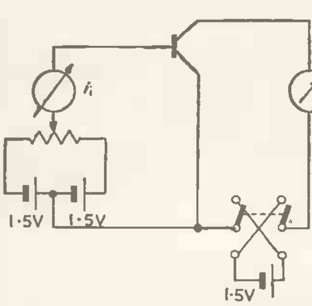


Fig. 10. Test set-up to check transistor symmetry

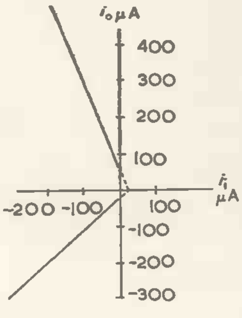


Fig. 11. Collector current symmetry in special transistors

involving a single unit. This characteristic may be best described by a simple experiment, using the test set-up shown in Fig. 10. A collector voltage with either polarity can be applied through a double-pole double-throw switch, and similarly the potential applied to the base can be varied continuously with either polarity.

The results of such a test are shown in Fig. 11. The current flowing in the base circuit is shown on the abscissa and the emitter-collector current is shown as the ordinate. When a p-n-p transistor is tested it is found that some emitter-collector current flows in either direction as long as negative bias is applied to the base. The extent of the

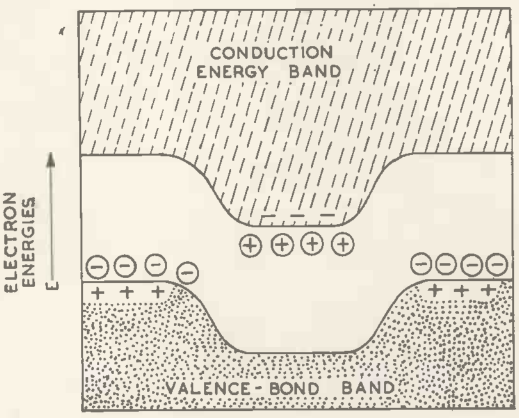


Fig. 12. Energy-level diagram of junction transistor

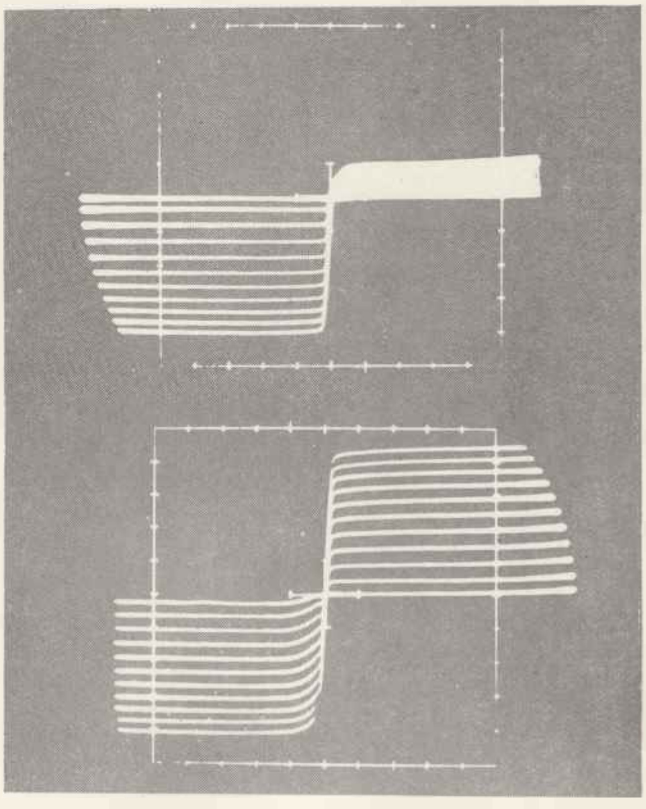


Fig. 13. Collector current/voltage curves for asymmetrical and symmetrical transistors

negative-emitter current may vary from unit to unit; however it is present in most cases.

There is no comparable action in vacuum tubes since this would require an anode emitting electrons and a thermionic cathode accepting them. With transistors, however, a reference to the energy level diagram in Fig. 12 clearly shows that units with a high degree of symmetry can be constructed. This is particularly true for the alloying process of transistor making.

In Fig. 13 there are two examples showing the variation that can be obtained with the alloying technique with respect to the single-unit symmetry. The curves are of the collector voltage and current family of two units, with the upper showing the curves of a unit made deliberately non-symmetrical, and the lower representing a symmetrical unit[4]. In both cases the abscissa represents the collector voltage, the ordinate the collector current, and the curves depart further from the origin as the base currents are increased.

The single-unit symmetry has many interesting applications since it provides a fast bi-directional switch, and a bi-directional switch can provide a sawtooth current with very high efficiency[5]. A single symmetrical transistor can provide this function with a minimum number of circuit components and with an efficiency considerably surpassing vacuum-tube circuits. Used to provide the horizontal deflexion voltage in a television receiver, the basic circuit is as shown in Fig. 14. When the base is biased negatively, the emitter-collector circuit is closed, and the current increases linearly in time according to the relation

$$di/dt = E_B/L$$

At some time the positive pulse opens the circuit and the energy stored in the inductance will be discharged through C_d in an oscillatory manner according to

$$L\,di/dt + (1/C_d)\int i\,dt = 0$$

Hence

$$i = \frac{E_B}{\sqrt{(L/C_d)}} \cos \frac{t}{\sqrt{(LC_d)}}$$

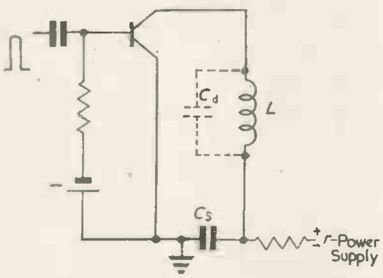


Fig. 14. Switching circuit for horizontal deflexion

If the transistor is made to be conductive again at $t - \sqrt{LC_d}$ (in other words at the half-period of the natural frequency of the yoke circuit), the energy is returned into the power supply C_s through the reverse path of the transistor. Current will change again in a linear manner until it drops to zero, when the cycle starts again.

The deflexion circuit, however, has some resistance, and the current actually follows the equations

$$iR + L\ di/dt = E_B$$

Hence $$i = (E_B/R)\,(1 - e^{-Rt/L})$$

This relationship suggests that a correcting input signal, which consists of a sawtooth as well as of a pulse, has to be applied to the base of the switching transistor, and resistive losses must be replaced by the current supply.

A Symmetrical Clamp Circuit

A conventional diode clamp circuit is shown in Fig. 15. At some given time push-pull pulses (blanking or the like) are applied to two diodes connected bi-directionally. The diodes connect the grid and charge the coupling capacitor to a predetermined potential applied to the centre tap of a transformer. After the pulse, the coupling retains its charge. The potential thus applied then forms the A.C. axis for the signal until the next pulse comes along. The push-pull transformer may be replaced by a triode phase splitter coupled through RC networks.

A symmetrical transistor connected as shown in Fig. 16 provides a simple clamp circuit. The circuit requires a single pulse (negative for p-n-p or positive for n-p-n transistors), which essentially short-circuits the emitter-collector path during the clamping interval.

Modulator and Detector Circuits

A symmetrical transistor can be arranged to provide a simple balanced modulator. The signal to be modulated is applied between the emitter and the collector in series with a load circuit, and the modulating signal is applied to the base with respect to either of the other two electrodes. The modulating signal can also be applied between the base

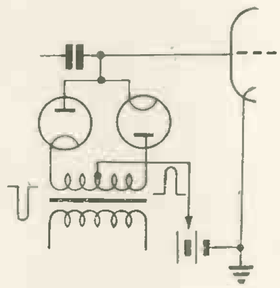


Fig. 15. Diode clamp circuit

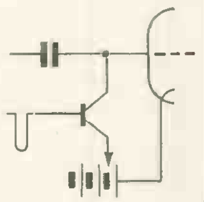


Fig. 16. Transistor clamp circuit

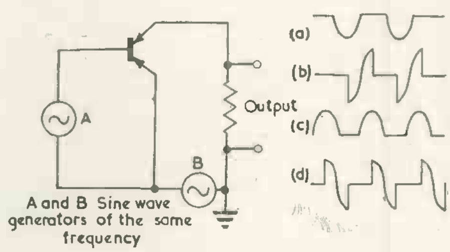


Fig. 17. Phase detector using symmetrical transistor

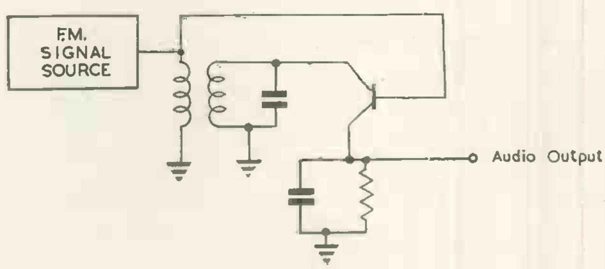


Fig. 18. Transistor FM detector

and the centre of the load circuit. Bias voltages can be applied in series with either of the two signals. When the absolute value of either of the signals is zero, the output is also reduced to zero.

The symmetrical transistor also provides a phase detector in the simple connexion shown in Fig. 17. When the signal sources A and B are in phase, the transistor conducts only during the negative cycle and the voltage drop across the load will be negative as shown in curve (a). When the source A lags source B by 90 degrees, the output wave will be as shown in curve (b) and the D.C. output is zero. Between 0 and 90 degrees the ratio of the positive and negative excursions and the D.C. output will change gradually. The same condition holds with the opposite polarity between 90 and 180 degrees. When the two sources are 180 degrees out of phase, the collector is positive. When the base is biased negatively the output wave will be shown in curve (c) and the D.C. output is positive. At 270 degrees an oscillographic indication as shown in (d) is obtained. This trace can easily be differentiated from that obtained at 90 degrees by its sequence. The D.C. output or amplitude ratio of positive and negative wave may be calibrated directly.

This type of phase detector can also be used for detecting the time relationship of a sawtooth and a pulse and thus provide automatic phase control in television synchronizing circuits.

The phase-detecting principle may also be used in F.M. reception. A simple F.M. detector using a symmetrical transistor is shown in Fig. 18. The transformer is fed directly from a signal generator, the intermediate frequency amplifier of a conventional F.M. signal receiver or the like. The F.M. signal output of the source appears across the base-emitter path of the p-n-p transistor. During the positive portion of the wave applied to the base the emitter-collector path of the p-n-p transistor is effectively open and therefore no current is flowing through the output resistor. During each negative swing the emitter-collector path is conductive. The opposite relationship applies to n-p-n transistors.

The magnitude and direction of the current flow during the conductive periods, however, will be determined by the signal developed across the secondary of the transformer. The voltage across the secondary is 90 degrees out of phase with the voltage across the primary when the applied frequency is equal to the resonant frequency. Under that condition the voltage drop across the load resistor will be zero, as it was shown in Fig. 17. As the applied frequency is changed the secondary voltage lags the primary voltage by an angle less than 90 degrees if the frequency is increased, or it will lag by an angle more than 90 degrees if the frequency is decreased. A reversal of the transformer secondary naturally will change the lag to a leading-phase relationship. As the phase relationship is changed the voltage developed across the load resistor will vary in accordance with the frequency modulation. It is of interest to note that as the signal amplitude is varied the current will tend to remain constant as is indicated by the collector current and voltage curves of Fig. 4.

Television Receiver Circuits[6]

After the discovery and a study of the symmetrical properties of junction transistors, several circuits were developed which are particularly useful in television systems. Parallel with this development some experience was obtained with point-contact transistors both in pulse and V.H.F. circuits, experience which appeared to be useful in television circuits. In view of these encouraging tests it was decided to make a general study of transistors in television receivers. For this purpose the construction of a transistorized portable television set was undertaken. The experimental model uses 37 developmental transistors and a 5 inch cathode-ray tube and is housed in a cabinet 13 by 12 by 7 inches. This portable receiver operates on a single channel using a self-contained loop, and has a total battery-power consumption of 13 watts, more than 25 per cent of which is consumed by the cathode-ray tube heater.

The signal for operation of the receiver is picked up by a loop antenna, mixed with the local oscillator in a pair of crystal diodes, and the resulting difference signal amplified by six stages of grounded-base point-contact transistors. Two second detectors provide independent signals for the sync and video amplifiers. Inter-carrier sound from the video second detector is amplified by four inter-carrier frequency stages and demodulated by a ratio detector. The resulting audio signal is applied to an emitter-follower stage driving a complementary type push-pull output stage.

The video signal is amplified by a system combining the higher input impedance of a grounded emitter junction transistor with the high-frequency response of a point-contact type.

Vertical Deflexion

The equivalent circuit of the vertical yoke consists of a resistance of 65 ohms in series with an inductance of 45mH. A peak-to-peak current of 100mA is required to deflect the beam 3 inches—the proposed picture height. Fig. 19 shows the ideal waveforms necessary to obtain linear deflexion.

The waveform to energize the yoke is obtained from a synchronized relaxation oscillator using one 2N32 point-contact transistor. The operation of this circuit is briefly

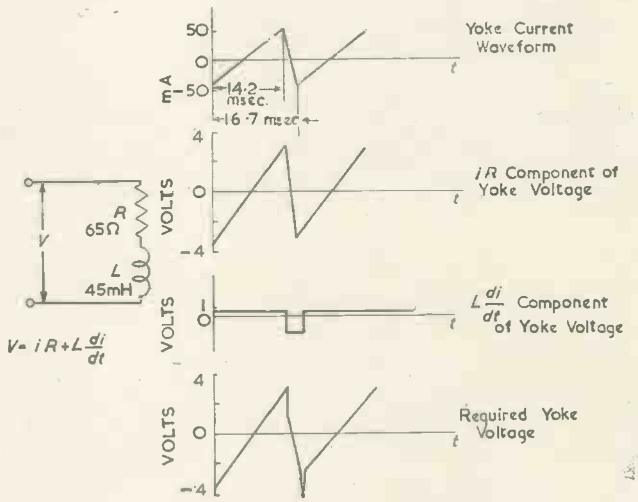


Fig. 19. Theoretical vertical yoke waveforms

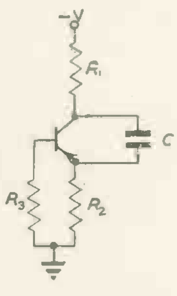


Fig. 20. Vertical oscillator

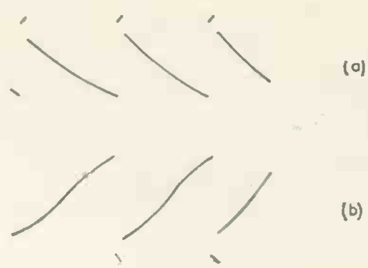


Fig. 21. Waveforms of (a) vertical oscillator output; (b) vertical yoke voltage

described by the simplified schematic shown in Fig. 20. The application of voltage causes the base of the transistor to assume a negative potential equal to the product of the base resistor R_3 and the leakage current across the collector rectifying contact.

The emitter voltage is also negative because of the current which flows through R_2 to charge the capacitor C. As long as the emitter is more negative than the base the conduction through the emitter-collector path of the transistor is negligible. As the capacitor becomes charged, however, the leakage current increases while the charging current through R_2 falls off exponentially. Eventually a point is reached when the emitter is positive with respect to the base. The capacitor then begins to discharge through the emitter-collector path. Since the current gain of the

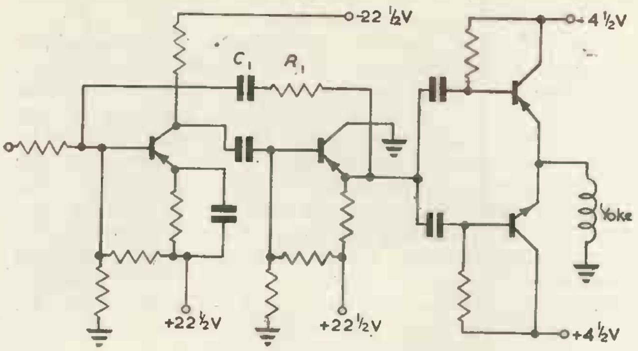


Fig. 22. Vertical amplifier

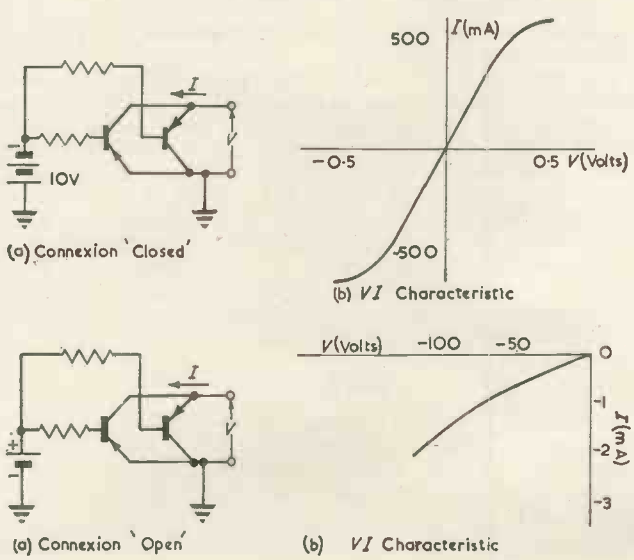
transistor is greater than unity the base current increases faster than the emitter current, and the action becomes self-sustaining and continues until the capacitor is discharged. The cycle is then repeated. Synchronization is accomplished by applying positive pulses to the emitter at a rate somewhat faster than the free running rate. Voltage waveforms are shown in Fig. 21. The ratio of sawtooth to pulse may be altered by changing the ratio of the collector resistance to the emitter resistance.

Fig. 22 shows the three-stage vertical amplifier. The first stage consists of a grounded-emitter stabilized class-A amplifier using a 2N34 junction transistor. The second stage is a grounded-collector power-junction transistor which provides the necessary drive at low impedance to operate the output stage. Frequency selective feedback, provided by R_1 C_1 serves to compensate for distortion introduced by the first two stages. The output stage consists of a grounded-collector complementary symmetrical push-pull amplifier using both a p-n-p and n-p-n transistor.

Horizontal Deflexion

The use of transistors as the switching mechanism results in a simple and highly efficient circuit which closely approximates to an ideal switch. In Fig. 23(a) two power-junction transistors with their emitters and collectors cross-connected are used in place of a single symmetrical transistor. Fig. 23(b) shows the output characteristics of the switching transistors in the "closed" and "open" conditions. The symmetry and low effective resistance of the "closed" condition and the high effective resistance of the "open" condition are apparent.

Fig. 23. Output characteristics of switching transistors



In order to control the output transistors it is necessary that their base be returned to a source of negative potential while they are conducting, and to a source of positive potential while they are cut off. This is accomplished with another form of switching circuit employing two transistors as shown in Fig. 24. The arrangement, essentially a single-pole double-throw switch, has come to be known as the "totem pole" circuit.

During the forward trace of the beam the first transistor T_1 is in a high conduction state because of the large negative bias applied through R_1, while T_2 is cut off. The bases of the output transistors are thus connected through T_1 to the negative 22½ volt supply. When a positive pulse is applied to T_1 through C_1, T_1 is cut off and the resulting negative pulse across R_2 is coupled through C_2 to T_2, causing T_2 to conduct heavily. For the duration of the positive pulse the bases of the output transistors are therefore connected through T_2 to the positive 22½ volt supply.

This indicates that pulses occurring at line repetition rate and equal in width to the retrace time are necessary to operate the horizontal deflexion system. These pulses are obtained from an oscillator essentially the same as is used for the vertical deflexion, except for the frequency-determining components. Amplification of the pulses is

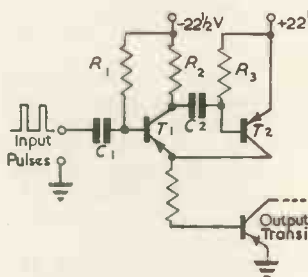


Fig. 24. "Totem pole" circuit

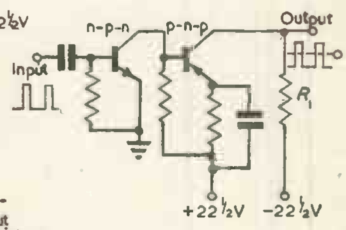


Fig. 25. Direct-coupled pulse amplifier

provided by the direct-coupled complementary symmetrical class-B pulse amplifier shown in Fig. 25.

With no input signal, both transistors are nearly cut off, and the output terminal is connected through R_1 to the negative 22½ volt supply. A positive pulse at the input causes both transistors to conduct heavily and connects the output terminal to the positive 22½ volt supply through the p-n-p unit. With this circuit it is possible to obtain approximately 40 volt peak-to-peak output pulses from a peak-to-peak input of 0·5 volt.

The preceding discussion has tacitly assumed that the elapsed time between the oscillator discharge and the opening of the switching transistors is negligible when compared to the time for one line retrace. If this were true, driven sync could be used. However, because of reactive effects associated with the input circuit of the switching transistors, there is appreciable time delay between the application of a positive pulse and the "opening" of the yoke circuit. Experimentally it was found that this delay is approximately 15μsec, which is about 5μsec longer than the time allotted for horizontal retrace in the video signal.

One effect of such a delay on the television picture is to place a black vertical bar at the right of the screen, which corresponds to the horizontal blanking in the video signal. In addition, a portion of the left-hand side of the picture appears immediately to the right of the black bar. Finally, the beam is not blanked during retrace and thus tends to smear that portion of the picture which does appear. It is apparent that an A.F.C. system can compensate for this delay by re-phasing the horizontal oscil-

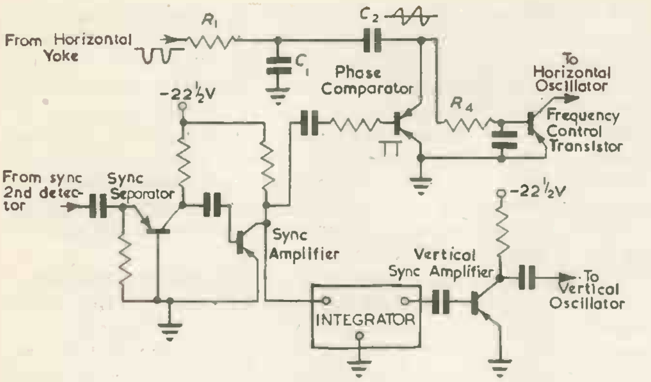


Fig. 26. Sync channel and A.F.C. circuits.

lator. A simple and reliable transistor A.F.C. system will be described in the next section.

Sync Channel

A schematic of the synchronizing circuits of the receiver is shown in Fig. 26. The sync separator and amplifier consists of two 2N34 junction transistors. The first of these is connected with grounded base. The emitter-base rectifier serves to separate the sync pulses which are coupled by the collector circuit to the following amplifier. After integration the vertical pulses are amplified by one more grounded-emitter stage and are then used to synchronize the vertical oscillator.

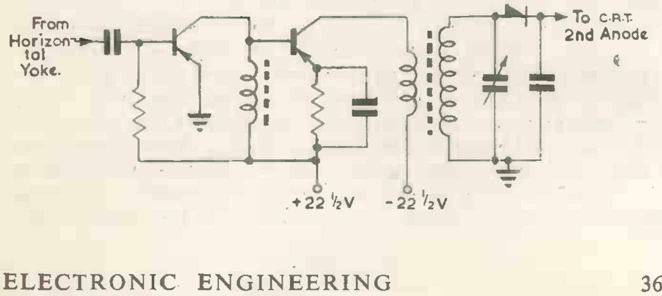
The horizontal A.F.C. system mentioned previously depends for its operation on the symmetrical properties of the phase comparator transistor. The retrace voltage pulse from the deflexion yoke is integrated by a network consisting of R_1 and C_1. In the absence of a negative sync pulse on its base, the phase comparator functions simply as a high resistance and the integrated retrace pulse appears as a symmetrical sawtooth wave, its D.C. component having been removed by C_2. There is therefore no D.C. on the base of the frequency-control transistor.

The operation is unchanged if the horizontal sync pulse arrives coincident with the time of zero voltage during the sawtooth retrace. If the sync pulse arrives when the retrace voltage is positive C_2 receives an incremental charge through the resulting low resistance of the phase comparator. During the time before the next sync pulse, this incremental charge drains off through R_4 and the frequency-control transistor. A small negative bias is thus generated on the base of the frequency-control transistor which lowers its effective resistance and causes the horizontal oscillator to fire sooner than on the previous cycle. The action continues until the sync pulse occurs simultaneously with the zero-voltage point of the sawtooth retrace. The sequence of operation is reversed if the sync pulse arrives while the sawtooth retrace is negative.

High Voltage Supply

Because the flyback voltage of the horizontal deflexion system is limited by the transistors, and since no transformer is used, it is not convenient to get the 2 000 volts

Fig. 27. High-voltage circuit



D.C. for the cathode-ray tube directly from the deflexion system. For this reason a separate system shown in Fig. 27 was used. This is essentially a two-stage class-B amplifier driving a tuned transformer in the output. The negative flyback pulse is lightly coupled to the amplifier and causes the first transistor to conduct, thereby applying a conduction bias to the output stage. The output transistor, a power type, conducts heavily and supplies power to the tuned transformer at the horizontal frequency. The choke in the base circuit of the output transistor provides a low-impedance D.C. path and ensures that any leakage current in either transistor does not bias the output to conduction except when the system is pulsed. A half-wave selenium rectifier and filter provide the 2 000 volts D.C. from the secondary of the tuned transformer.

Overall Data and Tests

The general layout of the receiver is shown in Fig. 28 The signal channel is built on the vertical plastic shelf in the centre of the receiver. The deflexion chassis is under the neck and the socket of the cathode-ray tube. Batteries are placed on the left-hand side of the plastic shelf, and the high-voltage supply is placed beyond the batteries to provide the least disturbance to the rest of the receiver. The total weight of the receiver with batteries is 27lb. The cathode-ray tube is the 5FP4 type and provides a

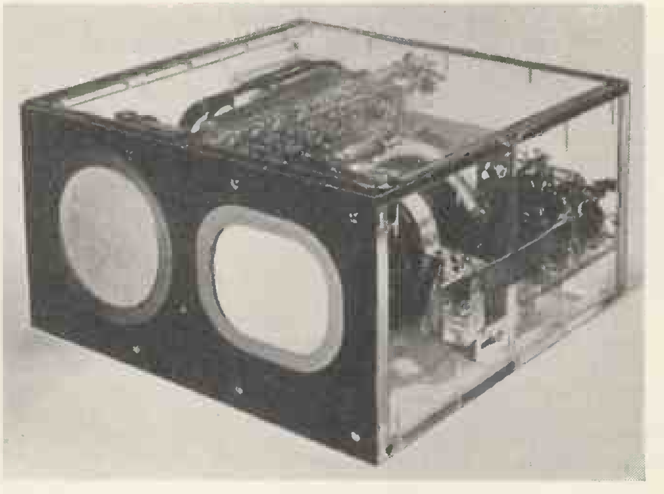


Fig. 28. General layout of television receiver

3 × 4 inch picture with a highlight brightness of approximately 10 foot lamberts.

The receiver was tested in the North Jersey area and satisfactory reception was obtained with the built-in loop within an area of approximately five miles' radius from WNBT. With the aid of a simple dipole held by hand about seven feet from the ground, satisfactory reception was obtained out to Elizabeth, N.J., about 15 miles away from the Empire State Building. The receiver was also demonstrated on many occasions from a re-radiated signal of approximately 5mV/metre in Princeton, N.J. This sensitivity was not limited by noise and could be increased substantially with additional stages or with the improved transistors which are now available.

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A Band Pass Filter for Low Frequencies

With Special Reference to Electroencephalographic Studies

By G. W. Morris* and P. G. M. Dawe†, M.Sc.(Eng.) B.A.

The band-pass filter described was developed to deal with frequencies of 8-13c/s and has been used to isolate the "alpha rhythm" in electroencephalographic studies. It is composed of a number of RC valve amplifier resonant circuits arranged to give a substantially flat frequency response within the limits of the pass band and a sharp cut-off beyond. Two alternative types of circuit are described and compared from the point of view of stability.

THE waveforms recorded in electroencephalographic studies have a complex character and usually consist of a mixture of low frequency sinusoidal waves together with aperiodic waveforms, more in the nature of pulses or spikes. Investigators working in this field have written at some length on the difficulties involved in analysing and interpreting such waveforms[1-2], and instruments have been designed to facilitate this work, for example, by providing a complete analysis of the frequency spectrum present in the record, over short time intervals. One well-known instrument for this purpose is the Ediswan Electronic Analyser, described by Baldock and Walter[3-4], but other investigators have developed alternative systems[5-8], using optical, mechanical or electronic methods of analysis.

