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Commentary

IN our last issue comment was made on the Jubilee of the invention of the thermionic valve and a brief account was published of the life and work of the late Sir Ambrose Fleming. Such anniversaries provide an occasion for a review of historical development and, in the better perspective which the passage of time makes possible, it not infrequently happens that there are lessons to be learned therefrom. For thirty years or more, Fleming has been widely acknowledged, at least in this country, as the inventor of the valve, but scientific history has so often been distorted by commercial interests, by the need to establish patent claims or by the needs of publicity that it is not unreasonable at least to enquire into the circumstances in which Fleming's invention was made. At the risk of being accused of detracting from the merit of Fleming's invention, let us briefly review the background.

Fleming's work in this field had begun as early as 1882 when, at the age of 32, he had joined the Edison Electric Light Company and become engaged in the many problems then awaiting solution with the early electric lamps. Among these problems was the prevention of the blackening of the inside of the bulbs and, in particular, the investigation of the causes of the clear white line which was sometimes noticed on the inside and which seemed to be a shadow of one leg of the filament. Edison himself carried out some experiments on the subject in 1884, in the course of which he sealed a sheet of metal foil inside a lamp and observed that if this electrode was connected to the positive leg of the filament through a sensitive galvanometer a small current could be detected but that no current flowed if it was connected to the negative leg.

During a period of eight or ten years, Fleming carried out many researches in this field and, in 1889, he had constructed a whole series of special lamps with variously shaped electrodes within the bulb. He thus became intimately familiar with the conditions under which a current could pass from the heated filament to a nearby anode. Even if the exact mechanism of thermionic emission was not fully understood—the electron was not discovered until 1897—Fleming's work on the Edison effect in 1889 had clearly shown that a current could only pass between filament and anode if the anode was positive in respect to the filament. He knew, in fact, in 1889 that an electric lamp with an electrode inside would act as a rectifier of alternating current.

In 1900, Fleming joined the Marconi Company as a consultant and he was largely responsible for the design of the transmitter at Poldhu, from which the first wireless signals were sent across the Atlantic. Intimately aware of the capricious behaviour of the coherer detectors then in use and of the urgent need for a more reliable detector, it seems remarkable that it was not until October, 1904, that he thought of trying one of his experimental lamps, made fifteen years earlier, as a rectifier of high frequency currents. The experiment was immediately successful—the Fleming Oscillation Valve was born—but if it is surprising that the idea should not have occurred to Fleming earlier, it is still more remarkable that he should have made so little effort subsequently to develop and improve on his discovery. With all the background of his previous work on the Edison Effect to guide him and with all the facilities that he had available, it does seem incredible that he should have done so little towards its further development.

The fact is, of course, that Fleming had invented nothing new ; all that he had done was to apply a known device to a new application and since the Fleming valve was no more sensitive than the magnetic detector which had been introduced two years earlier, it is not surprising that it was regarded at the time as merely an interesting but unimportant alternative.

Fleming's application was a novel one—but all wireless telegraphy was a novelty in those days. Nevertheless, if the Patent Office examiners had been able to view his application with the perspective we can apply today, we cannot help wondering whether they would have granted him a Patent at all. Provocative though this thought may be, it is a fact that Fleming's Patent of 1904 was declared invalid by the Supreme Court of the U.S.A. as recently as 1943 on the grounds that he had applied for a patent for the valve as a rectifier of high-frequency oscillations without disclaiming its already known use as a low-frequency rectifier.

Why did it take Fleming so long to apply his experimental lamps to the detection of high-frequency oscillations ? Why, with all his background of experience in this field and all the resources at his command did he so completely neglect the subsequent development of the valve ? We may never know the answers to these questions but surely the moral is clear. Let every scientist who is called upon to work in some new field consider whether he has not some special knowledge gained from past experience to apply and develop to advantage in his new surroundings and, having found some new and useful application, let him pursue it, develop it and exploit it to its uttermost capacity.

Three-Phase High-speed Magnetic Amplifiers

By A. E. Maine*, A.R.Ae.S., A.Brit.I.R.E.

The article describes the extension of the principles of "Half-Wave" High-Speed Magnetic Amplifiers to 3-phase circuits. It is shown that there are many arrangements possible for 3-phase operation and these offer scope for continued development and application especially in the field of high-speed, high-power control systems. A 50c/s multi-stage unit of this type is described giving a power output of about 1kW, and used for supplying the armature requirements of a ¾ h.p. D.C. motor forming part of a fast acting servo system. The article is based largely upon the results of an experimental study, and represents early work carried out in a relatively new branch of the magnetic amplifier field.