For some purposes, however, electroencephalographic investigations may be conducted in terms of a particular physiologically significant band of frequencies, and in making studies involving such a band it is then of considerable advantage to be able to record the frequencies within that band without their being obscured by other frequencies. There are generally considered to be four such bands between the frequencies of 0·5 and 35c/s, as follows:

Delta band	:	0·5 - 3·5c/s
Theta band	:	4·0 - 7·0c/s
Alpha band	:	8·0 - 13·0c/s
Beta Band	:	14·0 - 35·0c/s

The filter units here described were designed in the first instance to aid studies involving frequencies within the "alpha" band of from 8-13·0c/s inclusive, but with small modifications they can be used for the other bands.

Requirements

The ideal band-pass characteristic to be aimed at was given as one having an attenuation of 10 times at frequencies of 1c/s beyond the band edges, the response within the band being as uniform as possible and the phase-shift reduced to a minimum. It was also considered important to give attention to the factors determining the stability of the overall characteristic, the possibility of discriminating against ripple voltages and slow variations of the H.T. supply, the insertion loss of the filter and the overall power consumption. In E.E.G. work it may be necessary to use 6, 8 or more filters simultaneously, so that cost, power consumption, and the facility with which construction and adjustment may be carried out are important considerations. Finally, the units were required to work in conjunction with standard E.E.G. recording equipment which requires a push-pull input.

Rear view of the complete filter

Method of Construction

The filters were constructed by combining several resonant circuits of the type which utilize resistors and capacitors together with valve amplification. These were tuned to different frequencies, and the input signal was fed to each in parallel via a resistance network, as shown in Fig. 1. Two types of filter units were, in fact, developed, differing mainly in the type of basic resonant circuit employed, and a comparative account is given of their performance. In one case the basic resonant circuit consisted of a 2-mesh *RC* network giving zero phase shift at resonance, together with a two-stage amplifier, as shown in Fig. 2(a), while in the other case a 2-mesh *RC* network was used in conjunction with a single stage amplifier, as shown in Fig. 3(a). Both these types of circuit have been fully described in the literature on *RC* oscillators and resonant circuits[9-12], and it is therefore proposed to discuss here only the points and circuit details arising out of this particular design and application.

The method of connecting the circuits, indicated above, was chosen in preference to the more conventional method of cascading, or top-capacitance band-pass coupling, since it gives a somewhat smaller phase shift throughout the band, and a more flexible adjustment of the overall bandwidth. Further, an improved band-pass characteristic can be obtained with the aid of a push-pull feed to the filter or by making use of push-pull output connexions. With such anti-phase signals available, two of the resonant circuits tuned to frequencies just beyond the edges of the band, can be joined to the paralleling network in phase opposition to the other circuits. Fig. 4 illustrates diagrammatically how the phase and amplitude of the voltage from

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each resonant circuit will vary throughout the band and so contribute to the overall characteristic. At frequencies just beyond the upper band edge, a large leading voltage component will be combined with a number of lagging components to give a cancelling effect and a consequent reduction in output voltage. The cut-off at the band edges is consequently sharpened, and can, in fact, be adjusted to give a frequency of infinite attenuation in the manner of an *m*-derived filter characteristic.

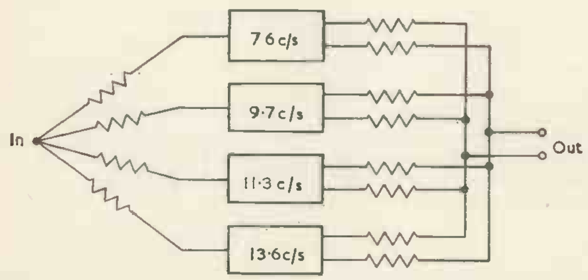


Fig. 1. General scheme of the band-pass filter, showing paralleling of a number of resonant circuits

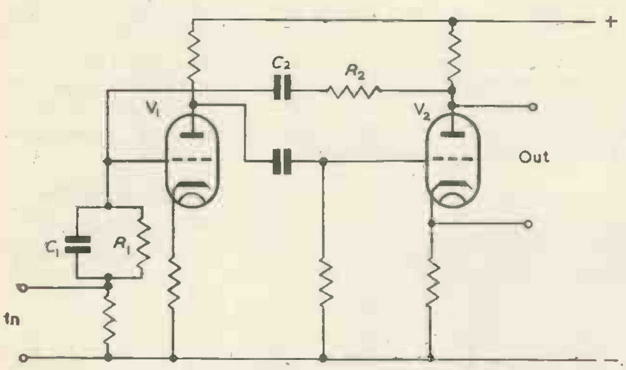


Fig. 2(a). The basic zero phase-shift resonant circuit

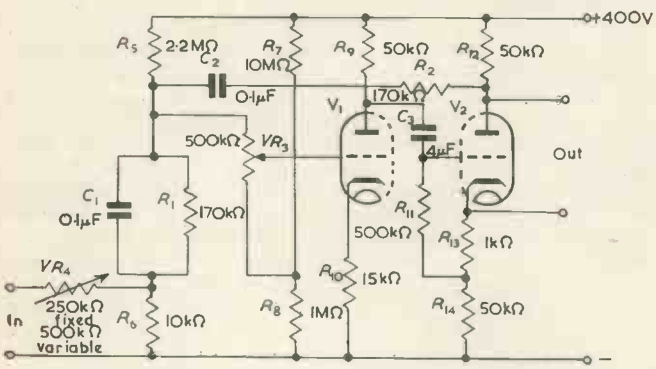


Fig. 2(b). Practical form of Fig. 2(a), using an ECC35 double triode valve

R_1 R_2 C_1 C_2 are typical values for a frequency of 7·6 c/s. Actual values depend on required frequency.

Circuit Details and Performance

Fig. 5 shows the overall response curve and the phase-shift characteristic obtained by combining four of the resonant circuits of the type shown in Fig. 2(b). The individual circuits in this case were tuned to frequencies of 7·6, 9·7, 11·3 and 13·6c/s, having Q's of approximately 6·3, 4·5, 4·5 and 6·0 respectively. These values were chosen empirically to give the best overall characteristic. Each

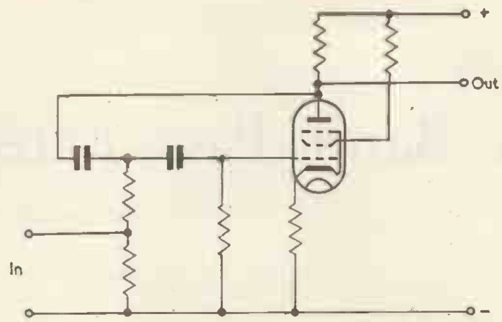


Fig. 3(a). Basic circuit of the single valve resonant network

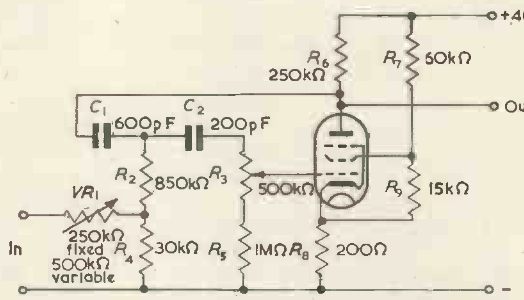


Fig. 3(b). Practical form of Fig. 3(a), using an EF91 pentode valve

R_2 R_3 C_1 C_2 are the values required for a frequency of 7·5 c/s, and would be varied for other frequencies.

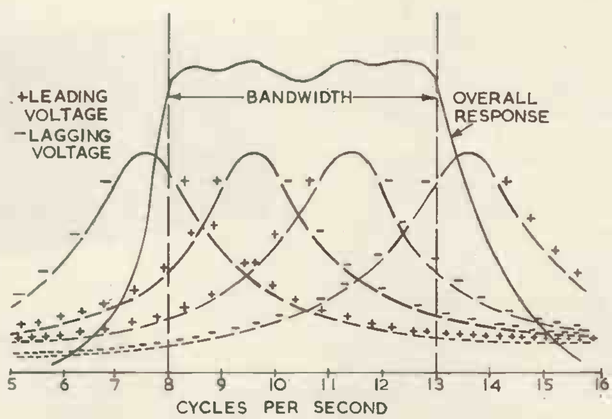
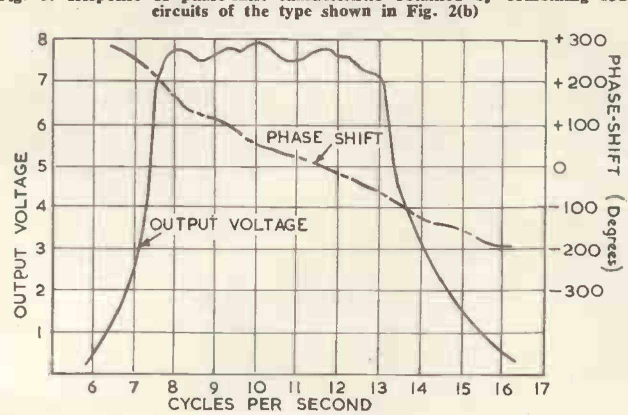


Fig. 4. Illustrating how the phase and amplitude of the voltage from each resonant circuit will vary throughout the band and so contribute to the overall characteristic

Fig. 5. Response of phase-shift characteristic obtained by combining four circuits of the type shown in Fig. 2(b)



circuit used an ECC35 double triode valve, the outputs being combined on two 3MΩ star connected networks, one being fed from the cathode loads and the other from the anode loads, giving a push-pull output. The Q values were adjusted by means of the potentiometer VR_3, which controls the overall gain of the circuit, and the level of input to each circuit was made variable by VR_4. In practice it was found that the best overall characteristic was obtained by working the two circuits at the band edges with somewhat higher Q values and levels of input than in the case of the other circuits.

With $R_1 = R_2 = R$, and $C_1 = C_2 = C$, the differential equation for the grid voltage of V_1 of the circuit of Fig. 2(a) is easily obtained from the mesh equations as:

$$d^2v/dt^2 + \left(\frac{3-k}{RC}\right) dv/dt + \frac{v}{R^2C^2} = 0 \quad \ldots\ldots (1)$$

where k is the overall stage gain from the grid of V_1, to the anode of V_2 and the parallel combination of the valve internal impedance and load resistance is lumped together as part of the series resistance R. This equation has complex conjugate roots for values of k between 1 and 5. From the roots of the equation, the frequency of the resonance is given as:

$$f = \frac{\sqrt{(6k - k^2 - 5)}}{4\pi RC} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

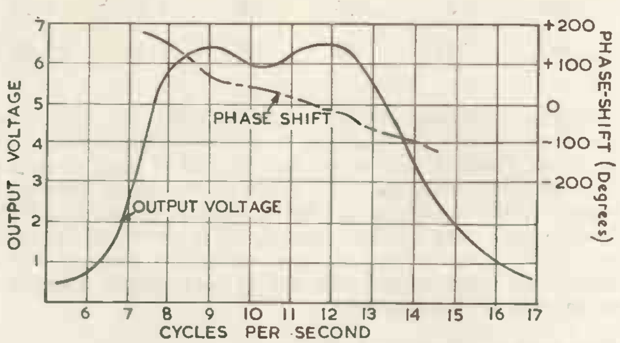


Fig. 6. Response of filter unit combining five resonant circuits of the type shown in Fig. 3(b)

and the Q value of the circuit, when being used as a selective amplifier, is given as:

$$Q = \frac{\sqrt{(6k - k^2 - 5)}}{2(3-k)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

hence for Q's of the order of 5, the amplifier gain is required to be about 2·8 times. A considerable amount of negative feedback may therefore be used with the circuit, serving to stabilize the gain, and hence the frequency and Q of the resonant circuit. This is most easily applied in the form of current feedback by using large resistors in the cathode circuits of V_1 and V_2.

To secure discrimination against ripple voltages and slow changes in the H.T. voltage the stage gain of V_2 was adjusted to unity. With equal anode loads, this arrangement provides for equal anode voltage and current swings for V_1 and V_2, (neglecting any voltage drop in the coupling network). A ripple signal appearing at the anode of V_2 would then be offset by the ripple signal fed in from the anode of V_1 via the grid circuit of V_2. The equal anode current swings, in phase opposition to each other, also prevent signal voltages from being developed across the internal impedance of the H.T. source.

The operating point for V_1 was secured by making the lower end of the grid leak, R_8, in the form of a potential divider across the H.T. An additional resistor R_5 was connected between the upper end of the gain control R_3 and the H.T. line in order to maintain a constant bias voltage on V_1 while varying the circuit Q by means of R_3.

A feature of this circuit is the ease with which a push-pull output may be obtained by taking the output from across the anode and cathode circuits of V_2. This is important when it is required to feed the output into a balanced E.E.G. amplifier. The insertion loss of the unit was measured as 20 times, an acceptable figure since the E.E.G. amplifier following the filter is a high gain unit.

Fig. 6 shows the overall response characteristic of a filter unit combining 5 resonant circuits of the type shown in Fig. 3(b). These were tuned to frequencies of 7·7, 9·2, 10·6, 11·8 and 13·9c/s, having Q values of 4·8, 4·3, 4·5, 4·6 and 4·0 respectively which are somewhat lower than

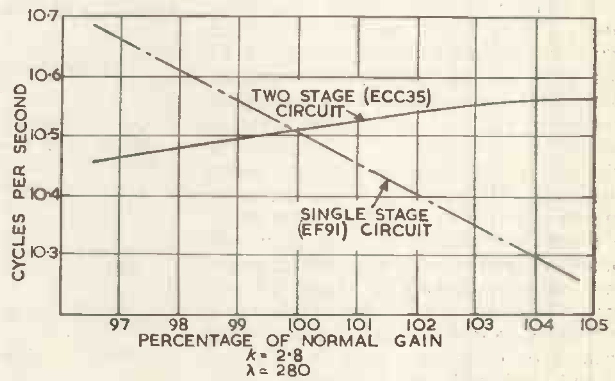


Fig. 7. Variation of frequency with stage gain

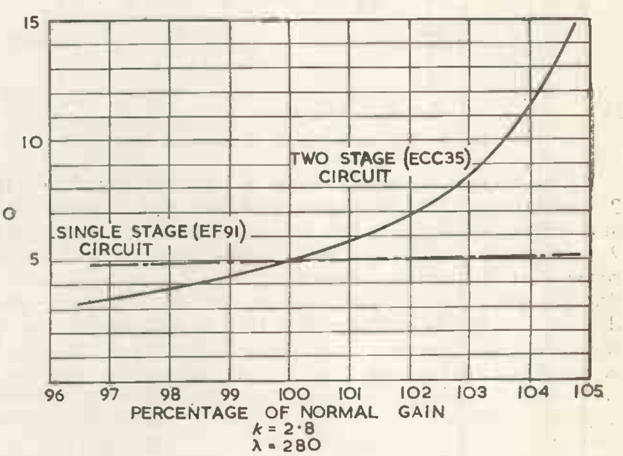


Fig. 8. Variation of circuit Q with stage gain

in the other unit. With this circuit, large values of stage gain are required in order to obtain Q values of the order of 5, and in practice there is a limitation of the Q obtainable as the frequency increases, due to losses in the capacitors. The maximum theoretically possible is given as $\sqrt{\lambda/2}$, where λ is the stage gain, but the values which can be obtained in practice are somewhat less than this, especially at the upper edge of the band. As in Fig. 2(b), R_3 controls the overall gain of the circuit, but in this case a change in gain has a larger effect on the resonant frequency of the circuit than on the Q, although both are affected. (See Figs. 7 and 8.) The optimum circuit Q is given from the differential equation of the circuit[9] as:

$$Q = \sqrt{\lambda/2} \cdot \frac{1}{1 + \sqrt{[R_0/(R_3 + R_5)]}} \quad \ldots\ldots\ldots (4)$$

where

$$R_o = \frac{R_6 R_a}{R_6 + R_a}$$

while the resonant frequency is given as:

$$f = \frac{1}{2\pi\sqrt{[\lambda C_1 C_2 (R_2 + R_4)(R_3 + R_5)]}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(5)}$$

Fig. 7 compares the variation in resonant frequency with stage gain for the two circuits, and Fig. 8 gives a similar comparison for the variation in circuit Q with stage gain.

Stability Considerations

It can be estimated from equation (3) that at a Q value of 5, a 1 per cent rise in the overall gain will produce approximately a 15 per cent rise in the Q of the circuit. Similarly, from equation (2) it may be estimated that a 1 per cent increase in overall gain will cause a 0·15 per cent change in resonant frequency. From equations (4) and (5) it can be estimated that for Q's of the order of 5, a 1 per cent change in stage gain will produce approximately equal changes in frequency and Q of about 0·5 per cent. Thus the Q of the single-valve circuit is inherently more stable with respect to stage gain than the 2-valve circuit, although the frequency stability is somewhat worse.

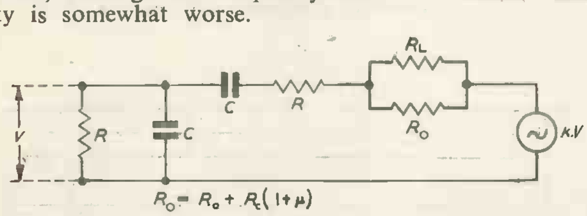


Fig. 9. Equivalent circuit of two-stage filter

At first sight it would seem as though the two-stage 2-mesh circuit will be highly unstable with respect to Q and output amplitude when compared with the single-stage 2-mesh circuit, although more stable than the latter with respect to frequency. However, Fig. 8 and equation (3) do not give the complete picture of the operating conditions. In the first place, the true equivalent circuit of the two-stage 2-mesh circuit is as in Fig. 9. From this it is apparent that as the valve μ increases due, say, to a rise in H.T. voltage, so also does the effective internal impedance of the valve V_2, which is raised by current feedback to an amount $[R_a + R_c(1 + \mu)]$, and this will increase the attenuation of the feedback network. This effect tends to compensate for the rise in stage gain, k, which is also a result of the rise in valve μ. Secondly, due to the presence of current feedback, the overall stage gain is found to remain very stable with respect to changes in H.T. and L.T. Where capacitance coupling is employed between the two sections of the ECC35, a 100 volt change may be made in the H.T. voltage (from 335 to 435 volts) before the stage gain changes by 1 per cent, and there is no observable change in stage gain produced by a variation in the L.T. from 5·0 to 7·0 volts. Overall measurements on the stability reveal that for a rise in the H.T. voltage, the change in the anode impedance of the valve V_2 is the predominating effect and there is actually a fall in the output amplitude and circuit Q with a rise in the H.T. volts, and vice versa. Thirdly, for the single-stage 2-mesh unit, the stability is worsened in practice because the high values of stage gain required leave little margin for stabilizing the circuit with negative feedback. In the circuit adopted, the effects of current feedback had to be minimized by using a small cathode resistor and passing the current from the screen potential divider through it in order to develop the necessary bias voltage. With this small amount of feedback it was calculated that a 1 per cent change in amplifier gain will result from approximately 1 per cent change in valve μ, or a 3 per cent change in valve R_a, and, in practice, can be produced by 0·5 per cent change in heater supply voltage. The EF91 valve is a miniature tube and the variation in stage gain with heater voltage over the range of from 5·0 to 7·5 volts in this particular circuit is found to be very considerable, and to follow an approximately linear relationship. With respect to variations in H.T. supply voltage, however, the circuit can be made very stable provided that the screen voltage is carefully chosen.

TABLE 1

Comparisons of the overall stability of the two units with changes in mains supply voltage

A.C. MAINS VOLTAGE	PERCENTAGE OF NORMAL MAINS VOLTAGE	H.T. VOLTAGE	L.T. VOLTAGE	PERCENTAGE OF FULL OUTPUT AMPLITUDE		BAND COVERAGE (c/s)	
				ECC35 FILTER	EF91 FILTER	ECC35 FILTER	EF91 FILTER
200	80·6	325	5·0	115	40	8–13	9–14
248	100·0	400	6·4	100	100	8–13	8–13
260	104·8	442	7·05	80	138	8–13	7·5–12

Figures obtained on the overall stability of both units (i.e. with respect to variation in mains supply voltage) are given in Table 1 and show that the EF91 unit compares unfavourably with the ECC35 unit. By using somewhat higher value resistors in the feedback network of the 2-valve unit (and smaller capacitors) it should be possible to reduce the effect of the changing internal valve impedance, and so stabilize the Q and output amplitude to an even higher degree.

Circuit Refinements and Comparisons

The triode sections of the ECC35 unit may be D.C. coupled using a potential divider returned to earth between the anode of V_1 and the grid of V_2. The frequency range can then be extended downwards by a simple change of R and C, and the values of these components calculated approximately from equation (3), the final values being found empirically. Such a D.C. coupling somewhat impairs the stability of the circuit Q, since the stage gain of V_2 must now be increased in order to compensate for the attenuation through the coupling path and so keep the anode swings identical. This implies a reduction in the current feedback on V_2; in addition, the D.C. coupling does not permit the bias voltage on this valve to rise so much with increasing H.T., hence a bigger change in mutual conductance is to be expected for any appreciable change in the H.T. voltage.

The number of components involved in the construction of a multi-band filter is somewhat less with the single-stage unit and this unit gives a simpler construction and adjustment. As against this, however, it is at a disadvantage from the viewpoint of stability, and also circuit Q, for it is more difficult to obtain a high Q to give a good cut-off at the band edges. Also an additional valve is required for the purpose of providing a push-pull output. With this valve the overall insertion loss is reduced to a figure of 8 times, compared with a figure of 20 times for the ECC35 filter.

With regard to power consumption there is little to choose between the circuits, since although the total anode current consumption of the EF91 is considerably less than

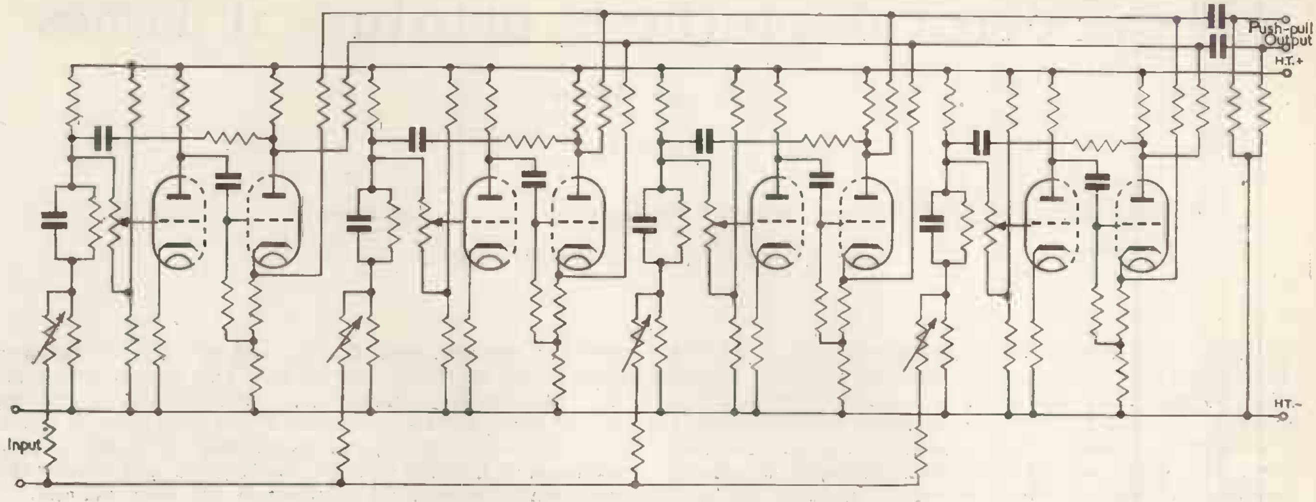


Fig. 10. Complete circuit of an alpha band filter

the total anode current consumption of both halves of the ECC35, there is the additional current consumption through the screen potential divider (necessary for developing the bias voltage) which adds to the total consumption of the stage. Fig. 10 shows the complete circuit diagram of an alpha band filter, as finally adopted.

The general method of using such a filter in E.E.G. practice has been to switch it into each channel in turn, taking the input to the filter from across the coils of the recording pens. When a larger number of units have been built it is intended to filter six channels simultaneously so that phase differences between channels can be observed and localization studies facilitated. Fig. 11(a) shows a section of an electroencephalograph record, the upper channel recording the normal unfiltered wave, and the lower channel the filtered signal from this channel. This arrangement facilitates the estimation of such physiologically significant factors as the percentage time " alpha " content of the record. Fig. 11(b) shows the effect of filtering a complex waveform, artificially produced by combining the outputs from several sinusoidal generators at frequencies of 1, 2, 3, 4 and 10c/s. During the recording the 10c/s wave was stopped and another wave of 8c/s put in its place. Measurements of both frequency and amplitude can be made from the filtered record, whereas such observations are quite impracticable with the original record. Both these examples show the great simplification which is possible when only frequencies of 8-13c/s are of interest. However, it is not intended that the original record should be dispensed with in analysis, but that it should be studied in conjunction with the filtered record.

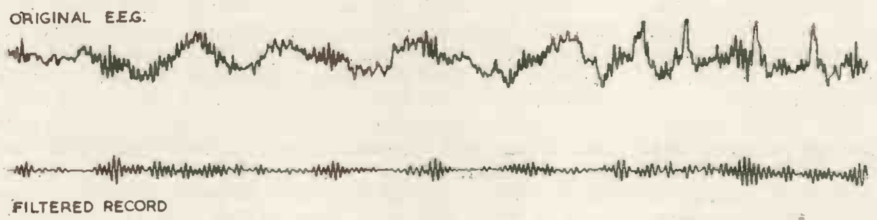


Fig. 11(a). The effect of filtering an electroencephalograph record

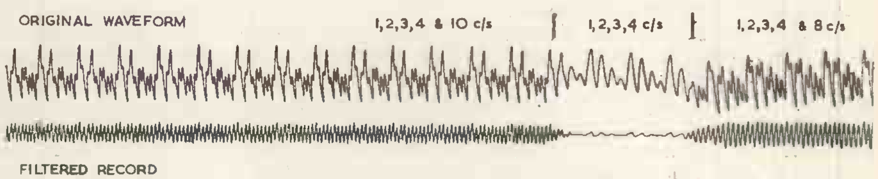


Fig. 11(b). The effect of filtering a complex waveform

Acknowledgments

In conclusion, the authors would like to thank the other members of the biophysics unit for their assistance in the construction of apparatus and in the making of experimental recordings, and grateful acknowledgment is also made to the Board of Management, Crichton Royal Hospital, for providing the facilities for this investigation.

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The Electrical Synthesis of Musical Tones

By Alan Douglas

(Part 3)

Intermodulation; Loudspeakers; Volume Level; Expression; Aesthetic and Commercial Considerations.

THE presence of many independent tone sources, whether electrical or physical, gives rise to complex beating effects as previously stated. In the case of pipes sounding together, for instance, the only coupling between them is that due to air pressure waves radiated from the resonating tubes. No energy can be transferred through the point of common coupling, i.e., the windchest. The same applies to groups of instruments playing in an orchestra. On the other hand, a high degree of controlled intermodulation exists in a piano, giving rise to tonal variations without number. The energy is transferred by air coupling between the strings and mechanical coupling through the soundboard. In a skilfully designed piano, the effect is colourful and rich; in a poor one, harsh and "jangly".

Electrical tone generators are frequently coupled by common circuits, but so far as the generated waveforms are concerned, this is the last thing required. The presence of small A.C. components in the H.T. supply line, stray fields, leakage through circuit wiring and valve capacitances, makes it difficult accurately to confine the entire signal to its proper path. With magnetic and electrostatic generators there will be leakage and fringing at high frequencies and if many circuits having differing impedances are combined there will be a redistribution of energy between the components. Screening may also present problems.

Many valve generators derive all their power from a common H.T. line. This is responsible for a good deal of mutual coupling if amplifiers are also fed from it. The latter should always have their independent source of power and the use of gas tube or hard valve regulators for the oscillators with their well-known reduction in internal impedance is strongly recommended. This greatly reduces any tendency to feedback.

In valve generators not accurately in tune, it is not so much the departure from pitch of the prime tone or fundamental which is objectionable, but the beating between adjacent high harmonics not in tune. These can form a whole series of sum or difference tones which will not be harmonically related to the series being generated and will in consequence introduce marked distortion even if of very small amplitude. This condition is aggravated if many pitches covering a widely-separated frequency band are in use together, e.g., an organ pedal note of D♯ 38·9c/s together with manual notes of D♯ 155·6c/s, F♯ 185·0c/s, A 220·0c/s and C 2093·0c/s. It is interesting to note here that while the tuning accuracy of sine wave generators for additive synthesis must be of the order of one part in 10 000 or better, if the generator supplies all its own correctly related harmonics the tuning for adjacent fundamentals need not be better than one part in 1 000. The highest stable accuracy so far attained commercially is two parts in 1 million of the interval of a semitone. This accuracy would not, of course, be found in a pipe organ or piano, and could not be retained; but the highest possible accuracy should be aimed at because as the volume level is raised the small components which are inaudible below a certain loudness may then obtrude and cause frequency distortion.

Some forms of distortion very difficult to remove can arise from the non-linear performance of iron-cored inductors supplied with D.C. as well as A.C.; for instance, some forms of Hartley oscillator. This is also noticeable in pulse-shaping transformers and frequency-dividing transformers. Careful selection of the iron may be necessary, indeed in winding iron-cored coils for complex wave oscillators it is difficult to get two coils to give exactly the same waveform without selection of the core material. A peculiarly irritating form of distortion can arise from shock excitation of iron-cored coils used in tone-forming circuits.

If valve oscillators are sufficiently well isolated from the load, very little tendency to "pull" or draw into tune will occur; circuits have been developed in which adjacent oscillators can be tuned to within a minute part of a semitone of the same pitch without pulling.

Main amplifiers are a prolific source of distortion, but this is only mentioned in passing since the literature now describes many circuits substantially free from distortion. Some applications of negative feedback tend to produce frequency distortion and this often occurs in the form of self-oscillation above the audible range, which however, manifests itself in the form of objectionable beats in the audible range. However, the sum of all the distortions can be kept quite small in a properly co-ordinated music generator; it is all the more regrettable that the haphazard application of loudspeakers can reverse this condition.

Undoubtedly no loudspeaker can simulate or re-create the spread of acoustic energy associated with some physical sound sources. On the other hand, it can be practically perfect for certain other physical sources. A great deal of work has been done on this subject, especially in the United States. It has been found that for moderate powers, that is, powers of the same order as the original physical source of sound, a single loudspeaker will reproduce any single source very well. In the frequency band 80-8 000c/s this applies to multiple sound sources also. It is in the range 32-100c/s, and especially when such frequencies are combined with those in the higher ranges, that the effect is different.

The economics of present-day design call for the lowest cost and the smallest number of parts in a reproducing system. For this reason, the frequency range of many loudspeakers has been gradually extended. This is in general a useful property where moderate powers are concerned, say up to 10 watts electrical input. In the case of simple melodic instruments, and those so far commercially produced for percussion tones, such wide-range loudspeakers perform adequately. The input power here never exceeds 12 watts. The real power level required is always closely related to the fidelity of the sound. For example, in instruments of a percussive type the means by which the strike tone is produced in a piano cannot be exactly simulated, since it contains many inharmonic components. To achieve the same apparent loudness with harmonically related components requires that these should have a greater energy content. Indeed, the purer the tone the greater the energy required of the initial waveform, and the sole example of a purely percussive keyboard instrument had to be withdrawn from manufacture because the energy of the charging pulses to the *RC* decay networks became so high that capacitor breakdowns were

A MODERN ELECTRONIC ORGAN USING VALVE GENERATORS

SPECIFICATION

SWELL

	ft.
Bourdon	16
Violin Diapason	8
Hohl Flute	8
Rohr Flute	8
Salicional	8
Orchestral Flute	4
Salicet	4
Flautino	2
Clarinet	8
Oboe	8
French Horn	8
Vox Humana	8
Trompette	8
Clarion	4
Tremulant	

Swell Octave
Swell Sub-octave
Swell Octave to Great
Swell to Great
Swell Sub-octave to Great
Swell Octave to Pedal
Swell to Pedal
3 combination pistons

GREAT

	ft.
Contra Dulciana	16
Open Diapason	8
Clarabella	8
Dulciana	8
Gemshorn	8
Principal	4
Gemshorn	4
Fifteenth	2
Tromba	8
Tromba Clarion	4

Great Octave
Great Octave to Pedal
Great to Pedal
3 combination pistons
Great to Pedal reversible piston

PEDAL

	ft.
Open Diapason	16
Contra Dulciana	16
Bourdon	16
Contra Bassoon	16
Octave	8
Violincello	8
Choral Bass	4

4 general pistons
4 general toe pistons

Swell, Great and General Crescendo Balanced Pedals with visual indicators.
All pistons immediately adjustable at console by motor-driven setter.

frequent at an acceptable loudness level; for to sufficiently increase the working voltage rating would have made the cost prohibitive.

Where large powers are required, or extreme bass notes, the loudspeaker problem is admittedly difficult and quite considerable intermodulation can result. Such conditions only apply to electronic organs. Here we not only have to contend with the range of power and pitch, but must consider the nature of the corresponding physical sound source.

A group of four trumpets playing together is practically a point source of sound. Thus such a sound is adequately reproduced by a wide-range single loudspeaker. The average power in watts would be about $1\frac{1}{2}$, the peak power perhaps 8 watts, the apparent loudness being due to the high sensitivity of the ear in the frequency range occupied by the principal harmonics of the trumpet. One large pedal pipe in an organ is, however, far from a point source. The energy from a 16ft pitch wood pipe is radiated

duplicate this train of events, therefore the applied waveform excites the coil more rapidly since even if the attack is gradual it is merely increasing the intensity of the first cycle or two of movement. There is no change in the mode of vibration, only the amplitude, and the effect on the ear is one of rather suddenly applied power.

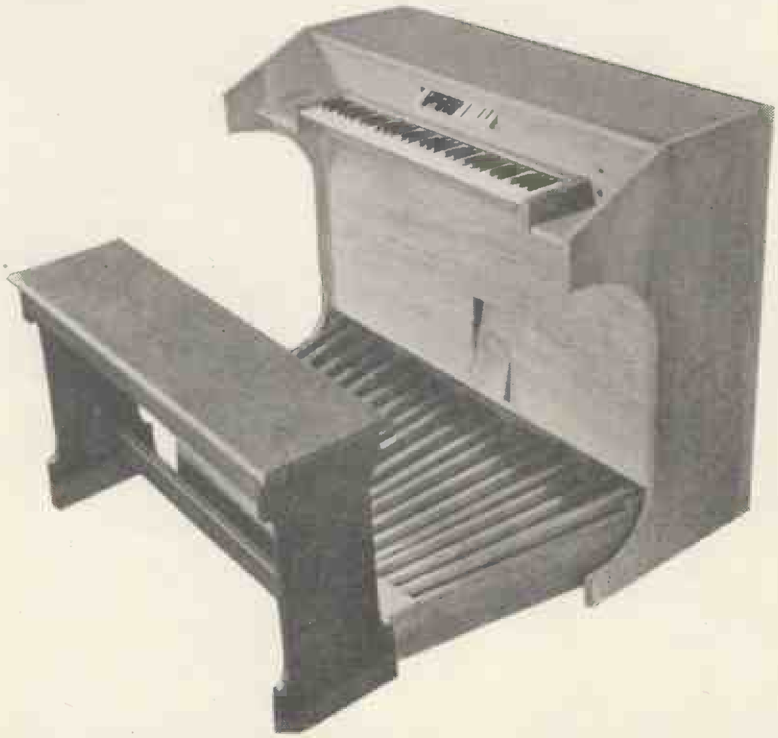
For moderate powers a multiplicity of small cones can be used. One effective unit comprises nine 10in cones placed as closely together as possible on a flat baffle. Each cone works at a low level, but the spread of acoustic energy is much superior to that from a single large cone. There is about 675 square inches available as against about 240 square inches for an 18in cone. The radiating area of the smallest pedal pipe having a pitch of CCC 32·7c/s is about 1 730 square inches. The overall cost of the group of nine loudspeakers does not exceed that of one 18in unit and the cost of a replacement in the event of a failure is low. The maximum power input for this group is 10 watts. Even at this low power it is clear that

A SMALL ELECTRONIC ORGAN

(Using the high-fidelity circuits mentioned in Part 2)

SPECIFICATION

MANUAL	*Flute, diapason, viola, oboe, trumpet, 8ft.*
PEDAL	*Sub-bass 16ft.*
ACCESSORIES	*Octave coupler, manual to pedal coupler, manual tremulant, balanced swell pedal.*



from two sides of the pipe only, the effective area being perhaps 30 square feet. The intensity of the sound may not be very great, but it is extremely pervading. This is because such a large quantity of free air is coupled to the resonating surfaces of the pipe. The tone is free and natural. It is not necessarily a matter of power, but simulation. If equal intensity is produced from a properly-loaded free cone of say 18in nominal diameter, the energy is found to be concentrated in far too small an area. The form of pressure wave produced in the surrounding air is quite different from that due to the pipe. The fallaciousness of claiming that an equivalent effect is possible is evident.

Tonally, however, such sounds are not initiated in a simple manner; the building up of the energy in the wood pipe is slow and introduced by a wind noise; then certain overtones of low intensity, leading to the main tone as the resonating tube achieves exact frequency coupling with the lip of the pipe. No electronic instrument can

the whole of the bass must be fed into an amplifying channel separate from the upper registers. It is impossible adequately to combine the frequency range of an organ-like instrument although at lower powers, perhaps 4 to 5 watts, acceptable results can be obtained with bass and treble loudspeakers fed through frequency-dividing networks; this system is liable to introduce intermodulation distortion if the frequency range is too extended. Large resonating chambers in the form of pipes and cubes have also been used with varying degrees of success.

An important aspect is that of volume or expression control. This is a function of the whole system, i.e. generator, amplifier, loudspeaker and room. A fundamental ruling is that the amplified sound must never exceed that of the corresponding physical source in loudness, otherwise coarsening of the tone and other forms of distortion may arise. One hardly ever hears an instrument of the orchestra in an average room. The impressions retained are derived from the concert hall. It is surprising how loud such instruments do sound in the house, and electrical simulations of such tones, even if perfect, would not be agreeable under these conditions. This scale factor is very difficult to adjust, except for the quieter sounds and those of an organesque nature—smooth tones. But caution should be exercised here, for there is a quite recognizable optimum loudness for the sound of any pipe organ conforming to a particular specification; this is too often overlooked.

Assuming that a satisfactory maximum loudness is attained, this must be subject to volume control. All musical instruments alter their harmonic content when caused to sound louder or softer. Such a procedure is impossible with electronic generators, although circuits are available which attenuate the bass to a lesser degree than the treble when the volume is diminished; but in a quite different way from the change in waveform with volume of physical instruments. In the case of the pipe organ, as the wind pressure cannot be varied, some or all of the pipes are enclosed in large boxes with louvred fronts, by means of which more or less sound is allowed to escape. In such instruments changes in loudness are accompanied by very marked changes in tonal texture. When the shutters are closed, the upper harmonics are greatly reduced, the middle ones less, the bass only a little or perhaps not at all; therefore opening the shutters not only increases the loudness but also the brilliance of the tone. It has so far not been possible to duplicate this effect with electronic generators, although skilful manipulation of the tonecolours, playing in certain parts of the compass, choosing ascending passages in which to increase the volume, heightens the æsthetic appeal and reduces any sense of artificiality which may be present in the tone.

While no special manipulative technique is required to play a well-designed electronic instrument, the maximum appeal or even usefulness can only be attained after some experiment. The comparatively weak initial envelope control or attack is partly responsible for this. Mention has been made of such circuits, but there is no known means of simulating the starting tones of orchestral instruments. We cannot imitate the fiery attack of the trombone or the windy attack of the piccolo. The only attack controls available are those of the plucked string or the slow speech of the organ. The reason for this is that there is no definite, exact or fixed way of starting the tone of any instrument under the control of a human being; an infinite variety of shades of speech is possible according to the method of manipulation. Were this not so, an orchestra would be no better than a series of keyboard instruments. So we must be content with the organ type of tonal initiation for most tonecolours, since an uncontrolled or abrupt start to the tone is unreal and irritating except for some forms of rhythmic or agressive playing.

In general, attack is best achieved in valve oscillator instruments by control of the grid or cathode bias. A superior but more costly arrangement is to use logarithmic variable resistance switches operated by the playing keys. In electrostatic instruments the H.T. charging potential may be delayed by an *RC* network; this cannot affect the pitch but may do so with valve oscillators unless special precautions are taken.

The ear tends to favour the slower attack of organ type and the above methods are widely used, in some cases with excellent results; in other instances, the kind of circuit and the time-constants are altogether incorrect and this most vital and important aspect of electrical tone production has not yet received the attention it requires.

There is no doubt, and it has been adequately proved, that if cost of manufacture is of no consequence, extremely satisfying synthesis of many sounds is possible by electrical means. Since there are limits set by the characteristics of existing physical instruments, it is only in the organ world that such costly experiments can be justified; apart from economics, the following reasons support this class of development:

1. Weight and floor area for installation greatly reduced.
2. Generator components proof against temperature changes and can be tropicalized.
3. Running costs low and conversion efficiency high, i.e., acoustic watts radiated for power consumed against acoustic watts compared with blowing fan watts in a pipe organ.
4. Loudspeaking equipment can be situated where most effective.
5. Power, tone, etc., can be regulated to suit acoustic environment actually on site.
6. Dismantling and re-erection costs negligible; it may cost hundreds of pounds to remove a pipe organ.
7. Specification can be altered without difficulty.
8. In some cases, tuning is permanent; in others, easily performed.
9. Servicing possible by radio engineer, often much more readily available than skilled organ hand.

From an economic aspect, all electronic musical instruments fall into the category of communication equipment. They could thus be subject to the same production techniques if the demand were sufficient, since with all circuit assemblies the cost is related to the rate of flow. The quality of the components and in particular the tolerances and ageing techniques, must be well above radio set level. Valves require care in selection and much ingenuity has been expended in obtaining the required results from simple triodes; for these are very reliable. A number of prototype generators using transistors and modern designs of gas tube are being tested, too.

Production in this country is on a small scale, but it is interesting to note that in the United States, where production of most makes runs between 120 and 400 a month, it has not been possible to reduce the price below roughly £1 000 for a full scale two manual and pedal organ. This is almost entirely due to the high quality of the individual components used.

It seems probable that the very largest organs will continue to have pipes, since here the complexity of the radiated energy is beyond the resolving powers of any loudspeaking system which can at present be envisaged; and there are certain well-defined characteristics of large organs which appear impossible of simulation by electrical means; but there is little doubt that smaller instruments of the organ type, and others for the production of orchestral tones of certain kinds, will develop intensively.

It will be observed that these articles began in general terms, but have concluded on an organesque note. This is simply because, from a technical and economic standpoint, it is only the organ type of tone which has made progress. It is evident that real simulation of all the important characteristics of orchestral instruments is impossible, and would always remain so if controlled by means of a keyboard. There is a remaining alternative in the production of new tones. This is a fruitful field and is bound up with new manipulative techniques which could start new fashions in modes of expression. Such an approach might do much to offset the inability to simulate many currently known orchestral instruments. However, it is always dangerous to attempt to run before one can walk, and it should not be forgotten that most physical tone sources have been under development for many years, in some cases hundreds of years. It is unreasonable to expect the same attributes in synthetic instruments within a couple of decades.

Readers interested in a comprehensive source of references to published information on electrical music are recommended to consult the bibliography on electronic musical instruments published by the Tottenham Public Libraries, London, N.17; and for the theory and practical design data, The Electronic Musical Instrument Manual (2nd edition), Sir Isaac Pitman & Sons, London W.C.2.

Control of Thyratrons by Small Signals

By R. Bailey*, B.Sc.

The use of a thyratron to control the power supplied to a resistive heater winding is discussed and it is shown that variation of the phase of the control grid voltage enables the power supplied to be varied continuously over a wide range. When the control voltage applied to the grid is not large compared with the critical grid voltage, the relationships between control signal voltage and the output power are markedly non-linear and the determination of the optimum operating conditions is greatly facilitated by graphical methods. The methods are illustrated by considering the problem of the temperature control of an oven. The effect of variation of valve characteristics are considered and it is concluded that although the use of control signals as small as one or two volts may be permissible if the thyratron forms part of a feedback system, large signals should be used whenever possible.

DURING the development of a small constant temperature oven, it was found convenient to control the power supplied to the heater winding by means of a thyratron which, in turn, was controlled by a signal voltage obtained from a temperature sensitive thermistor bridge[1]. The signal available from the bridge was quite small and to avoid the complication of an additional amplifier, it

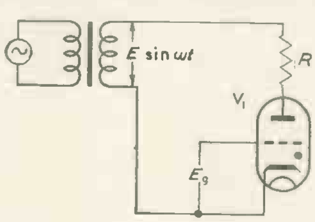
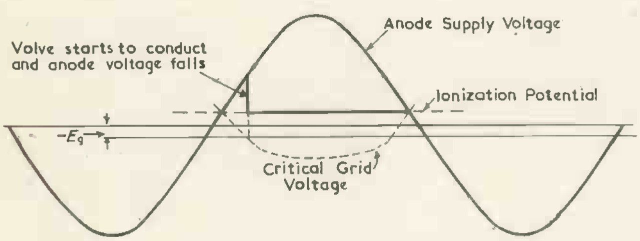


Fig. 1. Basic thyratron circuit

Fig. 2. Control of thyratron firing point using D.C.



was necessary to consider carefully methods whereby one could obtain the maximum possible change of power dissipated in the heater winding for the minimum signal from the temperature sensitive bridge. The graphical methods developed may have other applications where it is necessary to control thyratrons from small signals.

General Principle

Consider the circuit of Fig. 1. The thyratron V_1 has an alternating anode supply voltage $E \sin \omega t$ applied through a resistor R. The valve will conduct only if the anode voltage is sufficiently positive to overcome the effect of the bias on the control grid. Thus corresponding to a given anode voltage there is a voltage known as the critical grid voltage at which the valve will begin to conduct. The valve will continue to conduct until the anode voltage falls below the ionization potential later in the cycle. This is shown diagrammatically in Fig. 2. If the ionization potential were negligible and the control ratio a constant, the critical voltage curve would be half sinusoidal, but due to the non-linear valve characteristics, the curve is somewhat flattened.

Consider now what will happen if the control grid has a fixed negative bias E_g as shown in Fig. 2. When the critical grid voltage exceeds E_g, the valve will conduct, its anode voltage will fall to the ionization potential of the valve and the remainder of the supply voltage will

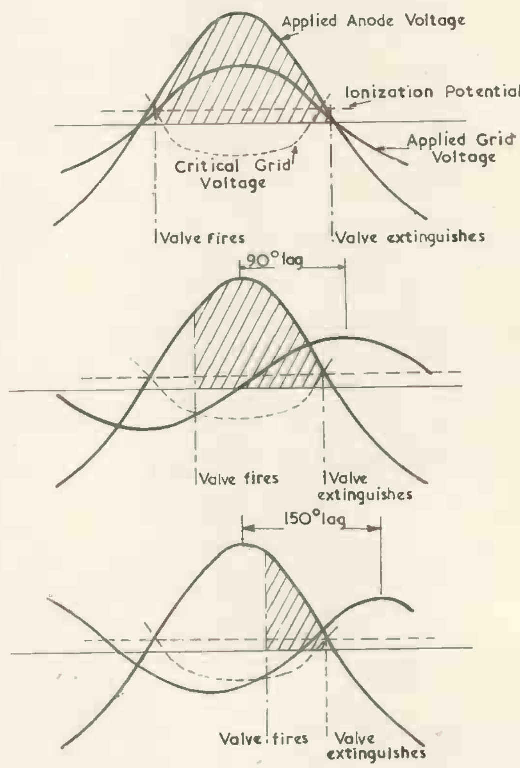


Fig. 3. Control of thyrtron firing point using A.C.

be dropped in the anode load resistor R. If one assumes that the ionization potential is small compared with the supply voltage, a little consideration will show that D.C. bias will allow the conduction angle, i.e., the proportion of the cycle over which the valve conducts, to be varied only between 90° and 180°, giving a change in output power of only two to one. In general, this is insufficient and phase control using A.C. bias on the grid is used.

Fig. 3 shows phase control in greater detail. In Fig. 3(a) the grid voltage is in phase with the anode voltage so that the valve conducts as soon as the anode voltage exceeds the ionization potential. Figs. 3(b) and 3(c) show the grid potential lagging 90° and 150° behind the anode voltage so that the valve conducts only over a part of the positive

* *Associated Electrical Industries Ltd.*

half cycle. It will be seen that by varying the phase of the grid voltage, any conduction angle between 0° and almost 180° can be obtained and, if the amplitude of the A.C. grid bias is large compared with the critical grid voltage, the firing angle is almost independent of the bias voltage and depends only on the phase.

To enable this method of control to be used in conjunction with a temperature sensitive bridge, the output of which is zero at balance, it is necessary to provide a fixed A.C. bias. This is combined with the A.C. output of the bridge, the phase of which differs from the fixed bias by about 90°, so that the resultant of the fixed bias and the output of the bridge has a phase which depends on the amplitude of the bridge voltage.

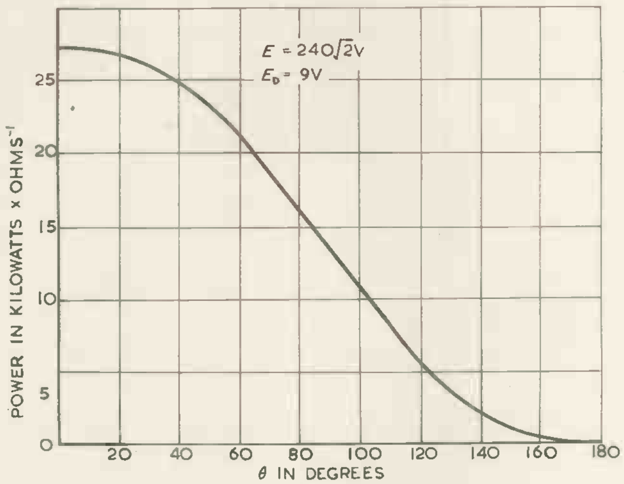


Fig. 4. Variation of power output with delay in firing (*linear plot*)

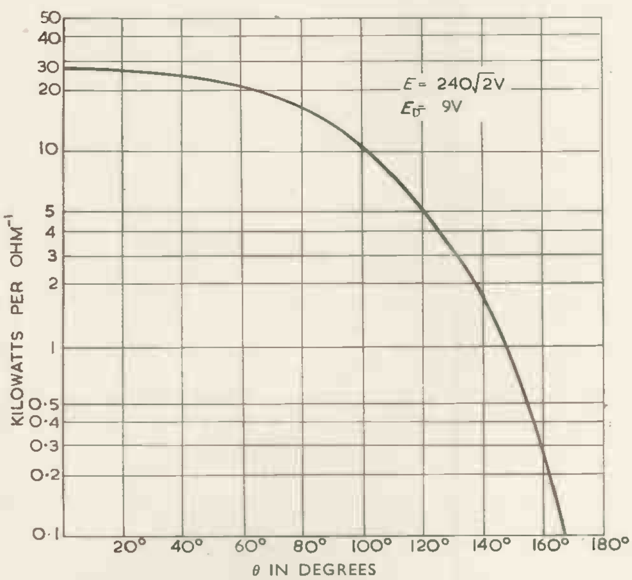


Fig. 5. Variation of power output with delay in firing (*logarithmic plot*)

To obtain a large phase change for a small output from the bridge, it is necessary for the fixed A.C. bias to be small and, under these conditions, the bias is no longer large compared with the critical grid voltage and the amplitude of the grid signal and the exact shape of the critical grid voltage curve affect the firing angle to a marked extent. Because of the non-linear relationships involved, graphical methods are the only practicable ones for determining the optimum bias conditions. In describing these methods, the data applicable to the constant temperature oven will be used as a practical example, but the methods are perfectly general.

Anode Circuit Design

The power dissipated in the resistor R in the anode circuit of the thyratron fed from an alternating voltage source of peak value E volts is given by:

$$\frac{1}{2\pi R}\int_{\theta}^{\pi-\alpha}(E\sin\theta - E_D)^2\, d\theta$$

where E_D is the voltage drop in the valve when conducting; θ is the phase-angle by which the start of conduction is delayed

$$\alpha = \tan^{-1} E_D/D$$

If $E \gg E_D$ this is almost independent of E_D.

The variation of power with θ is plotted for $E = 240\sqrt{2}$ volts, $E_D = 9$ volts in Fig. 4. From this, it will be seen that the useful range of values of θ is about 40 – 140° which gives range of power of 12:1.

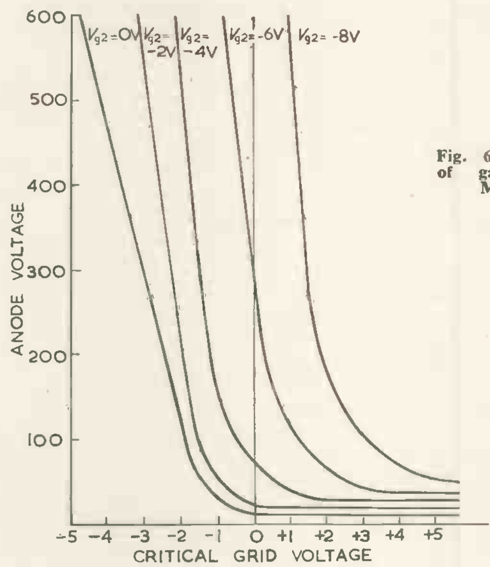


Fig. 6. Characteristics of gas-filled tetrode Mazda 20A2

In some circumstances, it may be desirable to work with conduction angles less than 150° because of the increased sensitivity which may be obtained. An oven will require a certain range of input power and the range of conduction angles required will be determined by the value of load resistor chosen. For example, consider an oven which requires 10 watts to maintain it at 10°C above ambient and 20 watts to maintain it at 20° above ambient. If the resistance of the heater winding is 1 000 ohms, a change of θ from about 63° to 103° (i.e. a change of 40°) would maintain the temperature of the oven constant against a change of 10°C in ambient temperature. If, however, the resistance of the winding were reduced to say 100 ohms, the corresponding values of θ would be 137° and 148° (i.e. a change of only 11° and an improvement in sensitivity of almost four times). This is clearer if Fig. 4 is replotted on logarithmic paper as in Fig. 5. In this curve, the sensitivity is proportional to the slope of the curve which will be seen to increase continuously as $\theta \rightarrow 180°$. An incidental advantage of using large values of θ is that when the oven is first switched on, the thyratron conducts for all the cycle and the input power is very large so that equilibrium is quickly obtained. It is not often that advantage

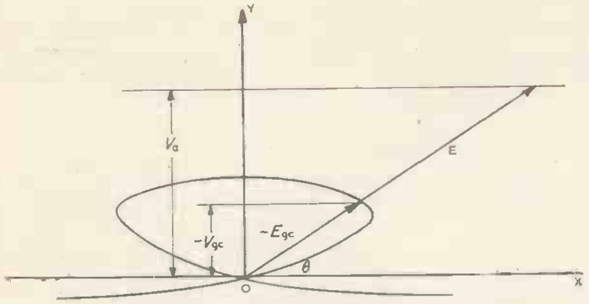


Fig. 7. Vector diagram showing relationship between the vectors and electrode voltages

can be taken of this fact since a low value of R means a high peak current both during normal operation and, more particularly, during the initial warm-up period and, unless the power requirements are very small, the cost of the larger thyratron which would be required more than outweighs any advantage gained.

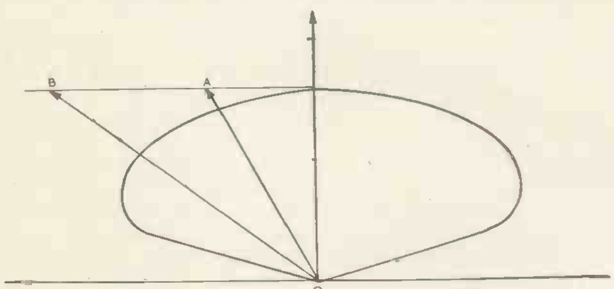
Grid Circuit Design

If the control voltage is small, a high control ratio is obviously desirable. The usual gas tetrode has this with the additional advantage that the control grid current before conduction is much smaller than for a gas triode and is therefore less likely to upset the predicted performance. The makers' control curves for a Mazda 20A2 are shown in Fig. 6, but for design of phase control circuits it is more convenient to replot them on polar paper.

In Fig. 7, let the anode voltage $E \sin \theta$ be represented by a rotating vector of magnitude E. The instantaneous anode voltage V_a may be represented by the projection of the rotating vector in OY. Corresponding to this anode voltage V_a, there will be a critical grid voltage V_{gc} which may be regarded as the projection of another rotating vector which is in anti-phase with the anode voltage. To bring the vector diagrams into the positive quadrants, this will be drawn as a vector $-E_{gc}$ in-phase with the anode voltage. Since the valve characteristics are non-linear, the length of the critical grid vector will change as it rotates. From the known valve characteristics, the locus of the tip of this vector may be drawn as shown.

The design procedure will be illustrated by taking the constant temperature oven as an example. This requires a range of 12:1 in input powers if it is to be unaffected by changes in ambient temperature. The control signal available from a temperature sensitive bridge is limited by the heating of the bridge elements and it is required to determine whether the maximum available signal of some 2V R.M.S. can be made to produce the required change of power, bearing in mind that the valve characteristics will change with both time and heater voltage.