SEVERAL papers have appeared in the literature over the course of the last few years dealing with high speed magnetic amplifiers of the "half-wave" type. Such amplifiers fall naturally into a class of their own since they possess several well-defined features. Probably their most important characteristic is their relatively fast response to a change of input signal. Here we find that full response can be obtained within the time of half a cycle of the A.C. excitation frequency. This fast action points significantly to the basic mechanism involved, i.e. a system wherein the control intelligence is impressed in a magnetic core during one half cycle (sampling) and acted upon in the half cycle following (gating). The medium involved in the process is, of course, the flux in the core, which latter is assumed to be highly remanent and sharply saturating.

The volt-seconds storage of a core is directly proportional to the maximum (knee-to-knee) flux swing and the number of turns, and usually this quantity is made to correspond to the voltage-time integral of the supply wave over one half cycle. It follows then, that if at the beginning of a gating period, the flux starting point is other than at the lower knee, a fraction of the applied voltage integral cannot be absorbed by the reactor, and must, therefore, appear across whatever impedance there is in the circuit. Load current consequently flows around the circuit, the reactor having substantially zero impedance in this interval.

The scheme of control may now be simply stated. An input signal is allowed to set the flux level in a "negative" direction during sampling. This level determines the angular instant in the following half cycle, with the flux going positive, when the reactor "fires". Since only magnetizing currents need flow during the sampling half cycle, and load currents limited only by the thermal rating of the reactor are possible during gating, it follows that the simple system described acts as a power amplifier. The separation of the two processes in time is accomplished by means of rectifiers connected in appropriate directions to the control and load circuits, both of which contain an A.C. source. The auxiliary A.C. voltage acting in the control circuit provides the necessary core magnetization to ensure full "re-set" of the flux during sampling, and therefore minimum load current, when the signal voltage is zero. It is to be noticed that during the sampling period, the control voltage is applied in series opposition to the auxiliary voltage and it is the difference between these which determines the re-set flux level, and subsequently the average value of the load voltage. A simple amplifier arrangement of the sort described is shown in Fig. 1(a) and represents a foundation for the development of more elaborate arrangements.

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The Three-Phase Connexion

A "full-wave" single-phase amplifier may be set out in bridge form as shown in Fig. 1(b). Here, four cores are used and because two coils are effectively in series in the two branches, each reactor is designed to absorb only one half of the applied voltage integral. Examining the circuit, it is evident that with the polarities given cores 1 and 4 are in a gating period while cores 2 and 3 are being re-set. Inverting

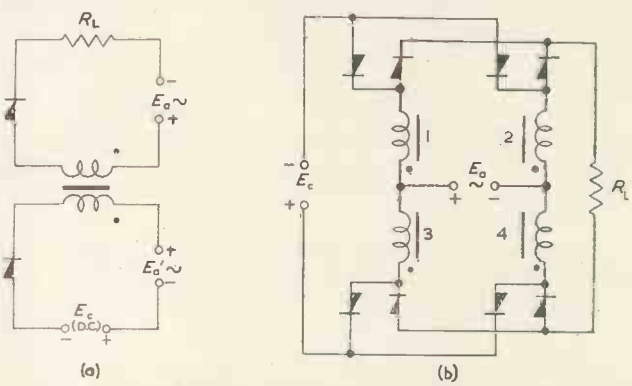


Fig. 1(a). Simple half-wave single core amplifier. (b). Bridge connected full-wave amplifier

the polarities of the supply terminals for the next half cycle, shows that the pairs of reactors exchange their operation. The load voltage will evidently be similar in form to that delivered by a single-phase grid controlled rectifier.

The functions described above can be secured with a more refined circuit but that shown in Fig. 1(b) has been deliberately chosen since it leads easily to the introduction of amplifiers for 3-phase operation. Here we add a third branch to the basic bridge and join its mid-point to the third phase of the supply. The new circuit is illustrated in Fig. 2(a) and this constitutes a full-wave three-phase amplifier possessing the usual half-cycle response and giving a voltage gain of unity.

An experimental amplifier of this sort operating at 400c/s and employing 1½in square H.C.R. laminations has been built and used to verify the dynamic response. The output power was about 10 watts and germanium junction diodes were used throughout. Photographs of the output voltage waveform during the application of step changes of control voltage showed clearly the half cycle response of the amplifier. The harmonic response tests yielded interesting results and suggest that the device behaves substantially as a pure time delay of $1/2f$ seconds. The amplitude ratio is shown in Fig. 3 and exhibits an unexplained rise before the fall away at approximately 20db/decade. The phase measurement shown in the inset was obtained by means of a Lissajous figure derived from the input and output voltages. The technique consisted of raising the test signal frequency and noting the points at which successive 90° phase shifts were obtained. It might be mentioned that a large number of "beats" were observed above a signal frequency of 400c/s, these being due to modulation products of the signal and sampling frequencies and their harmonics. In contrast to the single-phase version the beats were found to be of generally small amplitude. It appears therefore that the dynamics of the three-phase version differ somewhat from those of the single-phase arrangement and it is perhaps remarkable that the amplifier offers considerable response to signal frequencies well above that of the excitation source.