Fig. 8. A vector diagram showing two possible grid voltage vectors which would cause the valve to strike at θ = 90°



Using Figs. 4 and 5, it will be seen that the required 12:1 range of power can be obtained by varying the firing angle θ between 40° and 140°. When the oven is exactly at the designed temperature, there will be no output from the temperature sensitive bridge and the standing bias must hold the conduction angle at 90°. Fig. 8 shows two possible standing bias vectors OA and OB with different angles of lead. The larger the angle of lead, the greater the bias required and the greater the bridge signal required to change the conduction angle a given amount. The larger the standing bias, the less will the circuit be affected by change of valve characteristics. The actual angle of lead chosen is a matter for the designer's judgment but, for present purposes, an angle of 45° will be assumed to be a reasonable compromise. The method of determining the signal voltage required to shift the firing angle to 40° and 140° is shown in Fig. 9. The standing bias vector OA leads the critical bias vector by 45° and its projection is equal to the critical bias at ϕ = 90°. When ϕ = 40°, the OA is in position OA_1. If the valve fires at this point, the control bias added to the standing bias must have a resultant, the projection of which equals the critical bias at ϕ = 40°. Two possible control voltages are shown as

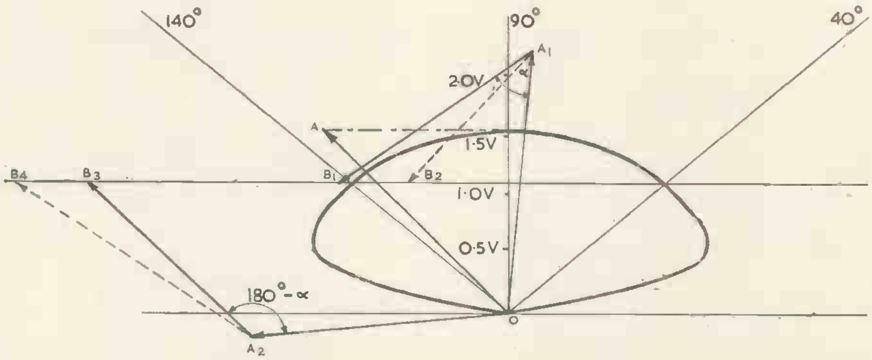


Fig. 9. Complete vector diagram showing two possible control signals to drift firing point from 40° to 140°

A_1B_1 and A_1B_2. Similar considerations apply to ϕ = 140° except that the phase of the control voltage is no longer arbitrary. Corresponding to vectors A_1B_1 and A_1B_2 are vectors A_2B_3 and A_2B_4 such that $\angle OA_1B_1 = 108° - \angle OA_2B_2$ and $\angle OA_1B_2 = 180° - \angle OA_2B_4$. An examination of Fig. 9 will show that there is only one angle OA_1B_1 for which $A_1B_1 = A_2B_3$ and this is the one usually used, although in some cases the non-linear control produced by using a different value for this angle may have advantages.

Effect of D.C. Bias

From Fig. 8, it will be seen that the magnitude of the control voltage required is determined largely by the magnitude of the standing A.C. bias and it is, therefore, obvious that the more this can be reduced, the less control voltage will be required. Examination of Fig. 6 indicates that the effect of D.C. bias on the shield grid is to shift the curves bodily sideways and is, therefore, very similar to D.C. bias applied directly to the control grid. Both can be dealt with in the same fashion. Fig. 10 shows a family of critical grid voltage curves plotted on polar paper and from these it will be seen that a bias of between −3 and −5V on the shield grid is a useful working range.

Effect of Grid Current

The effect of grid current is very difficult to calculate in detail, but if the following points are noted, a qualitative estimate of its effect may be made.

(1) Before firing the control grid current is small (usually less than 10^{-7}A), so that unless the grid leak is more than 1MΩ, its effect may be neglected.

(2) Unless limited by a series resistor, the control grid current after firing may be substantial (several milliamperes). It will be positive or negative depending on the sign of the grid voltage and it will also be influenced by the value of anode current. It is, therefore, extremely difficult to predict its mean value over a complete cycle.

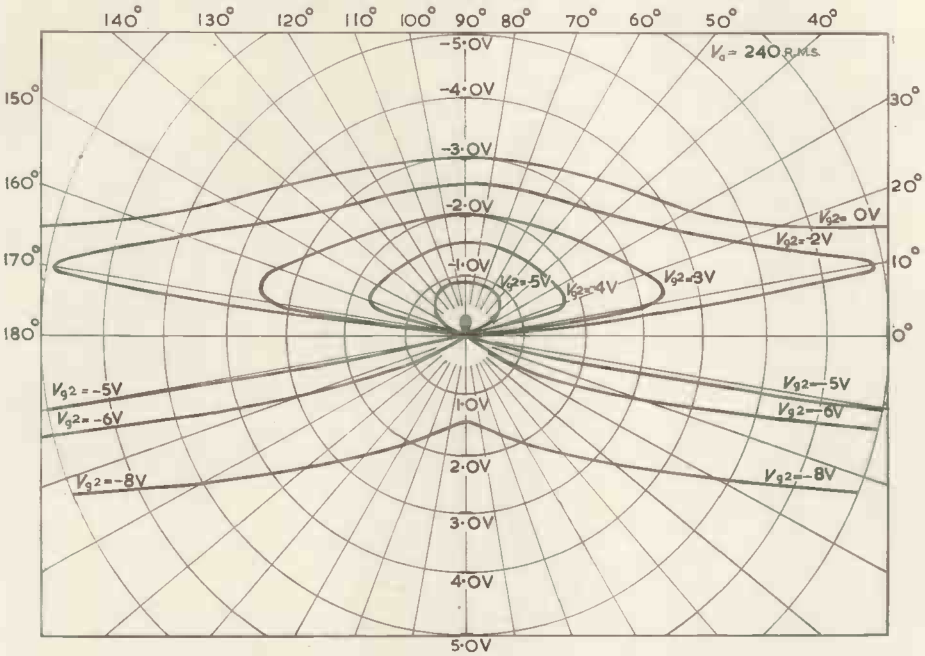


Fig. 10. Critical grid characteristics of Mazda 20A2 plotted in polar co-ordinates

(3) When non-conducting, the shield grid current resembles that of an ordinary valve and is fairly small (0·1mA at $V_{g2} = 0$) decreasing as the shield grid is made more positive. After striking, this current may also be several milliamperes and is influenced both by anode current and shield grid voltage.

If there are capacitors in either grid circuit, they will be charged by the grid current during conduction and alter the D.C. bias. This may either increase or decrease the sensitivity depending on the circuit arrangements and firing angle. Since these effects can not be predicted, capacitors in the grid circuits should be avoided if possible.

Effect of Heater Voltage and Valve Variation

If valve manufacturers provide limiting characteristics, the effect of valve changes can easily be estimated by drawing polar curves corresponding to the limiting conditions, but these are seldom available. Checks on a small batch of 20A2's showed a scatter of about ∓0·2V in the critical grid voltage under identical conditions. A change of 10 per cent in heater voltage changed the critical grid voltage by about 0·1V.

Conclusion

The graphical methods developed above facilitate the design of thyratron circuits working with small control signals. It is quite feasible to design a circuit which will change the conduction angle from 40° to 140° with a signal of less than 2V R.M.S.

The use of such circuits depends on the stability of valve characteristics and should preferably be used only when they form parts of feedback systems. Provided that the static bias be adjusted for the particular valve in use, it has proved practicable on a thyratron with signals obtained directly from a temperature sensitive bridge and for the system to work satisfactorily in spite of changes in the heater voltage of the valve.

Acknowledgments

The author wishes to thank his colleagues for helpful discussions and Dr. T. E. Allibone, F.R.S., for permission to publish this paper.

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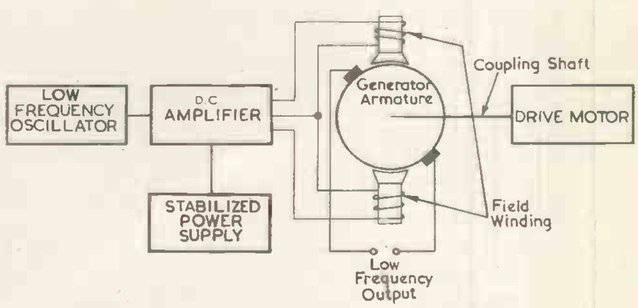
A Low Frequency Generator for Vibration Testing*

Vibrations for testing structures are conveniently produced with the aid of a moving-coil exciter. This is a large and powerful type of loudspeaker in which the cone is replaced by a rod which imparts mechanical motion to a structure to cause it to vibrate. This type of exciter requires a large sinusoidal energizing current, normally supplied by a high power amplifier. This means of supplying the current is not suitable for frequencies below 20c/s owing to difficulties encountered in the construction of the output transformer of the amplifier.

In the system to be described, the required current is obtained by modulating the field current of a D.C. generator at a low frequency. The low frequency supply is obtained from an *RC* oscillator and is applied to the field winding via a push-pull D.C. amplifier, as shown in the block diagram, the field winding being centre tapped. The output current, which is taken from the brushes of the generator, can be varied by rotating the armature at different speeds or, more easily, by altering the output of the D.C. amplifier; the output frequency is that of the oscillator.



The low frequency generator described

**Communication from E.M.I. Engineering Development Limited.*

The Vapotron

By G. Ashdown, A.I.Br., A.M.I.E.E., F.Inst.P.

The cooling of high-power valves by vaporization is shown to be a practical proposition. The advantages over circulatory water cooling or forced draught air cooling include higher anode dissipation (two to three times), reduced length of insulating sections and ease of valve replacement. Also, the heat extracted, being at a higher temperature, is of greater value and a supply of distilled water is readily obtainable.

THE power available from high frequency transmitting valves is limited, to a large extent, by the temperature rise of the anode. The method of cooling the anode therefore plays a very important part and the ancillary equipment to do this can represent a large proportion of the capital cost of the installation. Current practice in anode cooling is to use either forced draught air cooling or a water cooling system in which a definite flow is maintained by pumps, the water being raised some 40°C and passed through a suitable radiator to be cooled before returning to the anode.

Fig. 1. A typical Vapotron

At a recent electronics symposium held at Liège in Belgium, the French Thomson Houston Company gave details of a range of valves termed Vapotrons* so designed that the water in contact with the anode is raised to 100°C, circulates in the form of steam and loses its heat by condensation. A typical valve is shown in Fig. 1.

The complete system is known as the Vapodyne and is shown in schematic form in Fig. 2. The Vapotron, with its anode immersed in water, can be seen on the extreme left of this figure. Two heat exchangers, the first water- and the second air-cooled, condense the steam which then, in the form of water, returns by gravity to the valve. A simple two-stage alarm ensures correct water level.

Theory of Operation

The amount of heat dissipated by a metallic surface by evaporation of a liquid at atmospheric pressure increases as the difference between the surface temperature and the boiling point of the liquid is increased up to a critical "calefaction" temperature[1] which for water is 25°C corresponding to 135 watts/sq.cm. Any further increase in temperature difference produces a sharp decline in the amount of heat transmitted or, to use an expression more suited to the application in mind, any increase in the watts/sq.cm above the critical 135 will cause a rapid unlimited rise in surface temperature. It is therefore essential, as a first requirement, that the average power dissipation per unit surface area be kept well below the critical value under all conditions of operation.

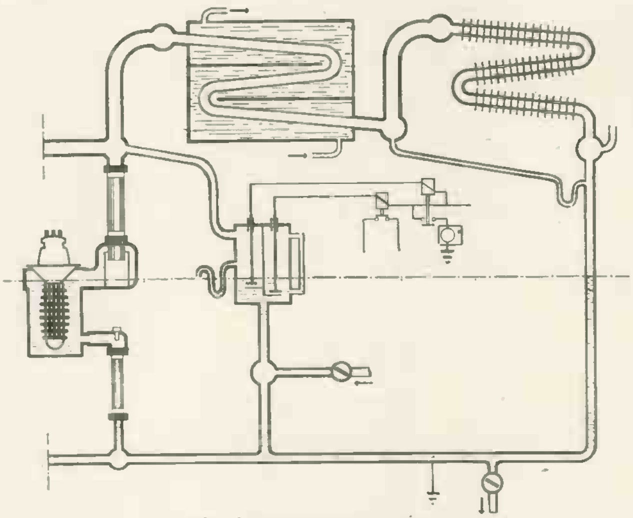


Fig. 2. The complete system

A further requirement is the prevention of hot spots which are liable to cause calefaction. From this point of view, the thickness of the anode wall is of considerable importance. As a thick wall will cause a temperature gradiant resulting in a higher temperature at the internal surface of the anode, it would appear desirable to have as thin a wall as possible especially if the valve is subjected to short-period overloads. The elimination of hot spots is, however, of far greater importance because once the critical temperature difference of 25°C is exceeded at any particular point, cooling by evaporation virtually ceases and the spot can only be cooled by conduction within the mass of the anode. If this conduction does not take place, calefaction will eventually spread over the whole surface and put the system out of action. By using a high thermal conductivity material like copper and by providing sufficient thickness, temperature differences on the surface are short-circuited and the danger of a serious local temperature rise is avoided.

Practical Results

The first Vapotron tests[2] were carried out with an anode having a smooth outer surface; at 84 watts/sq.cm calefaction set in. The smooth surface was then replaced by heavy milled teeth and it was found that calefaction did not take place, although the temperature of some spots

* *Registered trade mark, patents pending.*

exceeded the critical 123°C. Finally, by staggering the teeth, as shown in Fig. 1, a better circulation of the water-steam mixture was obtained, the hot spots disappeared and a further decrease in the average surface temperature was observed.

The curves of Fig. 3 show the various surface temperatures obtained with a TH445 valve, one of the four types being manufactured in France, for different values of anode dissipation. From these curves can be noted the very large overload capacity of the valve, the rating of which is fixed at 60kW, and the increased output obtainable over a valve of similar size using conventional water-cooling (2:1) or forced draught air-cooling (3:1).

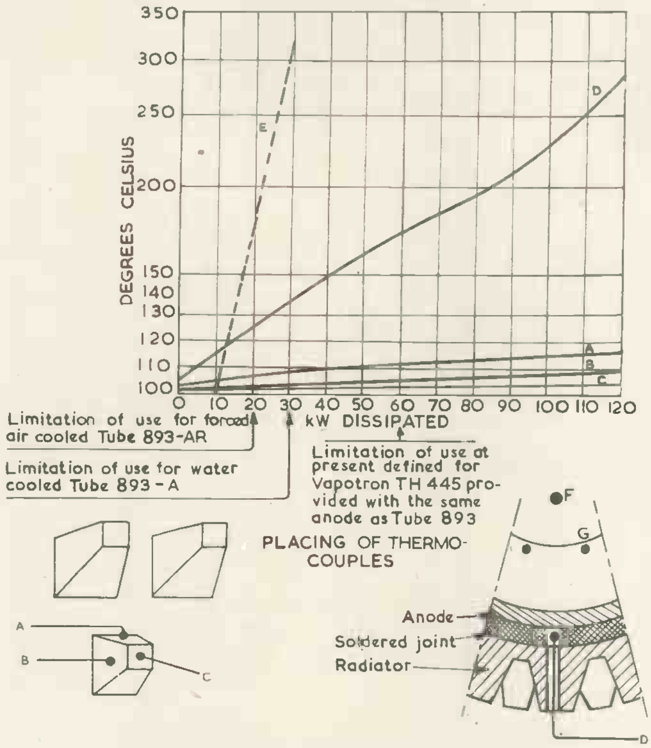


Fig. 3. Overload test in static operation (13kV D.C.) of Vapotron type TH445

Curve A—*Temperature at hottest point on external surface of radiator.*
Curve B—*Temperature in copper on the side of a tooth, i.e., approximate average temperature of useful part of radiator.*
Curve C—*Temperature at tip of a tooth.*
Curve D—*Temperature in joint at contact of tube anode, at hottest point.*
Curve E—*Temperature at the base of the teeth of a forced air cooled tube.*

Five Vapotrons have been installed in the Paris-Villebon transmitter[3], which started operation in 1951, three in the R.F. power stage and two as modulators. It can be seen from Fig. 4, which shows these valves installed, that the Vapotrons are held in place on the top of their water containers by their own weight and that the insulating sections in the cooling circuit are considerably shorter than those required with conventional water-cooling. This is made possible by the higher resistivity of steam, even if wet, compared with water. In the return circuit, less visible on the figure, shorter sections can also be used because the very low rate of flow (approximately 20gal/hour for 60kW) permits the use of a small bore pipe. Satisfactory operation up to 50kV peak has been obtained with 18in of pyrex tube on the outlet side and 12in in the return circuit, the latter tube having a cross sectional area of little more than 0·15sq.in.

Characteristics

Two of the valves at present being manufactured are the TH445 rated at 60kW and the TH456 rated at 40kW.

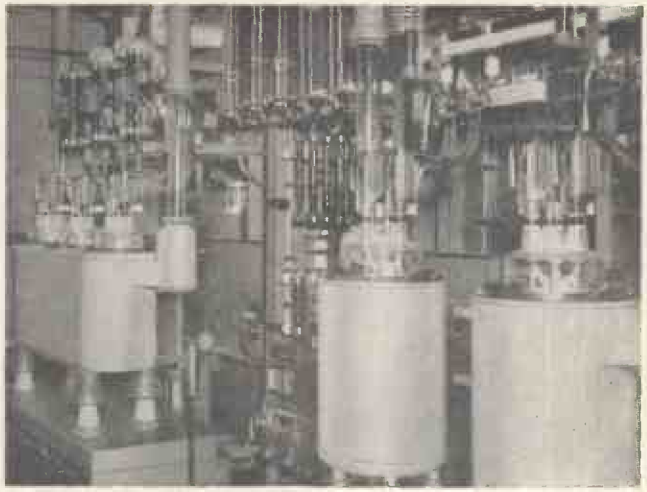


Fig. 4. A typical installation

The larger of the two valves has an overall length of approximately 21½in and weighs 31lb, and the smaller, a little over 17in and weighs 28½lb; the width over the handles is in both cases 11¾in.

The electrical characteristics are given in Table 1.

TABLE 1
Characteristics of Two Vapotrons

	TH 445	TH 456
Cathode	Tungsten	Tungsten
Filament voltage	15V	12·6V
Filament current	290A	300A
Interelectrode capacitances :		
Grid-filament	61pF	40pF
Anode-filament	6pF	4pF
Grid-anode	31pF	30pF
Amplification factor	22	12
Mutual conductance	18mA/V	12ma/V
Maximum ratings :		
D.C. plate voltage	15kV	12kV
D.C. grid voltage	−1·2kV	−1·5kV
Peak cathode current	30A	25A
Plate input	90kW	60kW
Plate dissipation	60kW	40kW
Plate output	60kW	40kW
Grid dissipation	700W	500W
Frequency at maximum ratings	10Mc/s	10Mc/s

By-products

In addition to the advantages of increased power output, simpler installation, rapid replacement of valves and quietness of operation, there are two interesting by-products: distilled water and a supply of heat at a high temperature. The latter, which is obtained automatically from the first heat-exchanger, amounts to more than 200gal/hour at 90°C for a 100kW transmitter such as the Paris-Villebon station. Distilled water is obtained by connecting the valve container to the water main through a long insulating section and a water softener and collecting the distilled water at a point beyond the condenser.

Acknowledgments

The author is indebted to the Compagnie Française Thomson Houston for permission to publish the technical details. He also wishes to acknowledge valuable assistance from M. Beurtheret, Chief Engineer C.F.T.H. and from the Edison Swan Electric Company with whom C.F.T.H. are associated.

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The Design of Electromagnets

(Part 1)

By L. R. Blake*, B.Sc., Ph.D., A.M.I.E.E.

An attempt has been made to present in this article data useful when designing magnet systems —D.C. magnets in particular, but also A.C. magnets and permanent magnets. The data included consists of fringing curves, showing how the field varies between the poles of a magnet; curves to assist in permeance estimates; magnetization curves of D.C. magnet materials; and heat dissipation curves. The principles of design of an electromagnet are outlined and various winding and magnetic circuit arrangements are examined and criticized. Various methods of estimating leakage flux are described, including the use of analogues, such as the electrolytic bath, and the use of scale models.

A D.C. magnet is one of the simplest electromagnetic devices, and it is not surprising that almost anyone with a background in electrical engineering is prepared to tackle its design with confidence. There are, however, more subtleties in connexion with magnet design than are at first imagined, and it is the source of much regret, especially with large and costly magnets, when these are appreciated too late. It is hoped that it will be of assistance, particularly to those who are not specialists in this class of work, to draw attention to some of these details, and to present in convenient form the more important design data.

Electromagnets are required for two main purposes: to provide a magnetic field over a certain restricted volume, or to provide a tractive force. Magnets of the second type deserve special attention† and it is not intended to include the special problems associated with these magnets within the scope of this paper.

Specification of the Magnet

The most usual method of specifying a magnet to provide a magnetic field is by saying that the field is required uniform to $\pm x$ per cent over a certain region of space (say the region a', b', g' in Fig. 1). If curves giving the variation of the field within standard pole configurations are readily to hand, it is possible to estimate the actual pole size (a, b) and the gap distance (g), but before this can be done it is necessary to know the rough outlines of the poles. It will be seen that it is necessary to proceed by successive approximation—a method necessary in varying degree throughout the design of the magnet.

Curves showing how the field varies between poles of various shapes are given in Figs. 12 to 16 at the end of the article, most of these curves being deduced using an electrolytic bath. For convenient reference, all the data useful in the design of a magnet has been separated from the text and put at the end. The field fringing curves which are given, show how the field varies in the mid-plane between the poles (Figs. 12 and 13) and how it varies across the gap, going from one pole to the other; curves are given showing the field variation in this direction at the edge of a pole (Figs. 14 and 15), and at the pole axis (Fig. 16).

When the pole dimensions (a and b, for example) are large in comparison with the gap distance g, and especially with straight sided poles, the estimation of the fringing of the field is a two dimensional problem and the curves of Figs. 12 and 14 can be used. When the pole dimensions are of the same order as the gap distance, then the curves of Figs. 13, 15 and 16, which give the variation of the field between cylindrical poles, will be helpful.

The fringing curves will rarely be directly applicable, but reasonable accuracy should be achieved in most instances if the information is applied carefully, appropriately modified as common sense dictates.

The final size of the magnet is, of course, closely dependent on the size of the poles themselves, so an accurate estimation of the minimum pole size tolerable using the fringing curves is an important part of the design. When this has been completed, one can proceed on the

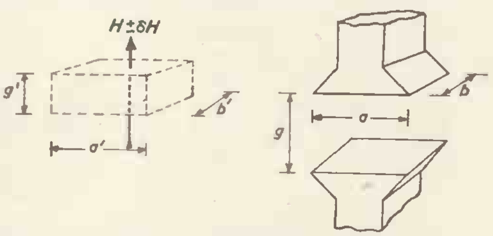


Fig. 1. How the specification of the field required is related to the actual pole dimensions

assumption that one knows the pole dimensions, the gap distance, the field in the gap and the variation of the field across the gap—at least in the centre of the poles—so that the mean gap field $H_m = (1/g) \int H\, dg$ is known.

Basis of Design

Consider an example in which the magnet is of the simple horseshoe type; other types of magnetic circuit will be considered later. The most important quantity to deduce —the total ampere turns required—can be obtained by the direct application of Maxwell's first equation:

$$\oint_c H \,.\, dl = 4\pi/10\, I_t \quad \ldots\ldots\ldots\ldots (1)$$

that is, the line integral of H around any complete circuit c is equal to $4\pi/10$ times the total current embraced by that circuit. I_t—the total current (or rather ampere turns) embraced by the circuit—is, of course:

$$I_t = \oint_s J \,.\, dS$$

where J is the current density in the direction normal to the surface, and the surface integral is taken over that surface of which the circuit c is the periphery. Equation (1) in its most fundamental form is, therefore,

$$\oint_c H \,.\, dl = 4\pi/10 \oint_s J \,.\, dS \quad \ldots\ldots\ldots (2)$$

To determine the total ampere-turns necessary, the most convenient circuit over which to integrate is around the

* The British Thomson-Houston Co., Ltd.

† For an excellent and thorough account of this type of electromagnet, see ROTERS, H. C., Electromagnetic Devices (Wiley, 1941).

magnetic circuit and across the gap near the centre of the poles, since this is the region where the field is most uniform and is known with greatest accuracy. Then:

$$(NI)_t = 10/4\pi \left(\int H \, . \, dg + \int H_i \, . \, dl\right)$$

Since from the previous section we now have a reasonably accurate knowledge of the field and how it varies across the gap, the first integral, $\int H \, . \, dg$, is simply $H_m g$.

The second integral, $\int H_i \, . \, dl$, taken around the iron circuit, can be evaluated provided that the flux density B_i is known everywhere in the iron and provided that the magnetization curve of the iron is available. An object of the design should be to shape the iron circuit so that B_i is everywhere constant, the full realization of this being prevented mainly by practical difficulties of manufacture and expense. It is therefore permissible to simplify this integral by assuming that B_i is everywhere at the design value so that H_i is known and is also constant. Hence we can write:

$$NI_t = 10/4\pi \, (H_m \, g + H_i l) \quad \ldots\ldots\ldots \quad (3)$$

where l is the mean path length in magnetic material, this equation being accurate and not normally requiring later modification.

The next question is: how large should B_i be so that $H_i l$ can be determined and NI_t can be calculated? Since the magnetizing coils have to be wound round the iron circuit, it is desirable to keep the iron section to a minimum everywhere, for then the weight of the iron, the weight of the copper and the ohmic losses in the copper are kept low. The copper losses are:

$$I^2R = \rho \, J^2 \, . \, \text{volume of copper},$$
$$= \rho \, J^2 A_c l_c \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

where J is the current density,

ρ is the resistivity of the copper at the working temperature

A_c is the net copper area and

l_c is the mean length of turn.

If J and A_c are fixed, increasing the iron section increases l_c and hence the losses. However, if the iron section is made too small and the flux density high, the term $H_i l$ will become excessive, leading to a large increase in the total applied ampere turns. Since H_i rises more and more rapidly with B_i as saturation is approached, it is essential to keep $H_i l$ to a reasonably low proportion of $H_m g$, as a slight increase in B_i can greatly increase NI_t without much gain from the small section. The calculation of the ideal value of B_i leading to a minimum overall weight or minimum copper losses is not normally worth the trouble; a reasonable estimate for B_i is that value which makes $H_i l$ from 5 to 20 per cent of $H_m g$.

Having selected the value for B_i the real problem of a magnet design begins: namely, to ensure that the iron sections are adjusted to achieve this density. Before considering this, it is necessary to complete the design of a magnetizing coil so that the window area required is known, and with it the general shape of the magnetic circuit.

Coil Design

The current density employed in the magnetizing coil is controlled by:—

(1) The power supply available; if this is limited the current density J may have to be low and the copper volume large.

(2) The method of heat dissipation.

The maximum temperature of the coils must be limited to a value depending on the type of insulation: typical figures are 70 to 110°C for the normal range of insulation, and up to 180°C for special insulation such as glass cloth with silicone varnish. The cooling must be such that copper losses do not raise the winding temperature above the prescribed limit. For example, with natural air cooling the copper loss per unit surface area must usually be less than 0·4 to 0·5 watts/sq.in.; with forced air cooling this figure can be increased and oil or water cooling improve the figure still further. Useful heat dissipation curves* are given in Figs. 19 to 21. As coil design is quite a standard matter this part of the magnet design will not be considered further, except to sum up by saying that the total magnetizing ampere-turns required of the coil now takes the simple form:

$$NI_t = 10/4\pi \, kH_m g$$

where k is a factor which takes into account the M.M.F. drop in the iron circuit, and is selected to have a value usually between 1·05 and 1·20. The copper losses are given by Equation (3), and can be evaluated as soon as J is selected.

Disposition of the Magnet Coils

Before passing on to the major problem of determining the iron sections, consideration will be given to the type of magnetic circuit to be employed and the disposition of the coils. Both of these factors can be considered from the viewpoint of how leakage is affected, for the lower the leakage flux, the smaller the iron section, which brings with it the attendant advantages previously outlined.

Consider first of all the disposition of the magnet coils. If we apply Equation (2) to the a circuit in Fig. 2(a), where $a_1a_2a_3$ is an imaginary flux line, we see that only a small net current is embraced (equivalent in magnitude to the

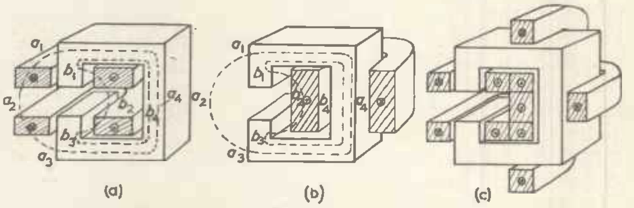


Fig. 2. Winding arrangements

ampere turns external to $a_1a_2a_3$), since the current in the left-hand section cancels much of that in the right-hand section. In Fig. 2(b) this is not so; here a circuit a embraces the full ampere turns, so in this case there will be more leakage flux of this type. For leakage on the inside faces as in the b circuits, it will be seen by a similar argument that there is little to choose between the two arrangements. The coil disposition of Fig. 2(a) is undoubtedly superior to that of Fig. 2(b) from the viewpoint of decreased leakage flux; another advantage of locating the coils near the gap is that the iron section is smallest in this position. However, localized coils wherever they are placed are often inferior to an arrangement as in Fig. 2(c) where the coils are distributed around the entire magnetic circuit, since this enables the mean length of turn to be a minimum, with the consequent low copper losses.

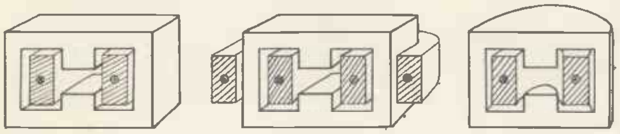
Magnetic Circuit Arrangements

An alternative to the horseshoe magnet of Fig. 2 is the arrangement of Fig. 3(a) or 3(b), which is virtually two horseshoe magnets or " C " type cores together. Different methods of disposing the windings are shown in Fig. 3(a) and Fig. 3(b). A further arrangement having complete cylindrical symmetry is shown in Fig. 3(c).

These arrangements are often inferior to the single "C" circuit, due to increased leakage flux, since minimum leakage flux is achieved when the iron circuit contains the maximum cross-sectional area per unit surface area. The arrangement of Fig. 3(c), although apparently ideal due to

* *From data compiled by F. Fitchett, Plant Engineers, BTH Co. Ltd.*

the small mean length of iron, is particularly bad in this respect, but whether it is better or worse than 3(a) depends on relative dimensions, especially on the ratio of the window width to the gap length.



(a) (b) (c)

Fig. 3. Magnetic circuit arrangements

Quite apart from leakage consideration, the arrangement of Fig. 3(b) is very poor since it makes for large copper losses and excessive weight. This is because in a parallel magnetic circuit arrangement both coils must each supply the same M.M.F. or ampere turns as the one coil in arrangement Fig. 3(a), yet the mean length of turn of either of these coils is not much smaller than that of the single coil.

Arrangement 3(c) has the advantage that the field in the gap is symmetrical and 3(a) and 3(b) also have a slight advantage in this respect. For very large magnets the arrangement of Figs. 3(a) and (c) may be preferred on the grounds of greater rigidity and mechanical strength, but normally the arrangements of Fig. 2 are to be preferred.

It is also worth while to consider the shape of the iron section, especially in terms of the arrangement of Fig. 2(c). Again the guiding principle is that of arranging for minimum leakage flux, hence the maximum cross-section area is required per unit of surface area. This leads naturally to a circular section which also has the further advantage over a square or rectangular arrangement in that there is less surface area on the inner faces where the leakage is greatest. On the whole, however, a circular arrangement is not very convenient for it is difficult to build up the cross-sectional area in the magnetic circuit as the distance from the gap increases, so a square or rectangular section is generally used. A convenient arrangement is to keep the width of the inner face of the iron circuit fixed and increase the cross-sectional area by increasing the depth of the iron circuit, again the aim should be to keep the average section as close to the ideal as possible: that is a square section or perhaps a slightly rectangular section with the smaller face on the inside, for to make the section too rectangular is of little assistance since the reduced leakage on the inner faces is offset by the increased leakage from the sides, and at the same time it leads to an excessive weight of iron.

Core Material

In D.C. magnet systems, the high permeability nickel iron alloys are desirable only if the gap is infinitesimal or if it is necessary to keep $\int H_i dl$ very low: in general, due to their low saturation flux densities, they are quite undesirable. The most common core material is reasonably pure iron with less than 0·1 per cent carbon, but ordinary mild steel is often suitable. As there is no close control on the impurities in mild steel the magnetic properties are very variable, so there is the danger that the permeability may turn out to be much lower than the design figure.

The ideal material should possess a high permeability at a high flux density. By far the best material in this respect is a cobalt-iron alloy in which the proportion of cobalt is 30 to 50 per cent. The highest saturation density is achieved with 34·5 per cent cobalt, for which the figure for the saturation ferric induction (*BH*) is 24·3 kilogauss. Two common cobalt-iron alloys sold commercially are known as Hiperco—a 35 per cent cobalt alloy, and Permendur—a 50 per cent cobalt-iron alloy. Permendur has a slightly lower saturation of 24·0 kilogauss. The resistivity of the simple binary alloys of iron and cobalt is low—about 7 to 12 microm/cm—so additional elements are added to improve the resistivity; chromium, vanadium and molybdenum are very useful in this respect. Hiperco has 0·5 per cent chromium which raises the resistivity to 20 microhm/cm at room temperature. The 50 per cent cobalt alloy (Permendur), which has a resistivity of 7 microhm/cm, has the best magnetic properties at high levels of induction, but an alternative alloy with 1·8 to 2 per cent vanadium is available and is known commercially as V-Permendur; this alloy has slightly poorer magnetic properties than the binary alloy, but the resistivity is much improved, being 26 microhm/cm. The addition of vanadium also improves the mechanical properties, making the alloy easier to fabricate and roll. V-Permendur is available rolled to 0·01in. strip and is useful also for A.C. applications.

Magnetization curves of the materials commonly used in D.C. magnets are given in Fig. 11. All the curves given must be regarded as approximate since impurities, state of strain, and heat treatment have a great effect on the magnetic properties. The difference in the magnetic properties of Permendur and V-Permendur is not intrinsically as great as the curves in Fig. 11 indicate. The V-Permendur curve is from a manufacturer's publication* and gives the average value to be expected in the commercial product: it should be possible with pure V-Permendur appropriately heat treated, to obtain magnetic properties as good as that of the Permendur curve.

Wide variations are to be expected in the case of iron, ranging from the curve for 99·99 per cent pure iron to poorer than the curves given for mild steel. The purest iron available commercially in quantity is Armco iron or Swedish iron and, for D.C. magnets, this material or low carbon steel with less than 0·1 per cent carbon is normally specified. Although the magnetic properties of these materials are much inferior to those of the Permendur type, the cost is much in their favour—Permendur is very expensive (about 12s. to 34s. per lb, depending on the quantity and whether the material is purchased in ingot form or in thin sheets). Very high maximum permeabilities are obtainable with very pure iron, but this is generally of little advantage since the high permeability comes at too low a flux density: Permendur is to be preferred to electrolytically pure iron in almost all respects.

To obtain good magnetic properties at high inductions, it is advisable to anneal the core material at the fairly low temperature of 800°C for the iron-cobalt alloys, and 800° to 950°C for iron, and to cool slowly at about 5°C per

* *Standard Telephones & Cables Ltd. publication, B/MB1.*

TABLE 1

Properties of commonly used materials

MATERIAL	COMPOSITION	SATURATION FERRIC INDUCTION (BH) (KILOGAUSS)	COERCIVE FORCE (oersteds)	ELECTRICAL RESISTIVITY (μ Ω-cm.)	RELATIVE DENSITY
Mild steel	<0·25C, 0·7Mn	21·2	2 approx.	10	7·8
Low carbon steel	<0·1C	21·5	1 „	10	7·88
Armco iron	0·1 impurities	21·5	0·05	10	7·88
Permendur	50 Co	24·0	2·0	7	8·3
V-Permendur	1·8V, 49Co	24·0	2·0	26	8·2
Hiperco	0·5 Cr, 35 Co	24·2	1·0	20	8·0

minute. The annealing should be as long as possible—at least 1 hour per inch of section at the thickest point. Heat treatment in a magnetic field appreciably improves the magnetic properties of the commercially available iron-cobalt alloys.

Useful properties of the materials most commonly used in D.C. magnets are summarised in Table 1.

The Magnetostatic Problem

The main difficulty in magnet design is in assessing the total flux at any point in the iron circuit. For most practical arrangements, the mathematical problem is virtually insoluble. Nevertheless, a mathematical statement of the problem is useful, if only to assist in discovering analogue solutions. The first equation which must be satisfied has already been given in Equation (2); it is:

$$\oint_c Hdl = 4\pi/10 \int_s JdS \text{ in integral form.}$$

The second equation to be satisfied is:

$$\int_s BdS = 0 \text{ in integral form} \quad \ldots\ldots\ldots\ldots (5)$$

Equation (2) says that on completing a circuit the gain in M.M.F. is proportional to the total current embraced; equation (5) says that flux is not created, or flux lines are continuous. The boundary conditions which are applicable are:

$$\text{and} \quad \left.\begin{aligned} H_{\mathrm{i(par)}} &= H_{\mathrm{air(par)}} \\ B_{\mathrm{i(perp)}} &= H_{\mathrm{air(perp)}} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots (6)$$

It is possible to determine the mathematical solution by relaxation or other methods, but these methods are exceedingly tedious, even with apparently simple problems, and they will not be considered further.

Three practical methods of solving the problem are:

(1) Intelligent estimating, either by estimating the permeance of component paths and the M.M.F. which are operative, by guesswork based on experience, or by breaking down the magnetic circuit into a number of standard configurations, the permeance of which are known, the effective operative M.M.F. again being guessed.

(2) The use of analogue methods such as electrolytic baths, resistor networks, etc., or

(3) The use of models.

Intelligent Estimating

The most simple method of determining the permeance of the leakage paths is to divide the magnetic circuit into different regions—for example, the regions numbered 1 to 7 in Fig. 4—then to draw flux lines which may be regarded as bounding the permeance path between similarly numbered regions, and then to estimate the mean area and mean length of these different paths. For example, for the arrangement of Fig. 4 we may write:

$$\text{Permeance of gap portion } P_1 = \frac{1{\cdot}2\,ab}{g}$$

$$\text{Permeance of sides } \quad P_2 = \frac{4(2ac)}{2{\cdot}5(g+c)}$$

$$\text{Permeance of sides } \quad P_3 = \frac{4\,bc}{2{\cdot}5(g+c)}, \text{ etc.}$$

The numbers 1·2, 4, 2·5, etc., are estimated factors.

Similarly, the mean effective M.M.F. operative over the different regions may be determined by estimating the position of the mean line of flux and calculating the current embraced by it and the iron. For example, $\Omega_1 = \Omega_2 = \Omega_3 = \Omega$, $\Omega_4 = 0{\cdot}9\Omega$; $\Omega_5 = 0{\cdot}6\Omega$, etc., where the factors 1, 0·9, 0·6, etc., are again estimated. The leakage fluxes $\phi_1 = \Omega_1 P_1$, $\phi_2 = \Omega_2 P_2$, etc., can be determined and the flux at any section deduced by summing the contribution from the appropriate components. This simple method if used alone is not accurate and so it is necessary to be fairly generous when deciding the size of the iron sections. If the estimates are corrected by tests on a model as described later, this method can prove very simple and is then not to be despised.

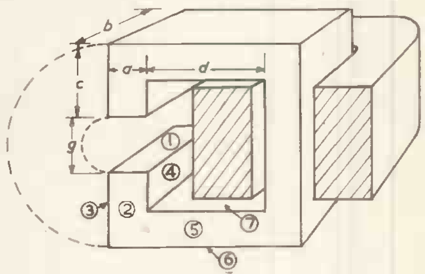
Another method of estimating the permeance of the component leakage paths, in which guessing does not play so important a part, is based on knowing the permeance of a number of standard configurations, which may be selected as similar to various components of the circuit. Again the effective M.M.F. must be estimated. In using this method the curves given in Figs. 17 and 18 will be found helpful.

The curve (a) in Fig. 17 was given by Carter in his classical paper in *J.Instn.Elect.Engrs*, 1926. It shows the increase in effective pole width (the poles being of infinite depth perpendicular to the paper and infinite length outwards from the gap) to account for the flux passing between the actual poles up to a position x from the pole edge. The field configuration in the equivalent hypothetical ideal pole is assumed to be uniform in the gap and of value Ω/g and zero outside the gap. When $x/g > 2{\cdot}5$, the actual fringing field takes the form:

$$H = \Omega/\pi x$$

that is, the lines of flux are semi-circles.

Fig. 4. The determination of the permeance between parts of a magnetic circuit by summing the contribution of easily defined components of the permeance



It will be observed that $\int_x^\infty Hdx$ does not tend to a limit, so the fringe Δ does not tend to a limit, but goes on increasing as x increases. Fig. 17(a) can be used to determine the permeance between long poles, especially with a straight gap edge, it being assumed that the flux passing between semi-infinite poles up to a distance x equal to the pole length will represent fairly accurately the actual fringing.

In a similar manner the equivalent fringe for the case of poles in the form of semi-infinite plates can be derived: this is given in Fig. 17(b). Naturally this curve would apply only to magnets in which the field between the poles is low (say less than 500 gauss), the pole structure then taking the form of plates at the end of a much smaller section yoke.

For short cylinders of small radius, the curves of Fig. 17 are not very helpful and so measurements were made in a wedge type electrolytic bath to deduce the variation of the effective fringe with both the pole length and the pole diameter. The family of curves in Fig. 18 gives the results of this test.

It is useful in estimating permeance, to remember the analogy between capacitance and permeance. Using the relation

$$P = \frac{9{\cdot}4\pi}{10} C = 11{\cdot}320\ C, \quad \ldots\ldots\ldots\ldots (7)$$

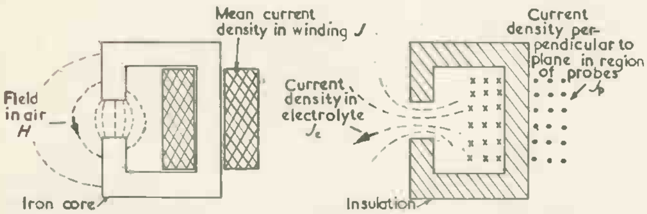
we can convert the capacitance of an arrangement, measured in pF, to the permeance of a similar arrangement, measured in cm. Hence formulæ for capacitance, listed in many places[1], can be utilized if appropriate. Obviously the inverse holds good, so that the curves of Figs. 17 and 18 can be used to deduce the capacitance between electrodes of the shapes given.

Analogue Solutions

No simple and convenient analogue appears to be available to solve the general electromagnetic problem of poles in the presence of conductors carrying current, although an exact analogy is possible using an electrolytic bath or a conducting sheet if the problem is restricted to one in two dimensions[2]. In Fig. 5(a) is shown the actual magnet system and in Fig. 5(b) is shown how this system is represented. A uniform depth of electrolyte is used: the winding is represented by probes carrying current proportional to the parent current and the iron core is represented by insulation. In the electrolytic bath, over a restricted region of the surface, the total current fed in is the surface integral $\int_s J_p ds$. This must equal the current in the electrolyte flowing parallel to the surface across the boundary of the region, i.e., $h\oint_0 J_e dl$. That is:

$$h\oint_0 J_e dl = \int_s J_p ds,$$

which it will be noticed is similar in form to equation (2). The iron is assumed to have infinite permeability(so that $H_{air(par)} = 0$; in the bath, however, $J_{perp} = 0$. The electrolytic bath representation is, therefore, orthogonal, with J_e representing H_{air} in magnitude but being perpendicular to H_{air} in direction. Lines of current flow in the bath represent magnetic equipotentials, and equipotentials in the bath represent magnetic lines of force. It is easy to verify with simple configurations of calculable permeance that resistance in the bath is proportional to permeance in the magnetic system. If the total current of all the probes representing one side of the winding is known, then a measurement of potential difference is all that is necessary to determine the resistance; consequently the potential difference between opposite sides of any section is proportional to the flux passing through that section.



a) Actual electromagnetic arrangment (b) Representation in electrolytic bath

Fig. 5. The orthogonal electrolytic bath for problems in two dimensions

One source of error in this bath could arise due to the conductance of the probes themselves: to represent the problem exactly the probes should not alter the "horizontal" conductivity, for the winding does not alter the permeability. Consequently, too many probes are undesirable and they should be reasonably small in diameter. In practice, this error can be made negligible.

Especially for permeance measurements, Teledeltos paper can be used instead of an electrolytic bath. The core can here be simply represented by cutting out the shape in the paper itself.

No other representation in the electrolytic bath gives so direct and useful an analogy to poles in the presence of conductors as the orthogonal bath just described. Poles distant from the magnetizing winding, however, can be simply and directly represented in an electrolytic bath by conducting electrodes, which outline the pole shapes near the gap only, the voltage across which being proportional to the M.M.F. (magnetic potential Ω) across the gap. Equipotentials in this bath are in a similar relative position to equipotentials in the actual system. In order to extend the analogy to the case of poles not distant from the winding, it is necessary to represent the iron circuit by a number of electrodes as in Fig. 6(a), each supplied at a voltage proportional to the M.M.F. at its position in the iron circuit. We must define the meaning of M.M.F. to suit our special purpose, for M.M.F. or magnetic potential difference has no meaning in the general electromagnetic problem. Here, the M.M.F. between any two electrodes connected by a line of flux is $4\pi/10$ times the total ampere turns embraced by the circuit formed by that flux line and the enclosing iron circuit less the M.M.F. dropped in that enclosing portion of the iron circuit. It is apparent that only near the gap will the M.M.F. be clearly defined: to determine the M.M.F. distant from the gap, an iterative procedure using the electrolytic bath is necessary, in which the electrode potentials are first guessed, the resulting flux lines deduced, a more accurate estimate of the potentials is then made using these results and applying equation (2), and so on. For the particular case when the coils are small in depth and wound close to the iron as in Fig. 6(b), the M.M.F. can be quickly and accurately deduced everywhere[3].

The electrolytic bath using conducting electrodes is a great convenience in determining leakage permeances, especially near the gap, and also for determining the field distribution in this region. Permeance is directly proportional to conductance in the bath, and the field is proportional to the potential difference between closely spaced twin probes.

The electrolytic bath is particularly convenient in deducing the attractive force between poles, or between poles and armatures, for if the applied M.M.F. is constant the force is proportional to $\partial P/\partial x$, which in turn is proportional to the rate of change with position of the conductance between electrodes. This type of bath has the great advantage over the previous type of Fig. 5 that it can be tilted and used for problems with cylindrical symmetry, the water's edge corresponding to the axis. A three-dimensional electrolytic bath can also be used, of course, with three-dimensional metal electrodes representing the poles. Teledeltos paper is also useful in the direct analogy, but only for problems in two dimensions. The electrodes can be painted on to the Teledeltos paper using a conducting paint, i.e. Aquadag.

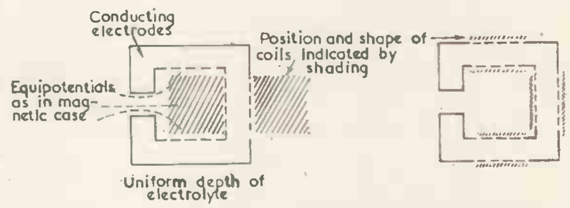


Fig. 6. Direct representation of a two-dimensional problem in an electrolytic bath in which electric potential in the bath is proportional to magnetic potential

It is obviously possible to utilize the analogy with electrostatics by making full scale or model electrodes similar to the poles and measuring the capacitance between them, when the permeance can be calculated by equation (7); this method is sometimes useful for three-dimensional problems where the poles are of awkward shape, as an alternative to the three-dimensional electrolytic tank. The electrodes can be simple structures covered with tinfoil or may even, perhaps, be plasticine dipped in conducting paint. A convenient method of measurement is by substitution. For example, if an accurate graduated variable capacitor is available, the capacitance of the electrode is measured by balancing a bridge with the variable capacitor in parallel set to a low value. The electrodes are then removed leaving all leads, which should be of small diameter wire, still in position and the variable capacitor is increased in value until the bridge is again balanced, the change in capacitance being noted. Convenient and manageable sizes of electrodes usually require capacitances of the order of 5 to 50pF to be measured. A symmetrical arrangement giving rise to an obvious equipotential plane can always be utilized to simplify the model making, the capacitance being measured between one electrode and a conducting sheet representing the equipotential plane. It is evident that the required capacitance is then a half of the measured capacitance.

MODELS

The usefulness of the methods of estimating the total flux at any position in a core which have been outlined in the previous two sections depend very largely on the facilities available to the designer. In some instances an awkward problem can be solved with a piece of Teledeltos paper, a dry battery and a multi-range D.C. meter; in other cases, it may be a ten minute job to measure the capacitance of a suitable electrode arrangement. In general, however, and especially for large and costly magnets, by far the best methods of designing, or checking a design, is using models. The procedure is then to design the magnet in the manner previously suggested, utilizing any information which may be available, then checking the design and resolving uncertainties using small scale models. If a scale model is to be made, flux calculations by the first method outlined by intelligent estimating is usually all that is necessary, and the whole design then becomes quick and sure.

Consider a scale model of a magnet, reduced in size in exact proportion, but in which the flux density in the iron and the field in the gap is the same. If the linear scale ratio of the model to the original is $L = l_m/l$, where the suffix "m" is used to refer to the model, it is simple to show that the ratio of the model to the original of the

permeance	P	$\propto L$,
flux	ϕ	$\propto L^2$,
M.M.F.	Ω	$\propto L$,
D.C. power	W	$\propto L$,
core section	A_i	$\propto L^2$
surface area	A_s	$\propto L^2$,
and watts/sq. in. of coil surface area		$\propto 1/L$

The size of the model is mainly influenced by:—

(1) Convenience of making;

(2) The sensitivity of the fluxmeter used in testing the model;

(3) The temperature rise of the exciting coils of the model;

(4) Whether field measurements in the gap are to be made.

From the point of view of convenience in making it is obviously desirable to keep the model of such size that it is easily and cheaply made.

The size dictated by the requirements of the fluxmeter is easily assessed. Consider a type of fluxmeter in common use having a full scale deflexion corresponding to 120×10^4 Maxwell turns (or line turns). If it assumed that the flux density in the iron is 10 kilogauss, and that winding 25 turns round the core is not regarded as being too laborious, and assuming that 25×10^4 Maxwell turns is sufficient deflexion (approximately 2 per cent accuracy), then 1sq.cm. of core area is required. Ballistic galvanometers can be obtained with a sensitivity 10 to 100 times that of this fluxmeter, so enabling the size of the model to be reduced, less search turns to be used, or the accuracy to be increased. It will be seen, however, that the core size demanded by the requirements of a normal fluxmeter does not conflict with (1) above.

Assuming that the temperature rise of the winding is proportional to the surface dissipation, as is the case over a reasonable range with natural and forced air cooling, then the temperature rise is inversely proportional to the scale factor L. This means that small scale models tend to get very hot, so that very small models are not to be recommended. It must be remembered that the excitation need not be continuous and the windings can be run very hot since it only has to survive the test; moreover, it is often convenient to cool the winding with blown air.

If measurements of field uniformity are to be made in the gap, this will require a more continuous rating for the coil, but in any case the gap will then need to be of reasonable dimensions to assist the actual field measurements. The field can be measured in various ways: the most common method is using a search coil and fluxmeter; or a magnetometer type field meter may be used, or a Hall effect gaussmeter which has an element of only a few square millimetres in size. Accurate positioning of the element itself requires the poles and gaps to be of reasonable dimensions.

One of the first tests to be made with a model is to measure the flux at any reasonable position in the iron circuit, preferably near the gap, and plot the flux to a base of magnetizing ampere turns. A well-known curve results, as shown in Fig. 7, which is similar to the open

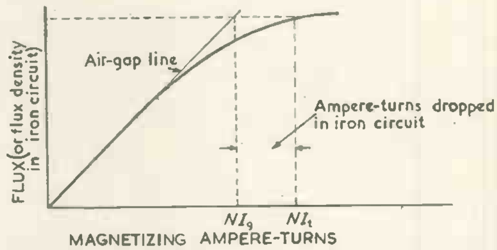


Fig. 7. Magnetization curve of a D.C. magnet from which can be deduced the field in the gap and the ampere-turns dropped in the iron circuit

circuit magnetization curve of a machine. The air-gap line drawn tangential to the curve at the origin gives the proportion of the applied ampere turns which are applied at the gap itself (this is accurate provided $(l/g)/\mu_{max} \ll 1$), hence the mean field across the gap is simply

$$H_m = 4\pi/10 \, . \, NI_g/g.$$

Since the mean field required is known, NI_g can be calculated and it can be checked to see if, at this value of NI_g, the total ampere turns NI_t is only 5 or 10 per cent greater, say, the value aimed for in the design. It is seen that unless field uniformity measurements are necessary, there is no need to measure directly the field in the gap.

(To be continued)

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GERMANIUM TRIODE PRODUCTION

A germanium triode of the point contact type which has for some time past been exhibited by the Research Laboratories of The General Electric Co. Ltd at the Physical Society and similar exhibitions, is now in pilot plant production in the company's works and is available to equipment makers in sufficient quantities for experimental work and prototype equipment. The triode uses single crystal germanium, and the unit is hermetically sealed in a metal can insulated from all electrodes. Flexible leads are provided for the connexions.

The triode is suitable for use in amplifiers, oscillators and for electronic switching applications. The current gain is greater than 2, the "knee" voltage is less than 3, and the collector current at -30V for zero emitter bias is less than 2mA. The maximum D.C. collector voltage is -50V, but normal operation is with much lower voltage, supplies of $22\frac{1}{2}$V being typical. Whatever the collector voltage used the dissipation should be kept below 100mW.

The Theory and Design of Cathode Follower Output Stages

By E. T. Emms,* A.R.C.S., B.Sc.

The cathode-follower output stage finds extensive use in cases where an output is required which may go both positive and negative with respect to earth. The author points out the limitations imposed on the design and gives a method of design which may be safely employed. The article also discusses the choice of valve to be employed.

THIS article deals with the design and behaviour of cathode-follower output stages feeding loads having an earthy end. This type of stage finds extensive use in buffer feedback amplifiers, integrating amplifiers, for driving servo-mechanisms, and in any application which requires outputs going both positive and negative with respect to earth. The circuit is shown in Fig. 1.

The chief limitations imposed on the design are three in number:—

(a) If reasonable linearity is to be achieved the valve anode current must not approach cut-off too closely;

(b) The maximum cathode current recommended by the valve manufacturers must not be exceeded;

(c) The maximum anode dissipation rating recommended by the valve manufacturers must not be exceeded.

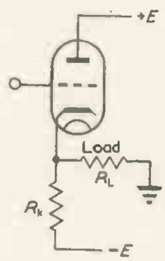


Fig. 1. Basic circuit

Limitation (a) may be explained in greater detail as follows. As the grid potential of the valve is reduced the anode current falls and the cathode potential falls correspondingly. Now let us assume that the grid potential is reduced sufficiently so that the valve is just cut-off. Since no anode current flows the current in the load is $E/(R_k + R_L)$. Further decrease of the grid potential now has no effect on this load current, and so the linearity (output voltage to input voltage) is radically changed from that obtaining when the valve is conducting. We may therefore assume that if good linearity is to be achieved, the valve cut-off must not be approached too closely.

Circuit Analysis

Let the cathode potential of the valve be V relative to earth. The current in the load is then V/R_L, while that in the cathode resistor $(V + E)/R_k$. The anode current is the sum of these two currents, giving:

$$I_a = \frac{E}{R_k} + V\left(\frac{1}{R_k} + \frac{1}{R_L}\right) \quad \ldots\ldots\ldots\ldots (1)$$

The anode-cathode potential of the valve is:

$$V_a = (E - V) \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

Now suppose that the voltage across the load must be varied between the limits $+V'$ and V'. When the cathode potential with respect to earth is $-V'$ the load current is V'/R_L, while the cathode resistor current is $(E - V')/R_k$ (Fig. 2).

If the valve is not to be cut off we must have:

$$(E - V')/R_k \geqslant V'/R_L$$

or

$$R_k \leqslant R_L[(E/V') - 1] \quad \ldots\ldots\ldots\ldots (3)$$

This equation gives the upper limit for the value of the cathode resistor R_k.

* *Air Trainers Ltd.*

Now let us consider the operation when the cathode is at a potential $+V'$ with respect to earth (Fig. 3). The load current is V'/R_L, while the cathode current is $(E + V')/R_k$, so that the anode current is:

$$(E + V')/R_k + (V'/R_L)$$

If the maximum permitted steady cathode current (recommended by the valve manufacturer) is I'_a we have:

$$I'_a \geqslant (E + V')/R_k + (V'/R_L)$$

or

$$R_k \geqslant (E + V')/[I_a - (V'/R_L)] \quad \ldots\ldots (4)$$

This equation gives the lower limit for the value of the cathode resistor R_k.