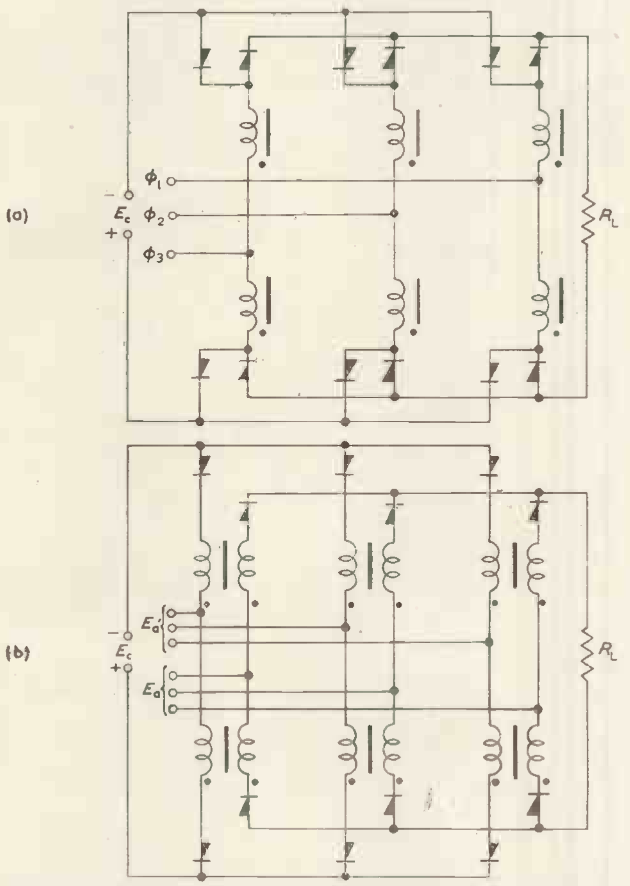


Fig. 2(a). Full-wave, three-phase magnetic amplifier. (b). Full-wave, three-phase amplifier with separate control coils

Often it is desirable to achieve galvanic isolation between the load and control circuits and, in addition, a voltage gain greater than unity may be required. Both of these needs may be fulfilled by the arrangement of Fig. 2(b) which employs double wound reactors. In this case a second three-phase re-setting supply is needed having a voltage $1/N$ of that of the load circuit, where N is the ratio between load and control turns. Both of the amplifiers described may be "resistance" or voltage controlled. In the first case a variable resistance is connected between the control terminals, and hence the magnetizing current flows through it during the sampling periods. The voltage developed across the resistance is not available for influencing the core flux, hence the greater the resistance, the less the re-setting action and the greater the load voltage. This type of operation is not often employed, voltage control being generally more useful. In this case the control source must have a sufficiently low impedance to prevent the loss of a significant amount of voltage from the reactors. The source voltage may take the form of (a) pure D.C., (b) three-phase full-wave, rectified and variable amplitude, or (c) three-phase, fixed maximum amplitude, rectified and variable conduction angle, e.g. the output voltage of a similar amplifier.

It will have been noticed that the amplifier configurations discussed so far follow the patterns of well-known rectifier