We must now take the remaining limiting factor into account—namely, the permissible anode dissipation of the valve. At any cathode potential V the anode dissipation is obtained using equations (1) and (2).

$$W_a = I_a V_a = (E - V)[(E/R_k) + V(1/R_k + 1/R_L)] \quad \ldots\ldots (5)$$

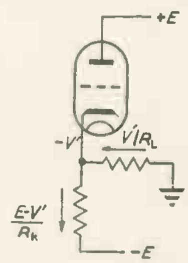


Fig. 2. Cathode $-V'$

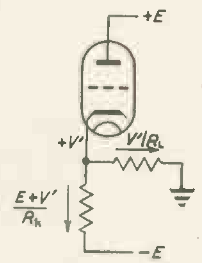


Fig. 3. Cathode $+V'$

We may now differentiate W_a with respect to V to find the value of the output voltage at which the valve dissipation is a maximum. We find that maximum dissipation occurs when:

$$V = (E/2)/[1 + (R_L/R_k)] \quad \ldots\ldots\ldots (6)$$

and inserting this in Equation (5) we find:

$$W_a = \frac{E^2}{4R_L} \cdot \frac{[1 + (2R_L/R_k)]^2}{[1 + (R_L/R_k)]} \quad \ldots\ldots (7)$$

In many cases the maximum output voltage required (V') is not as great as the value of V given by Equation (6). When this is so the maximum dissipation occurs with $V = V'$. Thus summarizing we have:

$$\left.\begin{aligned} W'_a &= \frac{E^2}{4R_L} \cdot \frac{[1 + (2R_L/R_k)]^2}{[1 + (R_L/R_k)]} & V' &> \frac{E}{2[1 + (R_L/R_k)]} \\ &= \frac{E - V'}{R_k}[E + V'(1 + R_k/R_L)] & V' &< \frac{E}{2[1 + (R_L/R_k)]} \end{aligned}\right\} \quad \ldots\ldots (8)$$

The Choice of Output Valve

The preceding analysis may be used to give a guide to the required characteristics of the output valve to drive a given load. Let us consider that the valve is just driven to cut-off when the output is $-V'$. Then from equation (3):

$$R_k = R_L[(E/V') - 1]$$

and inserting this in equations (4) and (8) we find that the maximum valve cathode current is:

$$I_a = (2EV')/[R_L(E - V')] \quad \ldots\ldots\ldots (9)$$

while the maximum anode dissipation is given by:

$$\left.\begin{aligned} W_a &= [E(E + V')^2]/[4R_L(E - V')] \qquad V' > E/3 \\ &= (2EV')/R_L \qquad\qquad\qquad\quad V' < E/3 \end{aligned}\right\} \ldots. (10)$$

We shall illustrate the use of equations (9) and (10) in determining the type of output valve required by an example. Suppose the load resistor is 400 ohms and that we require a voltage output across it ranging from −20 to +20 volts. The available H.T. supplies are +250 volts and −250 volts. Thus $V' = 20$, $R_L = 400$, $E = 250$. From Equation (9) we find that the maximum anode current required is:

$$(2EV')/[R_L(E-V')] = (2 \times 250 \times 20)/(400 \times 230) = 108 \text{ mA}$$

Since $V' < E/3$, the maximum anode dissipation is (from equation (10))

$$(2EV')/R_L = (2 \times 250 \times 20)/400 = 25 \text{ watts}$$

Thus a valve which has an available cathode current of 108mA or more, and an anode dissipation of 25 watts or more, is capable of serving as the output valve.

If in the interests of linearity it is desired to avoid approach to cut-off, the desired valve would have ratings somewhat in excess of those stated above. In fact if we choose a valve having ratings 20 per cent (or more) in excess of those stated above, we may expect good linearity.

A Complete Design

Let us suppose that we have to produce ± 100 volts across a 25kΩ load; the H.T. supplies are + 250 volts and −250 volts. Thus $V' = 100$, $E = 250$, $R_L = 25\,000$. Using Equation (9) we find that the valve required has a maximum cathode current somewhat in excess of:

$$(2EV')/[R_L(E - V')] = 13{\cdot}3\text{mA}$$

while since $V' > E/3$ the anode dissipation rating of the valve must be in excess of (equation (10)).

$$[E(E + V')^2]/[4R_L(E - V')] = 2{\cdot}04 \text{ watts}$$

Accordingly we might choose to use one section of the double triode 12AU7. This has a maximum permissible cathode current of 20mA per section and an anode dissipation rating of 2·75 watts per section.

From equation (3) we have:

$$R_k < R_L[(E/V') - 1] = 25 \times 10^3[(250/100) - 1]$$

so

$$R_k < 37{\cdot}5\text{k}\Omega$$

while from equation (4) we have:

$$R_k > (E+V')/[I_a - (V'/R_L)] = 350/(20 \times 10^{-3} - 4 \times 10^{-3})$$

or

$$R_k > 21{\cdot}9\text{k}\Omega$$

Thus we find that the value of the cathode resistor must be between 21·9kΩ and 37·5kΩ. We shall consider operation when the possible standard values are used, employing the equations derived in the analysis to build up Table 1.

TABLE 1

R_k (kΩ)	22	27	33
I_a MIN. (mA)	2·82	1·56	0·54
I_a MAX. (mA)	19·9	17·0	14·6
W'_a (watts)	3·14	2·63	2·24

The first case (with $R_k = 22\text{k}\Omega$) is not satisfactory as the anode dissipation rating of 2·75 watts is exceeded. The remaining cases are satisfactory from this point of view, but it is better to choose $R_k = 27\text{k}\Omega$ as it is to be expected that more linear operation will be obtained with the higher minimum current.

The circuit is shown in Fig. 4. We may actually proceed further, if necessary, and predict the linearity to be expected. Thus in this example Equations (1) and (2) become:

$$\left.\begin{aligned} V_a &= 250 - V \\ I_a(\text{mA}) &= 9{\cdot}26 + 0{\cdot}077V \end{aligned}\right\}$$

Thus we may find the values of V_a and I_a for any value of V and then using the $I_a - V_a$ characteristics of the valve find the value of V_{gk} required. This procedure is illustrated in Table 2.

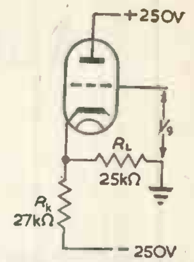


Fig. 4. Complete design

TABLE 2

V (volts)	V_a (volts)	I_a (mA)	V_{gk} (volts)	V_g (volts)
−100	350	1·56	−23·8	−123·8
− 80	330	3·10	−19·2	− 99·2
− 60	310	4·64	−16·4	− 76·4
− 40	290	6·18	−14·0	− 54·0
− 20	270	7·72	−11·6	− 31·6
0	250	9·26	− 8·8	− 8·8
+ 20	230	10·80	− 6·4	+ 13·6
+ 40	210	12·34	− 4·2	+ 35·8
+ 60	190	13·88	− 2·7	+ 57·3
+ 80	170	15·43	− 1·7	+ 78·3
+100	150	16·96	− 0·6	+ 99·4

These results are shown in Fig. 5, compared with the output-input voltage curve obtained experimentally. It will be seen that the experimental curve may be predicted to a reasonably high degree of accuracy using the methods explained in this article.

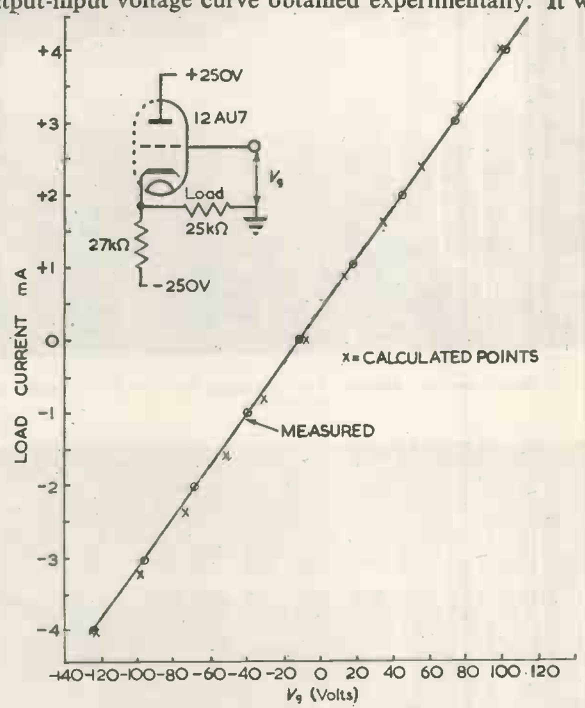


Fig. 5. Measured and calculated performance

Conclusions

The cathode follower output stage has been examined in detail. Procedures are given for choosing the most suitable valve and then proceeding to a final design. The designs are such that the maximum current specified for the valve is not exceeded; the maximum anode dissipation for the valve is not exceeded; the best linearity consistent with the two previous limitations is obtained.

Acknowledgment

Acknowledgment is due to Air Trainers Limited, Aylesbury, for permission to publish this article.

Automatic C.R.T. Trace Brightening

for Varying Amplitude R.F. Signals

By J. de Klerk*

IN photographing C.R.T. displays of short R.F. pulse envelopes of varying amplitude the correct film-time exposure is different for each pulse amplitude, due to the different degrees of brightness of the pulses. If the time exposure is chosen to give an accurately measurable photograph of the largest amplitude R.F. pulse, smaller amplitude pulses will be over exposed. This is illustrated in Fig. 1, where the exposure time was chosen to suit the brightness of the largest amplitude signal. Fig 2 shows that if the exposure time is correct for the smaller amplitude signals, the larger amplitude signals are underexposed.

This difficulty can be overcome by automatically increasing the brightness of the trace in proportion to the signal amplitude. Fig. 3 shows how an automatic brightening signal is obtained. The R.F. signal being displayed on the C.R.T. is rectified by V_1 and fed to an amplifier, whose output gives positive going pulses in proportion to the amplitudes of the R.F. signals. These rectified signals, whose amplitudes can be varied, are superimposed on the

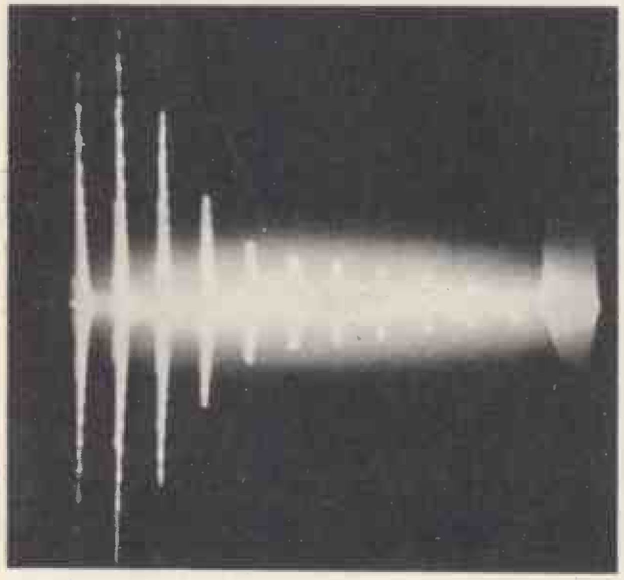


Fig. 1. Illustrating the different exposure time required for differing pulse amplitudes

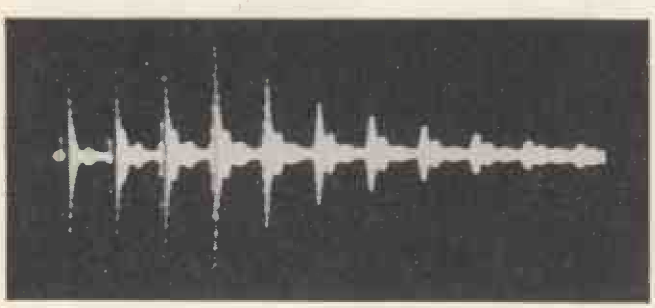


Fig. 2. Correct exposure for small amplitude pulses

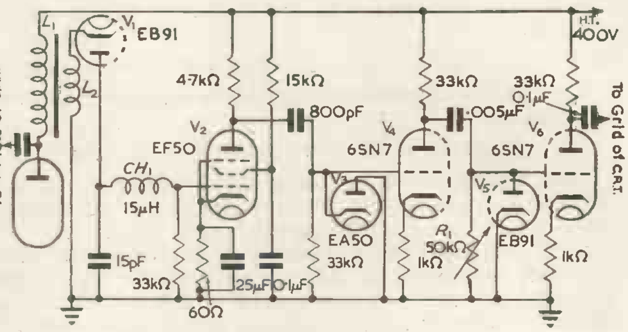


Fig. 3. The automatic brightening circuit

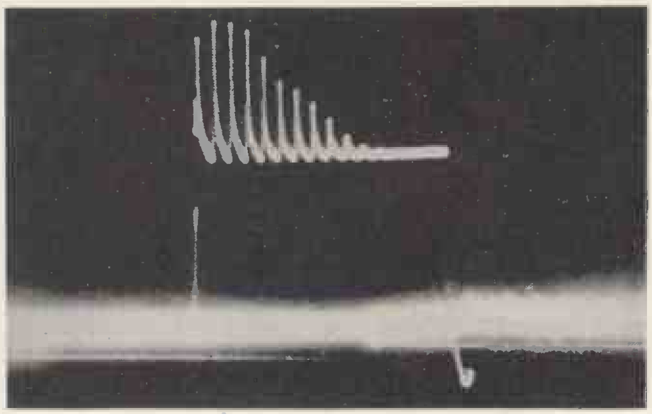
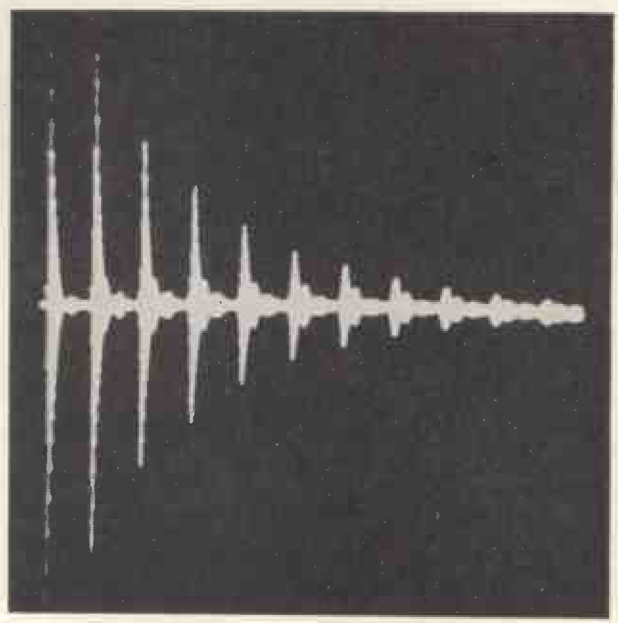


Fig. 4. The brightening pulse fed to C.R.T.

Fig. 5. Photograph obtained using the brightening circuit



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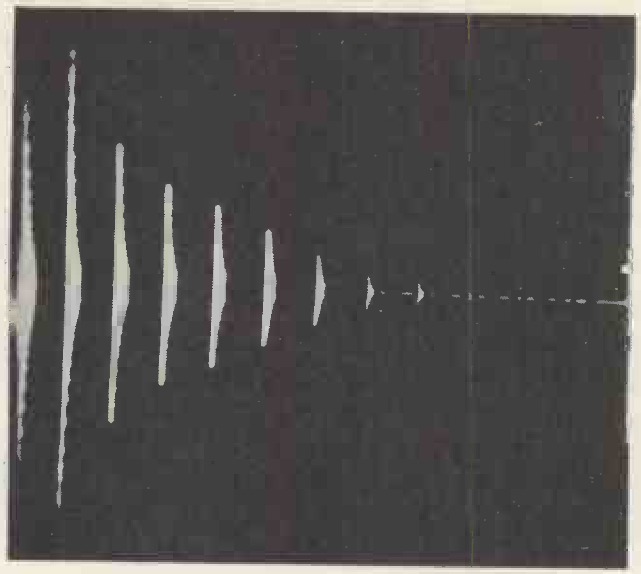

Fig. 6. Time-base suppression

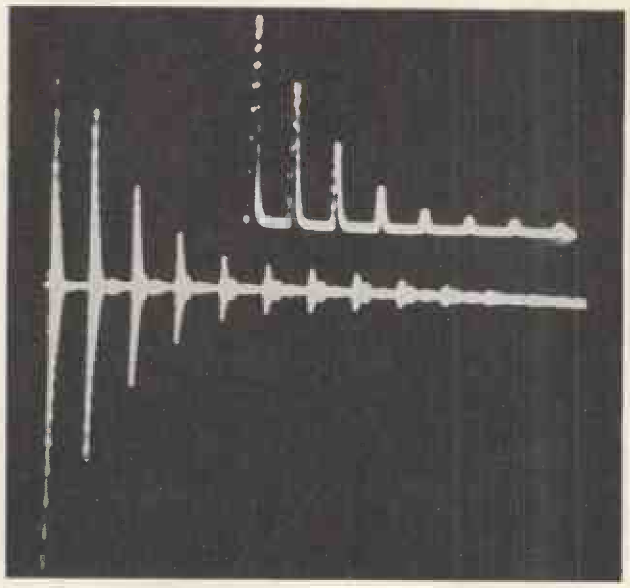

Fig. 7. Two methods of displaying ultrasonic pulses

positive brightening pulse normally obtained from the time-base circuit. Fig. 4 shows the new brightening pulse fed to the grid of the C.R.T.

By varying R_1 and the normal brightness control, it is possible to attain nearly uniform brightness of the R.F. pulse envelopes whatever their amplitudes may be, as shown in Fig. 5. It is also possible to display the pulses without showing the time-base, as shown in Fig. 6.

The coupling between L_1 and L_2 is chosen to reduce to a minimum the loading of L_1. The inductance of L_2 is chosen so that the maximum amplitude R.F. signal does not overload V_1.

Fig. 7 shows two methods of displaying received ultrasonic pulses, which have been simultaneously displayed on a double-beam C.R.T. Greater accuracy of amplitude measurement can be obtained from the R.F. pulse envelope method of display (lower trace) than from the rectified method as shown in Fig. 7.

A Wide-Range Oscillator for use with the Mass-Spectrometer Probe

By R. L. F. Boyd*, Ph.D., A.M.I.E.E., and D. Morris*, B.Sc.

The study of a wide variety of gas discharges by the R.F. *mass-spectrometer probe demands a source of radio frequency energy variable over a wide range.*

It is important, too, that the output voltage shall remain constant as the frequency is varied. The oscillator described in this article employs a self excited double beam-tetrode coupled to a capacitively loaded parallel line resonant circuit.

A range of 6-100Mc/s is provided without switching and the output voltage is maintained constant within ±5 per cent, at any predetermined level between 1 and 15 volts.

THE R.F. mass-spectrometer probe[1] was developed to provide a means of identifying and studying the concentrations of ions in gaseous discharges. Ions passing through a small orifice in a plane Langmuir probe receive an initial acceleration of several hundred volts and then pass through a system of some twelve coaxial disks or cylinders arranged along the line of flight of the ions and having an R.F. voltage fed in antiphase to alternate disks.

A definite proportion of those ions whose time of flight is six whole periods of the radio frequency receive sufficient energy from the R.F. to penetrate a grid at a retarding potential. The proportion depends on the relative magnitudes of the R.F. and retarding potentials and for this reason the output voltage of the oscillator must remain constant as the frequency is changed. The required R.F. voltage is about 5 volts peak-to-peak and the load (including the leads) is predominantly capacitive (about 12pF, although the inductance of the leads can be important at the higher frequencies).

The upper limit of the frequency range is set by the velocity of H^+ at 800 volts which is the normal accelerating voltage. For a twelve stage instrument having a path length of 2·5cm this gives a frequency of 98Mc/s. A frequency range of 17:1 gives a mass range of 1:290 thus taking in all atomic ions and the great majority of molecular ions likely to be encountered. A high frequency stability is not necessary as the resolution is low.

To reduce the leakage of the R.F. to other electrodes and to the electrometer used to measure the beam current, a balanced R.F. system is employed and adequate screening has been provided.

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General Description of the Oscillator

The problem of designing an oscillator covering the range 6-100Mc/s turns upon the choice of the form of the tuned circuit. It is clear that if the range is to be covered without excessive switching or coil changing both L and C must vary together.

The most successful arrangement (Fig. 1) has been found to be a resonant line whose electrical length may be varied in synchronism with a variable capacitor connected across its open end. At the high frequency end of the range the capacitance across the line is almost entirely the residual capacitance of the capacitor and the valve, and the frequency is largely determined by the line length. At the low frequency end of the range the capacitance of the capacitor is the principal factor controlling the frequency.

Fortunately no tuning of the grid circuits is required in this frequency range so that a very simple self-excited class-C arrangement suffices, the grids being fed through small adjustable capacitors from opposing anodes.

To make the instrument of manageable proportions the line of 16 S.W.G. silver wire is stretched around a pair of 24in diameter disks spaced 2in apart (Fig. 2). Its effective length is varied by a pair of silver contacts carried on an arm which rotates about the centre of the disks. The motion is also conveyed by a cam system to the rotor

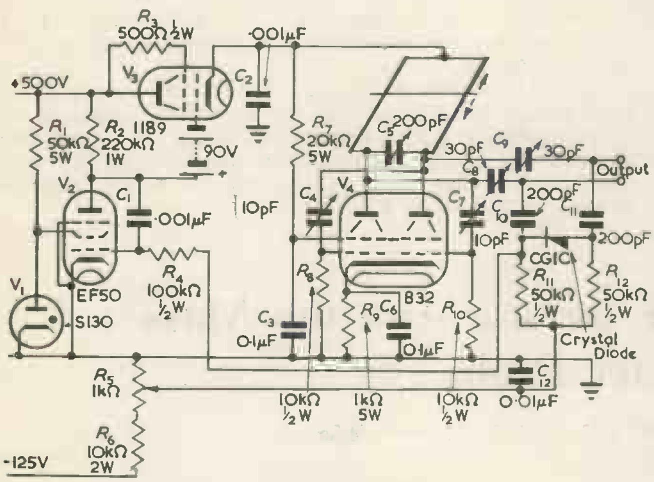

Fig. 1. Circuit of the oscillator

of the capacitor connected across the open ends of the line. In this way the capacitance is made to vary much more rapidly at the lower frequencies and so gives a better scale shape while permitting the retention of a compact capacitor construction.

The apparatus, apart from the power pack, is housed in an aluminium box 28in square and 6in deep giving a minimum clearance between the line and the box of 2in. In one corner with its anode pins very close to the line end is the 832 beam-tetrode. In the other corners are respectively a built-in wavemeter (necessary because the frequency depends on the capacitance of the load), an output meter and the self monitoring control circuit. The power pack is built on a separate panel mounting chassis and connected by screened leads.

The Control Circuit and the Ancillary Equipment

The amplitude of the R.F. output is controlled by automatic regulation of the H.T. supply voltage effected by supplying the H.T. from the cathode of an 1189 beam-tetrode. A germanium diode connected across the R.F. outputs, provides a D.C. voltage proportional to the R.F. voltage. This is amplified by a single EF50 pentode. The voltage so obtained is applied to the grid of the 1189 cathode-follower, D.C. coupling being provided by a miniature 90 volt battery. Amplitude control is effected by variation of the negative bias applied to the rectifier circuit. For the present purposes a single EF50 amplifying stage in the monitor is found to be sufficient.

The R.F. appearing across the output sockets of the oscillator is somewhat less than the generated voltage, by virtue of the series capacitors inserted in order to reduce the loading of the oscillator by the capacitance of the probe. It is this R.F. voltage which is monitored. Providing short leads to the probe are used, this is equivalent to controlling from the R.F. across the load itself. If long leads are used, the monitoring voltage may be taken from the R.F. voltage across the electrodes of the probe itself by mounting the rectifier circuit and its screened leads inside the probe head.

Performance

Because of the difficulty of matching a long cable to the impedance of the probe over a wide frequency band without tuning at the probe end, the oscillator is mounted on a rack very close to the probe so that the leads are

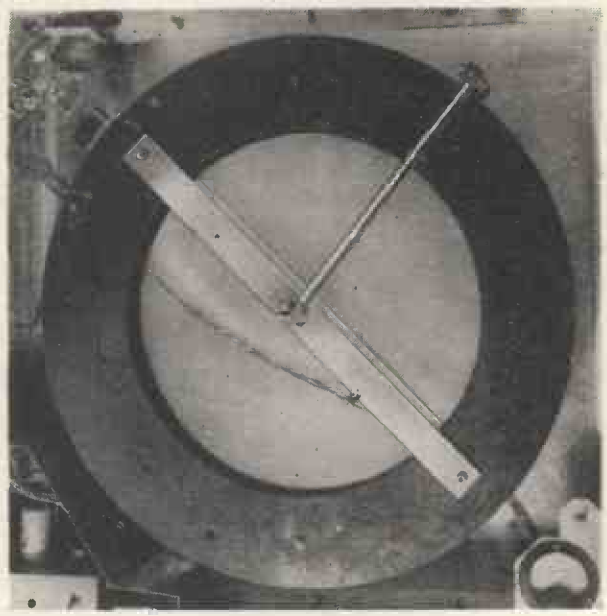

Fig. 2. The resonant line

only a few inches long. The anodes of the oscillator valve are coupled to the output by 30pF trimmers adjusted so that the oscillator attains a maximum frequency of 100Mc/s when feeding into a 12pF load.

The dependence of the output voltage on fluctuations of the mains voltage showed a change of about $1\frac{1}{2}$ per cent for a 10 per cent change in the latter. A drift in output voltage of the same order was found in the first hour after switching on, a few minutes having been allowed for initial warming up.

At a given setting of the output control the output voltage remains constant to within ± 5 per cent over the whole frequency range.

Acknowledgments

Acknowledgments are due to Professor H. S. W. Massey, F.R.S., the I.C.I. Fellowships Committee, and the Ministry of Education.

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A Simple Decimal Counter Using Binary Units

By W. C. G. Slatter*

A code for expressing the integers 0-9 in four binary digits is described. This code permits the use of a simple circuit arrangement for counting in decimal form. The circuits can be modified easily for normal binary counting.

THE advantage of using binary digits when using electrical circuits for arithmetic purposes has been thoroughly established. The expression of a result in binary form has disadvantages when its meaning is to be interpreted, or used for subsequent calculations, by human operators. A compromise is obtained by using a binary coded decimal system in which each digit in the decimal representation of a number is represented by four binary digits. The various combinations of the four binary digits are coded representations of the integers 0 to 9. The most obvious code is the normal binary one in which the binary digits represent in sequence ascending powers of 2 (Table 1). (In this and other illustrations the least significant digit is on the right, corresponding to the accepted convention for decimal representation of number.) To count in this system, binary counting circuits are taken in groups of four and a "feed-back" circuit introduced such that when the condition corresponding to a count of nine is reached, a further input pulse gives an output pulse and resets the group to zero.

Alternative Codes

The sixteen different combinations of four binary digits can be chosen to represent the integers 0 to 9 in approximately twenty-nine thousand million different ways (16!/6!). In the majority of cases the code is too complicated to have practical application. For most purposes it is essential that there should be a simple relationship between the representation of successive integer values, and it is always better and sometimes essential that each binary digit should always represent the same number in every combination in which it is used. It should be noted that if the binary digits have significance other than ascending powers of 2, values are not uniquely represented. There can be ambiguity in encoding a number but not in decoding.

Some of the possible codes have been adopted to meet special requirements. Where the results have to be read or used in decimal form it is usual to make use of resistance matrices. It has been shown[1] that permissible tolerances are greater with some codes than with others.

The simplest display is that in which the condition of each stage is indicated directly. In such circumstances the main requirements are simplicity of the circuits and ease of recognition of the pattern for each value.

Design of a Scale of Ten

The simplest combination of four binary circuits is in sequence producing a scale of sixteen in normal binary code. Any circuit arrangements of four binary stages to give a scale of ten is best considered as derived, by modification, from the normal binary combination. The modification can be either static in which in certain configurations the action of the next input pulse is modified by the intervention of gating, or inhibition of certain functions; or dynamic in which certain changes of configuration cause the generation of internal pulses to cause further change of configuration. A dynamic system may also require inhibition of certain functions, and this may be done temporarily as a direct consequence of the internal pulses. In general this will increase the operating time and decrease the maximum counting rate. Fergusson and Fraser[1], differentiate between "eight plus two" and "sixteen minus six" scales of ten; but this distinction is not fundamental, the difference being whether the change from the normal binary condition or operation takes place after registering nine, or earlier in the cycle of operation. All uses of four binary stages must in effect be considered as "sixteen minus six", and for ten input pulses, an additional six must be counted internally either by the generation of pulses, or by static changes of operating conditions that cause the same effect.