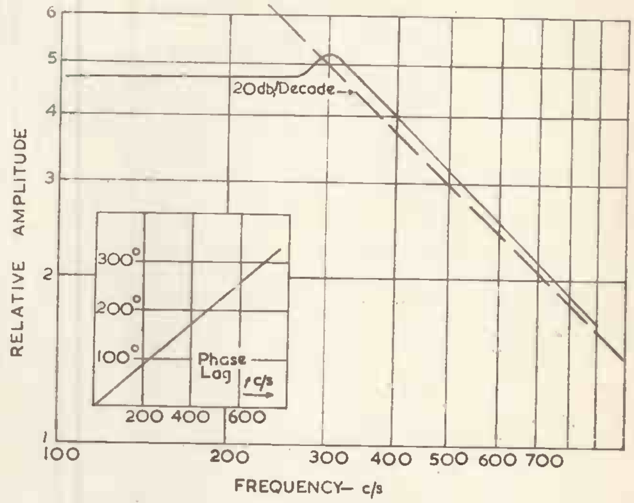


Fig. 3. Dynamic response of three-phase full-wave amplifier

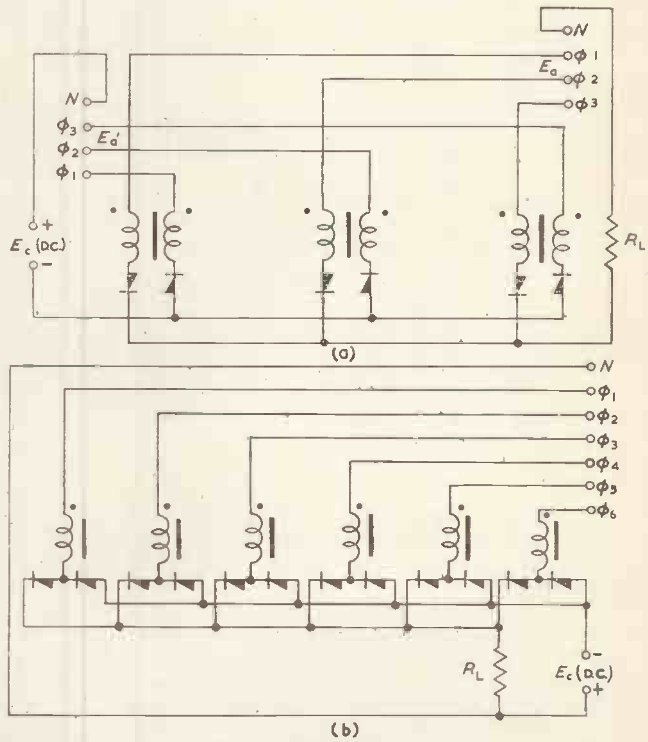


Fig. 4(a). Three-phase, half-wave amplifier with separate control coils. (b) Six-phase, half-wave amplifier

arrangements and one might expect other rectifier connexions to be equally applicable in this respect. This is indeed the case, and it is easy, for example, to set out half-wave versions such as those shown in Fig. 4. These amplifiers are perfectly sound in all respects but suffer the dual disadvantages that a neutral line is required, and the average load voltage, other things being equal, is reduced. In fact it might be said that the important features of the rectifier arrangements apply equally to high-speed amplifiers based upon them.

Core Materials and Constant M.M.F. Bias

Since the main application of the multi-phase amplifier is undoubtedly in the medium and high power fields, it follows that relatively large reactors with heavy cores are normally required. For the best possible results nickel-iron alloys, e.g. Permalloy " F " or H.C.R. are needed for the core construction and if such materials are used the reactors are liable to be rather costly. Reasonable alternatives to the nickel-iron group may be found in the grain oriented silicon steels, and type 62 Crystalloy is perhaps the most promising material in this category, at least for mains frequency operation. This material has a peak flux density of about 13 kilogauss and close upon 95 per cent remanence. In comparison with H.C.R. the hysteresis loop is very fat and the permeability at the flanks considerably smaller. From this it is evident that the standing control currents are relatively large and the power gain reduced. A worthwhile improvement may, however, be brought about by means of supplying the core with a constant M.M.F. bias. This has the effect of moving the hysteresis loop bodily along the *H* axis, the excursion being terminated when remanence loss becomes significant. At this point it is found that the control circuit magnetizing current is appreciably reduced, whereas the magnetizing current in the pre-firing gating period is correspondingly raised. The scheme serves the purpose of reducing the control current but renders the amplifier more sensitive to load rectifier leakage. In practice the required bias is obtained by joining a source of direct current to auxiliary windings on the reactors, the high circuit impedance ensures that alternating components induced into the bias windings are ineffective in altering the core state. Experiments show that a reduction of control current by a factor of two or three is possible.

Core Balance

For best results with the three-phase circuits, it is desirable to use sets of balanced cores in the amplifiers. In the case of laminated reactors, balance may be readily obtained by means of adjusting the number of laminations in each core so that they all saturate at the same value of applied voltage. Matters are not so straightforward if the spiral tape core construction is used, and here it becomes necessary to select similar cores and perhaps trim the number of turns on each. If large quantities of such cores are required it would be desirable to test each core before winding and make appropriate adjustments to the number of turns during winding. The balancing procedure eliminates departures from linearity of the voltage transfer characteristic, particularly in the region of low output voltage.