TABLE 1

Normal Binary Representation of Integers 0 to 9

0=0000	5=0101
1=0001	6=0110
2=0010	7=0111
3=0011	8=1000
4=0100	9=1001

TABLE 2

Code for Simple Counter Described

0=0000	5=0101
1=0001	6=0110
2=0010	7=0111
3=0011	8=1110
4=0100	9=1111

In circumstances where simplicity of circuit is the primary requirement, dynamic operation is preferred, requiring fewer components, and being less dependent on close tolerances. If the highest counting rate is to be maintained, there must be no inhibition of function and the internal pulses must be as short as those for normal operation. This requires that the internal pulses must be produced by a change, or changes, that occur only once in a cycle (ten external pulses) and that produce changes that themselves cause no further change, i.e., that put a binary stage "on". (The condition of a binary stage being "on" coincides with the corresponding digit having value one.) Now the only changes that occur only once in a cycle are those of the fourth, or most significant, binary stage, and the internal change which must correspond to an increase of count of six, are additional changes of the second and third stages, and which put these stages "on" when they are "off". Generation of an internal pulse when the last binary stage goes off would lead to one or two stages being on in the configuration corresponding to the representation of zero. It is considered best to avoid this, which limits the choice to a system in which internal pulses put the second and third stages on immediately after the fourth stage has come on, i.e., following the eighth input pulse of a cycle. This leads to the code shown in Table 2, which differs from the normal binary code only in the representation of eight and nine. It should be noted that by ascribing to the fourth digit (counting from right to left) the number two, the other three having their normal significance of 2^0, 2^1, and 2^2, the system is completely logical and the

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

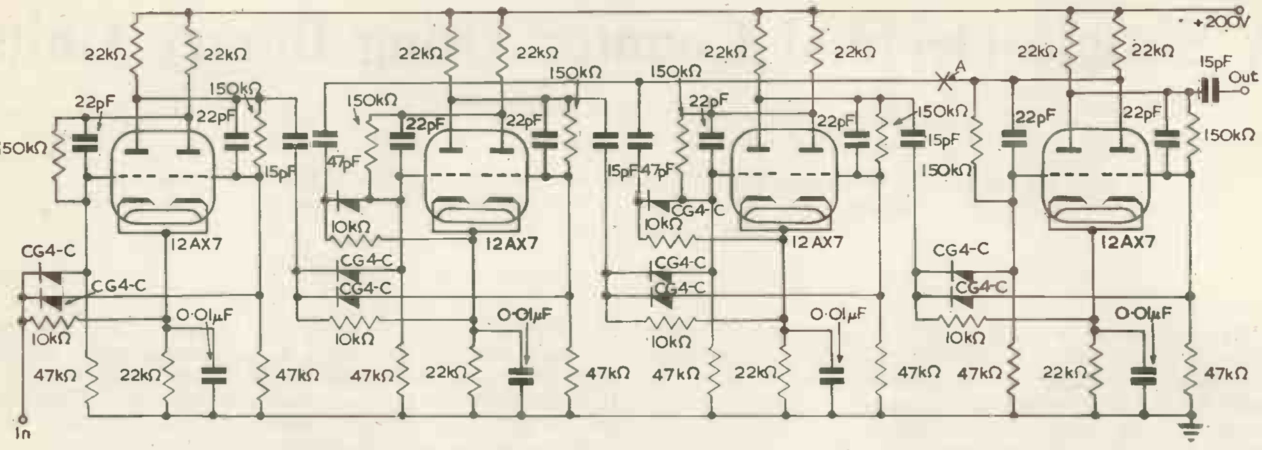

Fig. 1. A complete decimal counter *Breaking the feedback loop at A converts to normal binary counting*

code meets additional requirements to that for which it was derived.

A complete decimal counter, using this code, is shown in Fig. 1. This circuit operates up to 500kc/s, and has proved quite reliable in use. By removing the "feed-back" diodes the counter is directly converted to normal binary counting.

Acknowledgment

This article is based on work carried out at the Radar Research and Development Establishment, Ministry of Supply and is published by permission of the Ministry.

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A Method of inserting Blanking into a Television Video Waveform

Preserving the overall gamma

By J. E. Attew

THE insertion of blanking into a television signal must be performed so that the peaks of the blanking pulses coincide with black level, while the inherent curvature near cut-off of the "anode-current cut-off" type of switching circuit must be eliminated. This is necessary

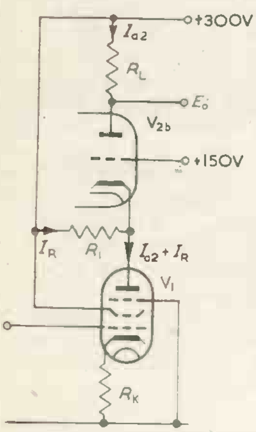


Fig. 1. Method of removing curvature of anode current near cut-off in V_2b

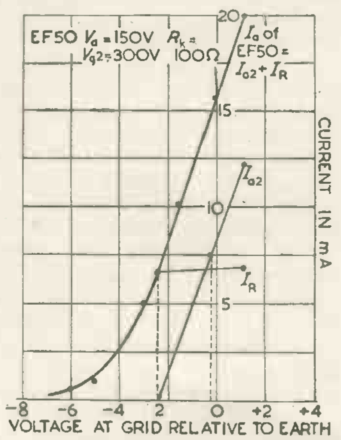


Fig. 2. Characteristics of circuit shown in Fig. 1

because such curvature increases the overall gamma of the system. Since the image orthicon type camera has unity gamma and the average viewing cathode-ray tube has a gamma in the neighbourhood of two, the resultant system gamma is higher than ordinarily necessary.

Fig. 1 shows a method of obtaining a linear relationship of anode current in V_{2b} for a change in grid voltage of V_1. A graph of currents in V_1, V_{2b} and R_1 are shown in Fig. 2, using an EF50 for V_1 with a cathode resistor of 100 ohms, one section of an ECC91 for V_2, and R_1 = 22kΩ. This shows the excellent linearity obtainable if zero

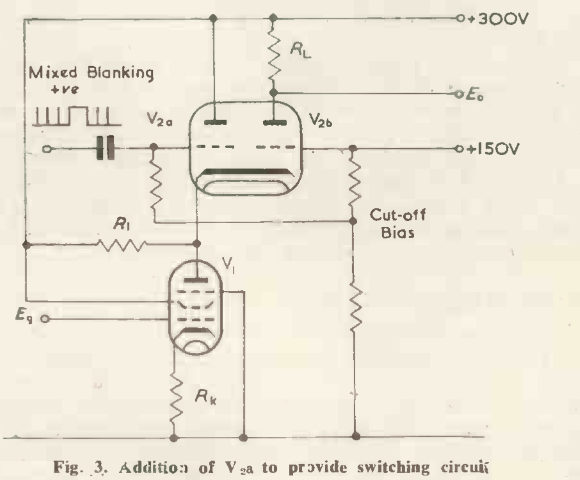


Fig. 3. Addition of V_2a to provide switching circuit

anode current in V_{2b} is established as black level.

By using the other half of V_2 in a switching circuit, where the anode current is transferred to this half during the blanking period, a very fast method of switching is achieved. This switching circuit also effectively clips off the top and bottom of the blanking signal, without grid current flow.

Fig. 3 shows the addition of this half of the valve to the circuit, where V_{2a} is normally biased to cut-off, and is caused to conduct during the positive going input signal, so cutting off the anode current of V_{2b}. The gain from input to output is approximately $(g_m R_L)/(g^k R_k + 1)$ where g_m is the mutual conductance of V_1, and $gk \approx g_m (1 + I_{g2}/I_a)$. ($I_{g2}$ and I_a are screen and anode currents of V_1 respectively).

The rigid establishment of black level at zero anode current in V_{2b} is achieved by means of a clamping circuit, using two diodes at the grid of V_1. Fig. 4 illustrates this circuit. The input signal must include a blanking reference, with white positive. The application of blanking to the camera tube or flying spot scanner cathode-ray tube, achieves this black level reference. The diodes are caused to conduct during the line blanking period by the clamping pulse, for a period sufficiently long to ensure that C_1 is charged to the bias level set by VR_1, which is adjusted to give black level as zero anode current in V_{2b}. The critical time-constant here is $(R_{S_1} + R_D/2 + R_{S_2}.\ C_1)$ where R_{S1} is the source impedance of the video input signal, R_d is the effective resistance of one diode, and R_{S2} is the source impedance of the circuit supplying the clamping pulse. The clamping pulse period must be at least four times the critical time-constant indicated.

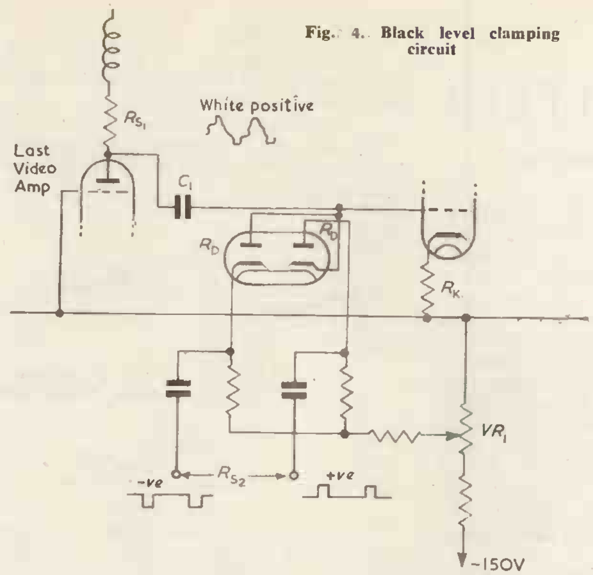


Fig. 4. Black level clamping circuit

This clamping circuit has the property of reducing hum and microphony introduced into the video signal by the preceding amplifiers to negligible proportions.

Fig. 5 shows a finalized circuit with the addition of a similar switching circuit to insert synchronizing pulses. This is used in a video amplifier with an overall bandwidth of 7·5Mc/s, giving excellent linearity and fast switching of the blanking and synchronizing signals. The composite waveform at the anode feeds an ECC91, arranged to drive an 80 ohms coaxial line.

Uses as a Linear Wideband Modulator

This circuit can be used to preserve the gamma of modulation of the output of a signal generator. Fig. 6 shows the proposed circuit.

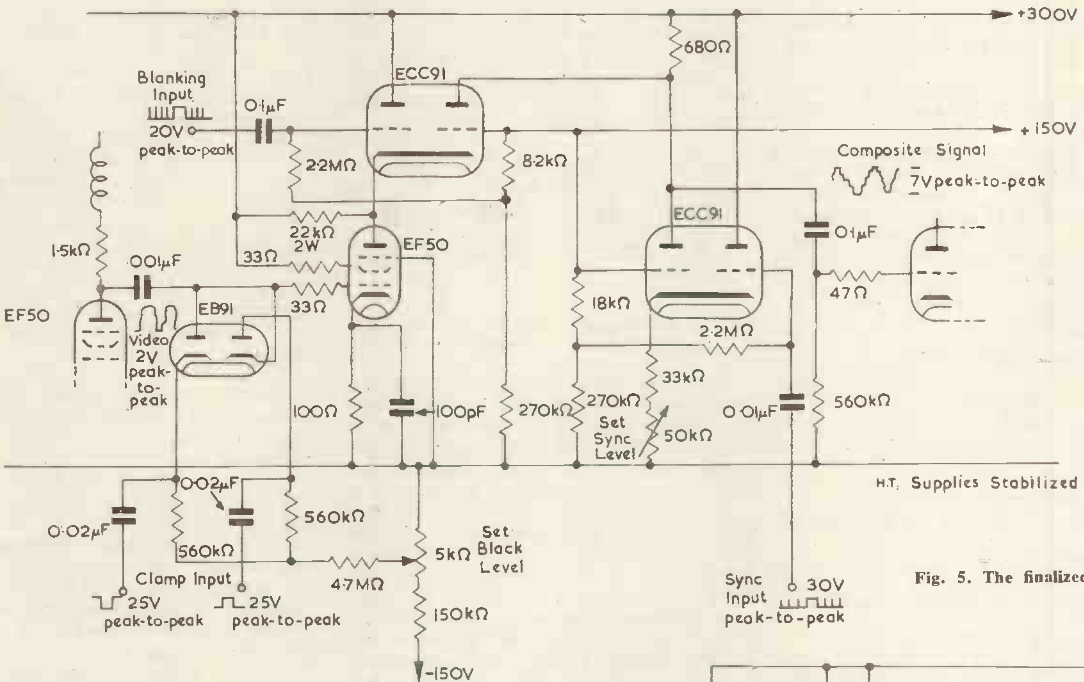


Fig. 5. The finalized circuit

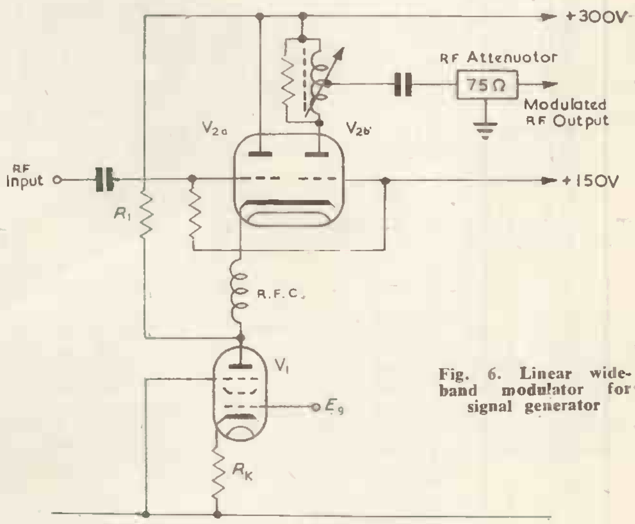


Fig. 6. Linear wideband modulator for signal generator

LETTERS TO THE EDITOR

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

The Equivalent Q of RC Networks

Dear Sir,—In Mr. D. A. H. Brown's interesting article contained in the July issue, there appears to be an error in the derivation of the basic formula from which the Q values are calculated.

Thus, starting from Mr. Brown's equation (1) and using the same notation

$$\phi = \tan^{-1} \frac{[\omega L(1-\omega^2 LC) - \omega CR^2]}{R}$$

$$= \tan^{-1}\left\{\frac{\omega_o L}{R}\,\frac{\omega}{\omega_o}\left[1-\left(\frac{\omega}{\omega_o}\right)^2\right] - \omega_o CR\,\frac{\omega_o}{\omega}\right\}$$

$$= \tan^{-1}[\alpha Q(1-\alpha^2) - \alpha/Q] \quad \ldots\ldots\ldots \quad \text{(A)}$$

which differs from Mr. Brown's equation (1).

However, differentiating and putting $\alpha = 1$, $\phi =$ tan ϕ, we get for large values of Q.

$d\phi/d\alpha_{\ (\alpha=1)} = -2Q$ which is in agreement with Mr. Brown despite his original error.

However, Mr. Brown then goes on to use this formula which is based on the fact that Q is much *greater* than unity, to obtain values of Q which are approximately *equal* to unity. This seems to me somewhat improper and a more accurate method is appended below.

Starting from my equation (A) and differentiating we get

$$\frac{d\phi}{d\alpha} = \frac{O - 1/Q - 3\alpha^2 Q}{1 + [\alpha Q(1-\alpha^2) - \alpha/Q]^2}$$

which for $\alpha = 1$ becomes

$$\frac{d\phi}{d\alpha}(\alpha=1) = \frac{-2Q - 1/O}{1 + 1/Q^2} = -\frac{Q(2O^2+1)}{Q^2+1} \quad \ldots\ldots\ldots \quad \text{(B)}$$

Applying this expression to the first of the *RC* circuits considered by Mr. Brown

where $\frac{d\phi}{d\alpha}(\alpha=1) = \pm \frac{\sqrt{3}}{4}$

for three sections we get the equation

$$2Q^3 - \frac{3\sqrt{3}}{4}Q^2 + Q - \frac{3\sqrt{3}}{4} = 0$$

This equation has only one real root, this being Q = 0·885 a value which is some 30 per cent higher than Mr. Brown's value. Since the solution of cubic equation is rather tedious a more approximate method is as follows:

Rewriting my equation (B)

$$\frac{d\phi}{d\alpha}(\alpha-1) = -Q(2 - 1/Q^2 + 1)$$

and for values of Q approximately equal to unity then

$$\frac{d\phi}{d\alpha}(\alpha-1) = \frac{3Q}{2} \quad \ldots\ldots\ldots \quad \text{(C)}$$

Applying this to the circuit considered above this yields Q = $\sqrt{3}/2$ = 0·866 which is only 2 per cent different from the value calculated above.

Thus a more accurate equivalent Q value is obtained by multiplying Mr. Brown's figures by a factor 1·5.

Yours faithfully,

P. Tenger,
Emsworth, Hampshire.

Dear Sir,—In his article, interesting especially because of the attractive and useful definition of the effective Q, in the July issue, Mr. D. A. H. Brown made a statement which is erroneous. On page 296 he writes: "It is interesting to note that though the phase of the output voltage from a twin-*T* or bridged-*T* network jumps continuously from 90° lagging just below the null frequency to 90° leading just above the null frequency, the rate of change of phase with frequency is continuous at the null frequency." Based on this conclusion is the remark at the end of the article that it is impossible to make a very stable oscillator, for instance with a twin-*T* network.

It is easily shown, however, that the rate of change of phase, far from being continuous, jumps from a finite value just below and above the null frequency to an infinitely large value at the null frequency itself. This is the case with any null-transmission network.

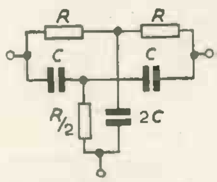


Fig. A. Null transmission network

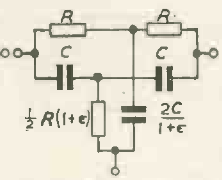


Fig. B. Minimum transmission network

Moreover, as a rule, a null transmission network cannot serve as a frequency determining network. There must be an output voltage for the operating frequency and for a good frequency stability, the phase characteristic should be as steep as possible at the same frequency. Therefore the null transmission network must be altered in such a way that it becomes a minimum transmission network. As a result the slope of phase characteristic is less steep, but when the alterations are small the steepness will be large enough to get a stable oscillator.

To prove this the null transmission network of Fig. A must be considered as the limit case of the minimum transmission network of Fig. B.

For the minimum transmission network the following formula holds, if $\epsilon << 1$,

$$\frac{E_o}{E_i} = \frac{1}{1 + \dfrac{4}{\epsilon + j(\alpha - 1/\alpha)}} \quad \ldots\ldots\ldots \quad (1)$$

in which $\alpha = \omega/\omega_o$ and $\omega_o = 1/CR$.

From formula (1) it can be derived that for the phase ϕ between the output and the input voltage the following equation holds for small values of ϵ

$$\phi = \tan^{-1}\frac{4(\alpha - 1/\alpha)}{(\alpha - 1/\alpha)^2 + 4\epsilon} \quad \ldots\ldots\ldots \quad (2)$$

From (2) it follows:

$$\frac{d\phi}{d\alpha} = \frac{16\epsilon(1+\frac{1}{2}) - 4(1+\frac{1}{2})\ (\alpha - 1/\alpha)^2}{16(\alpha - 1/\alpha)^2 + \left\{(\alpha - 1/\alpha)^2 + 4\epsilon\right\}^2} \quad \ldots\ldots \quad (3)$$

For the operating frequency $\alpha = 1$ this expression becomes:

$$d\phi/d\alpha = 2/\epsilon \quad \ldots\ldots\ldots\ldots \quad (4)$$

According to the definition of the effective Q this means:

$$Q_{eff} = 1/\epsilon \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

For the null transmission network ($\epsilon \to O$) we thus find an infinitely large value of Q_{eff}.

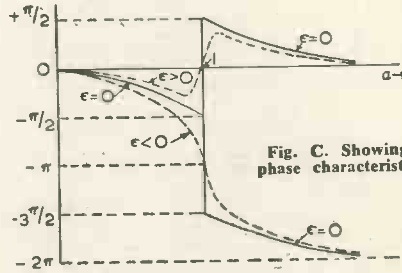


Fig. C. Showing phase characteristics

This can also be seen from the phase characteristic of the two discussed cases (Fig. C). It should be remarked that the phase characteristic of the case $\epsilon = 0$ can be drawn in two different ways: at $\alpha = 1$ jumping from $-\pi/2$ to $+\pi/2$ or from $-\pi/2$ to $-3\pi/2$. The phase characteristic for $\epsilon \neq 0$ can be drawn in one way only.

From the foregoing it follows that as to the effective Q, the twin-*T* network is equivalent to what Mr. Brown calls the capacitance bridged-*T* network of Fig. 11 of his article. As to the attenuation at the operating frequency the twin-*T* network is advantageous with respect to the capacitance bridged-*T* network. At the operating frequency the transfer voltage ratio of the twin-*T* network is given according to formula (1).

$$E_o/E_i = \epsilon/4 \text{ if } \epsilon << 1 \quad \ldots \quad (7)$$

For the transfer voltage ratio, of the capacitance bridged network, however, the following relation holds for large values of a

$$E_o/E_i = 2/a2 \quad \ldots\ldots\ldots \quad (8)$$

From the formulae (7) and (8) the following conclusion may be drawn:

(a) With the twin-*T* network the output voltage may be in phase ($\epsilon > O$) or in counterphase ($\epsilon < O$) with the input voltage. With the bridged-*T* network the output voltage is always in phase with the input voltage.

(b) To achieve an effective Q of 10 with the twin-*T* network an attenuation of 1/40 must be accepted (see formulae (6) and (7). For the same result with the bridged-*T* network the attenuation is 1/200 ($a = 20$). To get an oscillator with a certain frequency stability more amplification is needed in the latter case than in the former case.

From the foregoing argumentation it will be clear that the twin-*T* network as a frequency determining network is superior with respect to the bridged-*T* network.

A very small stable *RC* oscillator can be made with the twin-*T* network. For very small negative values of ϵ this can be done with one valve, as the output voltage is in counterphase with the input voltage at the operating frequency for this case. The more amplification available, the smaller ϵ can be chosen and as a result the oscillator will be more stable. Further theoretical and practical details about this problem are given in two articles in the Dutch Journal *Het P.T.T. bedrijf*, October, 1951, and February, 1952, respectively.

Yours faithfully,
A. P. BOLLE,
Central Laboratory of the
Netherlands P.T.T.
The Hague, Holland.

The author replies:

DEAR SIR,—I am glad to acknowledge the corrections made by Mr. Tenger. The approximation based on $Q \gg 1$ and its subsequent application to values of Q near unity had seemed to me unsatisfactory. Mr. Tenger's method avoids this.

However, the low Q case of the LC circuit is complex; it is no longer permissible to take the resonant frequency as $\omega_o^2 = 1/LC$, for the impedance becomes resistive when $\omega_o^2 = 1/LC - R^2/L^2$. Mr. Tenger's method is itself an approximation (based on $\omega_o^2 = 1/LC$). My approximations therefore have the virtue of simplicity until experimental work determines the absolute accuracy of the equivalent Q.

The method of Mr. Bolle which considers zero transmission networks as limiting cases of minimum transmission networks is most interesting. My equation (40) shows that for a Q of 10, A must be 40; for unity loop gain this implies a network attenuation of 1/40. This agrees with Mr. Bolle's conclusion (b) but his method of arriving at this result is much more illuminating.

Yours faithfully,
D. A. H. BROWN,
T.R.E., Ministry of Supply.

Investigation of Ionosphoic Absorption

DEAR SIR,—I read in your April and May issues the description, by Messrs. Jenkins and Ratcliff, of their interesting Ionosphere Absorption Measurement Apparatus.

Having worked in the same field for several years, I designed an equipment fairly different, in some ways, to that of Messrs. Jenkins and Ratcliff. but having quite analogous purposes. It is now in use at the Domont Ionosphère Station of the Bureau Ionosphèrique Français.

There is, between these apparatus, a difference worth noting. I don't use a "noise eliminator" and I think Messrs. Jenkins and Ratcliff are wrong with their one. I cannot agree that, if both signal and noise are present, the mean detected value of the composite signal is the sum of the mean detected values of the individual components.

Consider, for instance, the case of a single, c.w. interfering signal. The desired signal would be of the form*:

(1) $x = a \sin \omega t$

and the interfering signal:

(2) $y = b \sin (\omega + E)t$

we can write:

(3) $(x + y) = (a + b \cos Et) \sin \omega t + b \sin Et \cos \omega t$

As ω is very much larger than E, the detector stage will deliver a signal proportional to

(4) $\text{amplitude } (x+y) = [(a+b \cos Et)^2 + (b \sin Et)^2]^{1/2} = [a^2+b^2+2\, ab \cos Et]^{1/2}$

* (*Just before reaching the detector stage*)

and the mean detected value will be

(5) $\frac{1}{\pi}\int_0^{\pi} [a^2+b^2+2\, ab \cos Et]^{1/2}\, d\,(Et)$.

This integral is known as "E" integral, or "complete elliptical integral of the second kind." It is tabulated, and it is obviously an *even* function of b/a, the noise-to-signal ratio. For small values of b/a, it is equivalent to: $a + \frac{b^2}{4a}$

This result can be extended without difficulty to the case of any number of c.w. simultaneously (see *B.S.T.J.*, *23*, 282 (1944); *24*, 46 (1945); *27*, 109 (1946)). In this case, the mean detected value is still an *even* function of the noise-to-signal ratio, fairly similar to the above "E" integral.

The proposed method of noise elimination would be correct if a square-law or a peak detector were used, instead of the standard, linear one. With the square-law device, the ab term cancels out in the process of integration, and with the peak detector there is no integral left, but only the maximum detected value of $a + b$.

Several years ago, the peak detector was successfully tried in my laboratory, but it is not actually used with our standard equipment, because of the intricate circuit involved. The square-law device was not tried although it is likely to give the same good results. but with about the same degree of complexity.

It is worth noting that, in fact, the tables for the "E" integral show that, for $a = b$, the error made by ignoring the interfering signal is only 2db. So I am not sure that it is worth while designing such an elaborate eliminator. It is not impossible that the answer to this question would be chiefly a matter of severity of interference at the location of the station.

Yours faithfully,
S. J. ESTRABAUD,
Laboratoire National
de Radioélectricité,
Bagneux.

The author replies:

DEAR SIR,—We were interested to hear from M. Estrabaud of the apparatus which he has designed for measuring ionospheric absorption.

It is, of course, true that a noise eliminator of the type which we have described would give an incorrect result in a completely linear system. This would mean that the noise sample would be arithmetically subtracted from the vector sum of signal and noise and would tend to give a reading which was too low.

The use of a square law detector will give a correct result and in our apparatus the linear detector is followed by an amplifier working near the bottom bend of its characteristic which yields a result essentially the same as that obtained from a square law detector.

The theoretical analysis of the problem of the interference between a wanted signal and an unwanted signal which may be c.w., pulse or random noise is a matter of some complexity.

As described in our article in which results obtained using the manual technique were compared with simultaneous measurements using this automatic equipment, it appeared that under quiet conditions no serious discrepancy occurred.

Under conditions of heavy interference it is clear that the manual technique yields values which are in excess of the true signal intensity and some method of balancing out the interfering signal is highly desirable. The extent to which the automatic equipment in its present form compensates or over compensates for the effect of noise is a matter for further experiment.

The decision as to whether or not a noise eliminator is required will depend on the signal-to-noise ratio at the station which will be determined by the power available at the transmitter as well as the severity of interference. The equipment we have described was designed so as to be capable of use under the most difficult conditions—conditions which would normally preclude manual measurement.

Incidentally, a further disadvantage of equipment which does not provide noise balance is that it would be necessary to maintain a constant vigilance that the width of the selector gate corresponded closely with the width of the echo pulse.

Yours faithfully,
G. RATCLIFF,
J. B. JENKINS,
Department of Engineering,
University College of Swansea.

Noise Free Instrument Cable

DEAR SIR,—The short article entitled "Noise Free Instrument Cable" by M. Lorant (July, 1953), on work by the American National Bureau of Standards, seems to provide yet further evidence of the duplication of effort and work which the limitation of technical information originates.

The theory postulated and the practical conclusions made, which the article appears to infer to be new, have been accepted generally in the electrical cable industry in this country for a number of years. "Low Noise" and "Antimicrophonic" cables have been advertised and manufactured by leading British cable makers, using the technique described, for numerous instrumentation uses, amongst which investigation of the propagation of shock waves associated with explosions was one of the earliest applications during the war years.

May I suggest that mechanical considerations and the requirement for minimum capacitance implied by the use of piezo-electric sensing elements and cathode-follower stages, impose limitations on the smallness it is possible to achieve with this type of cable, and also on the materials which may be used in their construction. Hence the final sentence of the article should only be accepted with reserve.

Yours faithfully,
D. POLLARD,
Leigh, Lancashire.

The author replies:

DEAR SIR, — I have no comments to make. The research on noiseless or low noise cable has not ended yet; it will go on for a long time to come and will, very likely, meet again with "duplication of effort" in various lands.

Yours faithfully,
MICHAEL LORANT,
London, W.14.

BOOK REVIEWS

Fernsehtechnik (Television Engineering)

By Dr. F. Kirschstein and Dr. C. Krawinkel. 288 pp., 236 figs. Medium 8vo. S. Hirzel Verlag, Stuttgart. 1952. Price DM.25.

ALTHOUGH there exists quite a large number of treatises on the ever increasing field of television engineering, this book, written by two experts in the theory and practice of this field, will be welcomed by all those interested in television and able to study a German text. It is primarily intended as an introduction for the student of physics and telecommunication engineering already well acquainted with the principles of optics and electronics. The treatment is to some extent mathematical but without making too high demands on the reader in that respect. A large number of well chosen illustrations, diagrams and graphs serves for making also more complicated methods and systems well comprehensible. One hundred and fifty-four references to literature enable the student to find further details on the subjects treated here.

As also in Germany the development of television was interrupted during the war years much use is made of foreign literature particularly as regards the progress made in the U.S.A. But reference is made in some instances also to British developments, e.g. mention is made of the CPS Emitron, of the wide band coaxial cable connexion between London and Birmingham and of the Manchester-Edinburgh television radio relay system. On the other hand, no mention is made, e.g. of the Skiatron described in this journal in January, 1948, and which has found wide application in radar engineering.

In its eleven chapters the book deals with the following subjects: (1) Fundamentals of television engineering, particularly the choice of the number of lines and its influence on the image quality. (2) The transformation of light into electric current, giving the characteristics of the various photo-electric cells and a clear account on secondary electron multipliers. (3) The pick-up systems without and with image storage. A brief digression is made on electron optics. The iconoscope, super-iconoscope, orthicon and super-orthicon are discussed in detail and briefly also recent pick-up systems like the isicon and the vidicon. (4) This chapter deals with image reproduction and discusses the Braun tube for direct observation and for projection. The principle of Fischer-Thiemann's eidophor-projector is briefly discussed, but it is not stated whether since its presentation at the 1948 television congress at Zurich it has found any practical application. (5) The next chapter is devoted to a detailed discussion of the methods used for deflecting the electron beam, particularly those for producing the saw-tooth characteristic by means of relaxation or blocking oscillator. (6) The requirements which the synchronizing impulses must fulfil, their generation and transmission are discussed at some length. (7) Video frequency amplifiers and tuned amplifiers are dealt with and special attention is given to the difficulties encountered with high frequencies in the first case and to the use of staggered pairs and double-tuned circuits in the second case. (8) The next chapter deals with the modulation of carrier waves with image currents. The use of the Nyquist filter for vestigial sideband transmission is explained. Chapters X and XI discuss the problems of television broadcasting, the transmitters, aerials and receivers and the general arrangement of a television broadcasting station, the means for transmission over great distances, relay stations, cables, the costs of transmission and finally a brief account on colour television.

The book is written in a lucid style and very well produced.

R. NEUMANN.

Television Picture Faults

By John Cura and Leonard Stanley. 68 pp., 138 figs. Crown 8vo. Television Times Ltd. 1953. Price 3s. 6d.

ON the cover of this interesting booklet, probably the only one of its kind published in Great Britain, it is stated to contain 150 actual screen photographs. This reviewer counted 149 screen photographs and one photograph of an actual "c" card, but they are so interesting and well done that it is unlikely that there will be any complaints about the error.

The photographs show practically every kind of distortion which can occur in a received picture due to incorrect setting of controls, breakdown of components and design faults. Each distortion is described, generally non-technically and then technically, and possible causes and cures given. In addition there are several pictures showing results which can be expected from a correctly designed and adjusted receiver and the effects of various forms of interference upon a picture.

This reviewer would have preferred to have seen a few of the photographs rather larger, or significant parts of them only shown in the available space, such as the bandwidth bars in photograph 94, which is supposed to show all the bars clearly except the 3Mc/s group. In fact the 2Mc/s bars seem poor and the 2½Mc/s bars are practically undetectable. Picture 19, which is intended to show uneven focus is also much too small to show the fault, and is in addition a quite unsuitable picture. The left-hand side which is stated to be out of focus shows a man in the background and out of focus due to insufficient depth of focus of the camera.

Two photographs showing "pulling on whites" and "triggering on picture" are of a small section of the "c" card only and are extremely informative.

There are a few parts of the text which are not at all clearly written, such as the statement on page 52 that "the black edge of the left-hand side of the picture is the 5μsec delay period which allows time for the flyback to end before the commencement of the scan." It would be difficult to imagine the scan starting before the end of the flyback. Under picture 16 showing mush it is stated that "the only remedy for improvement of picture quality in fringe areas . . . .". Why remedy an improvement of picture quality?

The booklet is, however, very good value indeed, a prodigious amount of patience, labour and skill must have gone into its preparation and it should be of great interest to technical and non-technical viewers.

C. H. BANTHORPE.

The Design of Electronic Measuring Instruments

By F. G. Spreadbury. 102 pp., 63 figs. Demy 8vo. The Association of Engineering and Shipbuilding Draughtsmen. Session 1952-3. Price 4s.

WITHIN the limits of 100 pages the theory and design of many electronic measuring instruments is well given, although it is possible that much of the mathematical work will be beyond the reader for whom the book is published; it is worthy of a wider circulation than appears likely to be the case. To some extent the pamphlet (so-called by the publishers) is a companion to the author's book "Electronics." Valve voltmeters are thoroughly treated, but the C.R.O. is dismissed in one paragraph. Headings are: Valve Characteristics; Valve Voltmeters; Current Measurement; Wattmeters; Frequency Measurement; Stroboscopes; Tachometers and Flashlamps; Time Measurement and Control; Insulation Testing; the C.R.O.; Applications of Electronics to Bridges; Power Supplies. Many of the instruments described are of the author's design. The text and mathematical work are well printed, but many of the line diagrams are reduced so much as to be nearly illegible without a lens. Some mental agility is required in reading, since p is used instead of ω or 2πf, and ω for Ω or ohms; both symbols as used here having been obsolete for about thirty years. This is one of the few British publications on the subject.

E. H. W. BANNER.

Télévision Dépannage

By A. V. J. Martin. 176 pp., 180 figs. Demy 8vo. Editions Radio, Paris. 1953. 600 fr.

THIS is the first book in the French language devoted to the installing, focusing and repairing of televisors. The author has himself had many years of experience in servicing televisors and he uses his practical knowledge together with that of his colleagues in compiling this work.

The book is divided into three sections and systematically deals with installation and the actual workings and gives a description of each instrument in a televisor. It also lists nearly every possible cause of breakdown together with their remedies.

This work should prove to be of great use to any French television enthusiast.

Basic Electronic Test Instruments

By Rufus P. Turner. 256 pp., 172 figs. Medium 8vo. Rinehart Books Inc., New York. 1953. Price $4.

THIS book is directed primarily to the man the Americans call a technician, to enable him to understand the measurements he has to make in various applications of his art. The use of commercial products or kit-built equipment is given first place, except for some special purposes, where home constructed units offer the best proposition.

The book follows a logical sequence, building up from simple meters to multi-range instruments, including ohmmeters, then vacuum tube voltmeters, and meters for the measurement of power in different circuits, and impedance. Various types of bridge circuit are followed by oscilloscope applications, and various types of signal generator. Devices for frequency measurement and analysis, special equipment for overall testing, distortion measurement, and tube checking, finish out quite a comprehensive coverage.

Most of the sixteen chapters have review questions at the end of them, so the reader can test the knowledge he has gained from the chapter, but this reviewer feels that, although the numerical calculations involved are simple, the value of these questions to technicians would be enhanced if answers were provided at the back. Another useful feature is the list of references for further information, given at the end of several chapters.

In the earlier part, the author carefully breaks down multi-purpose circuits, so his reader can take in each principle by easy stages. Towards the end of the book, it would seem that the author felt the pressure of either space or time, with the result that the amount of attention given to such detail seems to vanish. The sections dealing with the use of Lissajou's figures and with displays using square waves, give rather poor illustrations, some of which are very unlike any possible display, and therefore would not help the reader in identifying his findings.

For showing calibrations, the author mostly gives tabulations of figures. This reviewer feels that a graphical presentation is better for this purpose, with actual readings distinguished from the theoretical curve in the usual way, where appropriate—it would take no more space. But maybe some technicians prefer masses of figures in a table.

The book is handsomely produced, and in view of its level of presentation, it should prove a good investment to the technician.

N. H. CROWHURST.

Essentials of Microwaves

By R. B. Muchmore. 236 pp., 201 figs. Royal 8vo. Chapman & Hall. 1952. Price 36s.

THE extension of the radio spectrum to frequencies above 1 000Mc/s has led to many interesting developments and given rise to problems which at first sight bear little relation to those of ordinary radio frequencies. The basic laws which govern the behaviour of electro-magnetic waves are the same whatever the frequency, and for an understanding of microwave systems it is not necessary to introduce new principles but sufficient to make appropriate use of those already well understood at the lower frequencies. This is the essential theme on which the author of the book under review has based a very readable account of the mechanism of a wide range of microwave appliances.

The principal difference in outlook between the radio and the microwave engineer is that the former usually works with the circuit concepts of voltage and current, while the latter thinks in terms of electric and magnetic field strengths. That these stem from the common basis of the fundamental laws of electricity and magnetism is well brought out in the early chapters. The generation of waves on a transmission line is used to provide a convenient transition from the circuitry to the field approach and paves the way for an explanation of the operation of the basic microwave elements—waveguides and cavity resonators.

In the remainder of the book microwave oscillators and the miscellaneous components which are needed to make a practical system, such as a radar or a relay chain, are described in a similar way. The corresponding circuit element is examined, the factors which determine the highest frequency at which it may be used are explained and then field concepts are used to suggest the microwave equivalent. One chapter is devoted to the use of microwaves as a tool in aiding research into some of the most important problems of physics and finally an outline is given of the measurement techniques available to the microwave engineer.

The above summary gives some indication of the wide range of material covered in this volume. Throughout it is well written and the essential ideas are clearly explained without at any time introducing mathematical arguments. This book is admirably suited to a beginner in the subject, or to the engineer, used to ordinary radio frequencies, who wishes to find out what happens at the higher frequencies.

J. BROWN.

The Magnetic Circuit

By A. E. De Barr. 62 pp., 19 figs. Crown 8vo. The Institute of Physics. 1953. Price 5s.

THIS monograph is intended for general reading by students in Higher National Certificate in Applied Physics courses or the first two years of a degree course. An account is given of the way in which the properties of a magnetic material are affected by the way in which it is built up into a magnetic circuit. This is followed by descriptions of the properties of those materials which are most influenced in this way.

Soft Magnetic Materials used in Industry

By A. E. De Barr. 62 pp., 35 figs. Crown 8vo. The Institute of Physics. 1953. Price 5s.

THIS is the companion volume to "The Magnetic Circuit." The emphasis has been placed on the physical basis of magnetic properties and the aim has been to try to account, in terms of a physical theory of ferro-magnetism, for the particular properties of each of the modern soft magnetic materials with which the book deals.

ELECTRONIC EQUIPMENT

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

Psophometer

(Illustrated below)

THE S.T.C. 74142-A psophometer is designed to measure electrical noise levels in telephone and high-quality programme circuits in accordance with the latest (1951) C.C.I.F. specification. It incorporates weighting networks for both types of circuit, with facilities for switching in either of these, or an external network, or for switching to a position giving a flat response.

An additional facility provided by the psophometer is the measurement of alternating electromagnetic fields by means of a calibrated search coil.

The measuring range is 25μV to 10V while the input impedance is not less than 6kΩ in the frequency band of 40c/s to 5kc/s. The frequency range, as a flat amplifier, is approximately 50c/s to 12kc/s with a measuring accuracy of better than $\pm$1db over the entire range.

**Standard Telephones and Cables Ltd,
North Woolwich,
London, E.16.**

Close Accuracy Capacitors

(Illustrated below)

THESE T.C.C. close accuracy plastic film capacitors have been developed from the existing Plastopack series primarily for use in tuned filters for carrier telephony and in remote control gear for certain specialized military equipment. After a series of tests including temperature cycling between 15°C and 60°C the manufacturers can guarantee a stability of capacitance which is better than 0·1 per cent, while the power factor is down to ·0005 over a wide range of frequencies.

Since it is anticipated that designers will specify very precise values of capacitance they will normally be made to order and no definite range of values is published. For values greater than 0·02μF a tolerance of $\pm$0·125 per cent can be guaranteed, for lower values the tolerance will, of necessity, be greater.

The working voltage is 350V D.C. and the maximum working temperature 60°C.

**The Telegraph Condenser Co. Ltd,
North Acton,
London, W.3.**

Wideband Oscilloscope

(Illustrated below)

THE type 830 oscilloscope has been designed for general wideband frequency work and is particularly suitable for observing pulse waveforms with very fast rise-times. The frequency response of the Y amplifier is flat from 30c/s to 20Mc/s and the time-base provides writing speeds up to 30cm per microsecond.

The same mechanical design as that employed in the Airmec oscilloscope type 723 has been adopted, the cathode-ray tube being mounted vertically and viewed through a surface aluminized mirror. The instrument may therefore be used in conjunction with the Airmec oscilloscope camera type 758. The standard instrument is fitted with a fine focus cathode-ray tube having a 4 inch diameter flat screen providing a green trace. Blue (photographic) or long-persistence cathode-ray tubes may be fitted to special order.

The Y plate amplifier is a six stage unbalanced to balanced circuit. The sensitivity with a final anode voltage of 2kV and at a frequency of 100kc/s is greater than 75mV/cm on the normal position and approximately eleven times greater on high gain, the rise-time being 25msec.

The X plate amplifier has a frequency response of 30c/s to 1Mc/s $\pm$3db and a sensitivity at 100kc/s greater than 1V/cm.

The hard valve time-base has a sweep time range of 0·04sec to 1μsec, the highest writing speed fully expanded being 30cm/μsec. An expansion control is provided to expand the trace by up to five times. Single sweep triggered operation can be obtained using either internal or external syncronizing signals.

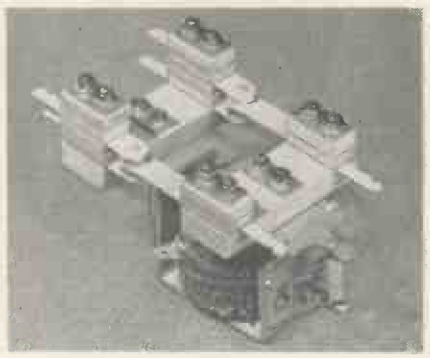
Alternative final anode voltages of 1, 2, or 4kV can be selected by means of a switch on the front panel.

**Airmec Ltd,
High Wycombe,
Buckinghamshire.**

High Voltage Relay

(Illustrated below)

THIS two pole change-over relay type 350 was designed primarily for aerial switching, but since a very high degree of insulation has been achieved, it has obviously many other applications. The design is based on the manufacturers standard 400 relay series and by using a new insulation material and wide spacing of contacts the following figures are achieved.



Breakdown voltage between fixed contacts and moving blades	3 000V.
Breakdown voltage between fixed contacts	12 000V.
Breakdown voltage between moving blades	14 000V.
Breakdown voltage between closed contacts and frame ...	7 000V.
Capacitance between open and moving blade	5·5pF.
Capacitance to frame when relay closed	4pF.
Operating time ...	13msec.
Time off	14msec.

**Electro Methods Ltd,
Caxton Way,
Stevenage, Hertfordshire.**

Vacuum Impregnator

(Illustrated top right)

THE Blickvac model 4 high vacuum impregnator has a chamber 14in in diameter by 14in deep and will impregnate coils to R.I.C. Specification 214. It is suitable for either laboratory or small scale production use. It will deal with waxes, hot or cold varnishes and thermosetting resins and has adequate trapping

arrangments for any of these materials. It has facilities for admitting dry air and for preventing moisture entering the pump. Direct indication of vacuum to 1mm of mercury absolute is provided.

Blickvac,
505 Lordship Lane,
London, S.E.22.

Coaxial Relay

(Illustrated below)

THE coaxial change-over relay type A.01 has a magnetic movement that provides adequate contact pressure over the normal range of aircraft supply voltages (18-28V). The contacts, which are rhodium plated, are rated at 5A. The relay can be supplied with a characteristic impedance of 45 or 70Ω. The R.F. losses are usually under 0·5db while the standing wave ratio measured at 200Mc/s is 1·08:1, the capacitance between open contacts being 0·05pF. It can be supplied fitted with either British Inter-Services or American JAN type sockets.

Besson and Robinson Ltd,
6 Government Buildings,
Kidbrook Park Road,
London, S.E.3.

Vitreous Enamelled Resistors

(Illustrated below)

THE construction of these resistors takes the form of a nickel-chrome winding on a ceramic former, capable of withstanding severe thermal shocks, covered by a protective vitreous enamel. They are available in the range of 5Ω to 33kΩ at a rating of 6 watts and 5Ω to 47kΩ at 10 watts. A 3 watt version is also manufactured. The normal tolerance is ±5 per cent of the stated value.

Labgear (Cambridge) Ltd,
Willow Place,
Cambridge.

Germanium Diodes

THE range of G.E.C. germanium diodes which are of the point-contact glass capsule type has recently been revised and extended by the addition of several new types.

Of the existing types, the GEX00 has been discontinued and the GEX44/1 replaced by the GEX34 for all purposes. The GEX34 is intended primarily for use as a television sound detector and sound noise limiter, and as a high level vision detector. It is capable of driving the sound output stage direct, where a sufficiently great R.F. input is available. The GEX35 remains the recommended type for low level vision detection and general purpose use.

The GEX45/1 and GEX55/1 high back resistance general purpose diodes are continued unchanged as is the GEX66 low-resistance V.H.F. mixer which will operate efficiently up to 1 000Mc/s and give a useful response at frequencies as high as 10 000Mc/s.

The GEX64 low resistance mixer is now supplied in groups matched for 5mA forward current in the voltage range 0·2 to 0·3V, for use in telephony modulators and similar bridge circuits.

The new types include the GEX36, a mixer diode and telephony modulator for use at higher voltage than the GEX64; it is available in groups matched for forward current at 5mA in the voltage range 0·675 to 0·875V.

A completely new group of diodes, the GEX54 group, comprises a number of high back voltage diodes, GEX54, GEX54/3, GEX54/4 and GEX54/5, which will operate at voltages of 80, 100, 150 and 200V respectively.

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

Miniature Transformers

TWO new miniature transformers have recently been introduced by John Bell & Croyden. The type O unit measures $\frac{3}{8}$in × $\frac{3}{8}$in × $\frac{1}{4}$in and it is an interstage transistor transformer or general coupling transformer. The inductance is 4H at 0·4mA over the normal audio frequency band. The step-down ratio is 4·5:1 with a D.C. resistance of 870 ohms primary and 170 ohms secondary. This transformer has a mumetal core, and can be supplied with a screening can if required.

The type A measures $\frac{3}{4}$in × 9/16in × 7/16in across the bobbin. It is an

Type O (*below*)
Type A (*right*

interstage transformer for matching a high gain pentode to a transistor, and has a primary of 125H at 50μA. The step-down ratio is 30: 1; the D.C. primary resistance is 6 000 ohms; the secondary resistance is 80 ohms.

Both of these transformers are illustrated at approximately their actual size.

John Bell and Croyden Ltd,
117 High Street,
Oxford.

Independent Sideband Receiver

(Illustrated below)

THE Mullard independent sideband receiver has been designed for continuous use on long distance circuits passing telephone and telegraph traffic, and it provides for the simultaneous and independent reception of four telephone channels of 3kc/s bandwidth, or two channels of broadcast quality of 6kc/s bandwidth. Alternatively each sideband can be used to accommodate several voice frequency telegraph channels.

The receiver covers the band 4-30Mc/s in three ranges. It is of the double superheterodyne type with intermediate frequencies of 3·1Mc/s and 100kc/s. The first oscillator can be used with crystal control, providing a choice of nine spot frequencies, or alternatively as a high stability variable oscillator. The second oscillator is self-excited, and operates at a nominal frequency of 3·0Mc/s. It has an A.F.C. motor-operated tuning capacitor covering a frequency range of ±4kc/s, and the design is such that the control is operative even during deep fades.

The output from the second I.F. amplifier is applied, via hybrid transformers, to upper and lower sideband and carrier filters employing X-cut crystals.

Each sideband amplifier is preceded by an attenuator so that sideband levels of up to 4db higher or 16db lower than normal can be corrected to give normal level at the demodulator. Forward acting A.G.C. is applied to the second amplifier valve to compensate for the rise of sideband level of about 6db that remains after the normal A.G.C. has operated on the R.F. units. Balanced diode demodulators, fed from a single hybrid transformer are employed in both channels, the A.F. output being almost independent of carrier level.

The equipment is mounted in a standard rack and cabinet, with doors at both sides and at the back.

Mullard Ltd,
Century House,
Shaftesbury Avenue,
London, W.C.2.

Notes from the Industry

Dr. B. V. Bowden, of Ferranti Ltd., has been appointed Principal of the College of Technology, Manchester, and will take up this new post on 1 September.

The Plessey Company Ltd. announce that arrangements have now been completed whereby Messrs. Amplion (1932) Ltd. are appointed distributors to the wholesale and retail trade for the sale of the comprehensive range of Plessey radio and television components.

The Canadian Broadcasting Corporation has selected Vancouver as the site for its fourth television station, and has ordered complete transmitter, studio and mobile broadcasting equipment from Marconi's Wireless Telegraph Co. Ltd., through the Canadian Marconi Company. The station will be ready to go on the air by the end of this year.

The British Standards Institution are now accommodated in their new premises at 2 Park Street, London, W.1. (Telephone Mayfair 9000.)

The Royal Society and Nuffield Foundation Commonwealth Bursaries Scheme. The objective of the scheme is to provide facilities for increasing the efficiency of investigators by enabling them to pursue research, learn techniques or follow other forms of study where either or both the physical and personal environment overseas in the Commonwealth is peculiarly favourable. The main difference between this scheme and the ordinary research fellowship is not merely one of duration but of emphasis as the bursaries will aim not so much at obtaining the answer to a particular question as at improving the powers of the recipient to extend the bounds of knowledge. Each bursary will provide for the cost of travel and maintenance normally for periods of two to twelve months. The applicant must be sponsored by a recognized research authority and must produce evidence that he or she has prior permission to work in the laboratory or other scientific institution chosen. It is proposed initially to consider applications at six-monthly intervals beginning early in 1954. Application forms containing further details will be obtainable from the Assistant Secretary, The Royal Society, Burlington House, Piccadilly, London, W.1, and must be submitted not later than 15 March and 15 September in each year.

Wild-Barfield Electric Furnaces Ltd. are again extending an invitation to senior students of technical colleges and similar institutions to visit their works during the coming winter season from October to May. The tours include an inspection of their Research and Development Departments as well as the production side and will take place during the afternoon of any week day. Those wishing to take advantage of this facility should apply to the Publicity Manager, Elecfurn Works, Watford By-Pass, Watford, Herts.

Seventh Electronics Course at Harwell. The Atomic Energy Research Establishment at Harwell is to hold its seventh specialized course on the design, use and maintenance of electronic instruments used in nuclear physics, radio-chemistry and work with radioisotopes. It will take place at the Isotope School from Monday, 28 September to Friday, 2 October. Applications are invited from physicists and electronic engineers holding a degree, or equivalent qualification. The fee for the course is 12 gns. Application forms can be obtained from the Electronics Division, A.E.R.E., Harwell, Nr. Didcot, Berks. These must be returned by Friday, 11 September, 1953.

The Ninth Annual National Electronics Conference will be held on 28, 29 and 30 September at the Hotel Sherman, Chicago. The technical programme offers 99 papers covering a broad field of electronic research, development and industrial application and is supplemented by over 140 exhibits by manufacturers foremost in the electronics field. The president of this year's conference is Dr. J. D. Ryder of the University of Illinois.

Meetings. The British Institution of Radio Engineers is holding a meeting on 30 September at 6.30 p.m. at the London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1. Dr. D. A. Bell will lecture on "The Impact of Communication Theory on Television".

The Electro-Physiological Technologists' Association is holding a General Meeting at the Burden Neurological Institute. Stoke Lane, Stapleton, Bristol, on Saturday. 19 September. at 10.30 a.m. There will be papers and demonstrations of interest to electro-physiologists and those interested in the application of electronic methods to statistics. Non-members are welcome at this meeting and should write to Mr. G. Johnson, the Honorary Secretary, at Hurstwood Park Hospital, Haywards Heath, Sussex.

The Television Society will exhibit their 405 line experimental transmitter which is being installed at the Norwood Technical College later this year for educational purposes and for the use of members wishing to gain experience on ultra short-wave reception. The vision carrier is 427Mc/s and the sound carrier 423.5Mc/s. An adaptor for reception on standard television receivers will also be shown.

Errata. Since we received the original information on the Bryan Savage V.L.F. Amplifier as published on p. 266 of the June, 1953, issue, we have now been informed by W. Bryan Savage Ltd. that the range has been extended as follows:

	Originally	As now supplied
Frequency response at 1000 watts	10c/s-1000c/s	6c/s-2000c/s
Maximum permissible output into zero power factor load	350VA	400VA

PUBLICATIONS RECEIVED

SELECTED PROBLEMS IN THE PREPARATION, PROPERTIES AND APPLICATION OF MATERIALS FOR RADIO PURPOSES is the twenty-fifth special report on radio research and has been produced by the Radio Materials Committee of the Radio Research Board of the Department of Scientific and Industrial Research. It is based on the work of several groups of experts who have studied the present state of knowledge of ceramics, organic polymeric dielectrics, magnetic materials and semi conductors. The report is divided into sections dealing with these materials, each section outlining existing knowledge and stating the research problems which most urgently need attention if the material is to be fully exploited for radio purposes. The report is published by Her Majesty's Stationery Office, price 1s. 6d.

WIGGIN NICKEL ALLOYS NO. 19 contains material of general interest to electrical and mechanical engineers including articles on pumping problems, furnace belts, industrial immersion heaters, diesel engine efficiency, recording thermometers and precision dispensing. A description of the Fawley Petroleum Refinery with statistical data is also included. Copies of this journal may be obtained from Henry Wiggin & Co. Ltd., Wiggin Street, Birmingham.

THE PETROLEUM INFORMATION BUREAU have published a list of their range of literature now available. This explains in non-technical language what is involved in the search for crude oil, its production, transportation and refining. The Petroleum Information Bureau, 29 New Bond Street, London, W.1.

EVERSHED ELECTRONIC REPEATER, Publication No. 266, describes the range of electronic repeater equipment made by Evershed & Vignoles Ltd., and gives examples of its various uses.

EVERSHED "DUCTER" HANDBOOK, Publication No. 269, describes the Evershed "Ducter" Low Resistance Test Set, printed in pocket book size, giving full details of its operation and examples of its uses. Evershed & Vignoles Limited, Acton Lane Works, Chiswick, London, W.4.

TEMPERATURE TESTING WITH THERMOCOLOR and THERMOCRON are two booklets describing temperature indicating paints and crayons recently introduced by Allied Colloids (Bradford) Ltd. These products are applied in the form of a paint with industrial methylated spirits and give accurate indication of thermal zones under observation. The Thermocolors are particularly useful for non-destructive material testing and for control work in industry generally, wherever heat distribution presents a problem. Allied Colloids (Bradford) Ltd., 11 Great St. Thomas Apostle, Queen Street, London, E.C.4.

CODES DIMENSIONS AND WEIGHTS OF RECTIFIER STACKS gives the dimensions and weights of SenTerCel spindle mounted rectifier stacks. It deals primarily with standard stacks and contains a complete explanation of the coding system which is used to describe them. Standard Telephones and Cables Limited, Rectifier Division, Warwick Road, Boreham Wood, Herts.

ENGINEERING EDUCATION IN THE REGION describes the engineering education in technical institutes and colleges and university colleges in London and the Home Counties. Copies of this booklet may be obtained from the Secretary, Regional Advisory Council for Higher Technological Education, Tavistock House South, Tavistock Square, London, W.C.1, price 1s.

REPORT ON RESEARCH WORK IN THE CITY AND GUILDS COLLEGE describes the research carried out in the Departments of Aeronautics, Chemical Engineering and Applied Chemistry, Civil, Electrical, Mechanical Engineering and Mathematics from 1946-51. Imperial College, South Kensington, London, S.W.7.

THE GRADUATE IN THE G.E.C. is a booklet giving details of the graduate's prospects with this company and the facilities for training and experience. The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2.

1953 CATALOGUE OF BOOKS ON ELECTRICAL ENGINEERING, RADIO AND ELECTRONICS includes the titles of works issued by various publishers. The catalogue has been produced by H. K. Lewis & Co. Ltd., 136 Gower Street, London, W.C.1, and orders for any of the publications listed should be placed through them.