Rectifiers

The rectifiers employed in Ramey[1-5] type half-wave amplifiers have a dominating effect on overall performance. This is especially apparent when using rectangular loop nickel-iron core materials. Spurious re-setting will always occur if the reverse leakage current is equal to or greater than half the width of the hysteresis loop when the *H* axis is calibrated in current for a given number of turns. To underline the importance of this point the case may be quoted of a high power amplifier designed to yield a maximum load current of 2A where the " ½-loop " value of magnetizing current was only 2mA. In cases of this sort it is necessary to either halve the manufacturers' voltage ratings for selenium rectifiers or possibly to use modified forms of amplifiers such as those proposed by Scorgie and House. Increasing the number of rectifier plates involves a worsening of overall efficiency which may drop well below 70 per cent. In high power applications the loss is undesirable to say the least and may well be serious. The capacitance of selenium rectifiers, even at power frequencies can be troublesome—in one case a 1kW 15A rectifier set was found to have a capacitance of 2μF per arm and this, too, produced unintentional reset thereby reducing the maximum output voltage. As a general rule, whether or not special circuit techniques are used it is desirable to work metal rectifiers at considerably reduced voltage ratings and to select stacks having the smallest possible leakage current. The control circuit rectifiers are somewhat less of a problem, especially as germanium junction units can usually be used here. These rectifiers in comparison with selenium types have very much improved characteristics and possess a high front to back ratio and very low forward power loss. Reviewing the whole question of rectifiers it seems that very much better units are needed if the potentialities of the circuits discussed are to be fully exploited. Large germanium[6] and silicon junction units may be promising in this respect.

Flux " Pre-Set " with Three-Phase Amplifiers

It has been stated previously that a highly remanent core material is necessary for the basic modes of action discussed. However, it is shown in a paper by House[7], that single-phase half-cycle response amplifiers may be constructed using core materials possessing zero remanence. Simply stated the necessary flux conditions are brought about by means of

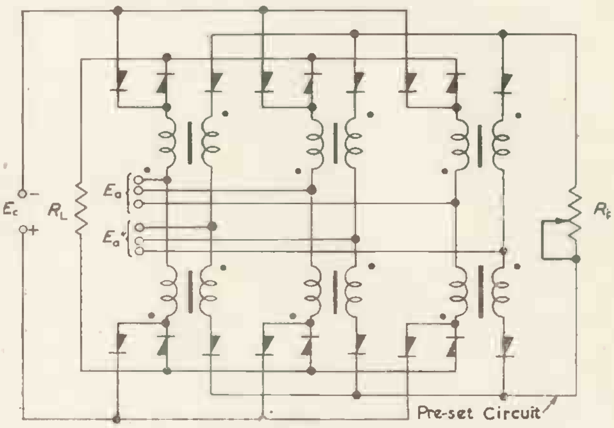


Fig. 5. Three-phase amplifier showing flux ' pre-set ' proposal

reversing the sense of the control coils, thus permitting an upward movement of the flux towards positive saturation during sampling. House describes the new operation as *pre-setting* and shows that the flux may indeed be set to any desired level by the onset of the gating period. It is also shown that partially remanent cores may be used under specified circumstances in arrangements which contain both re-set and pre-set circuits. A study of these proposals has indicated that such methods should be equally valid when applied to three-phase systems and hence it seems more likely that less refined core materials may be used successfully for large power applications. Furthermore, the spurious re-setting arising out of leaking main rectifiers can be largely off-set by these methods. A proposal showing the addition of flux pre-set to a three-phase circuit is given in Fig. 5. It is implied that only a relatively small remanence loss is involved, the pre-set circuit being adjusted to give the required amount of " flux-lift " to enable full output voltage to be achieved. Schemes of this sort are still under development.

The Three-Phase Amplifier with Negative Voltage Feedback

In the circuit types considered in the preceding paragraphs, a limitation to low-level signal working is imposed by the control rectifier characteristics which become unsatisfactory when the circuit voltages are small. The maximum voltage gain normally obtainable is usually determined by these considerations alone. Scorgie[8] has set out a means of overcoming this difficulty which consists of eliminating the control rectifiers altogether, and applying 100 per cent negative voltage feedback over the amplifier. The arrangement, which in some respects is more fundamental than

Ramey's schemes does not require an additional magnetizing source in the control circuit. In operation, the magnetizing voltage is obtained jointly by induced voltages from other non-fired cores while in their gating periods, and by the load voltage itself : the two components are furnished in time sequence. Such a scheme is applicable to the three-phase amplifier and a circuit modified in this way is shown in Fig. 6. It is to be noted that for a turns ratio of N (previously defined) the fraction of the load voltage required to give 100 per cent negative feedback is $1/N$, if the turns ratio is greater than unity, and the load appreciably inductive, difficulties arise since the arrangement becomes unstable, behaving as a second order system with only slight damping. A possible way out of this trouble is to wind the reactors with extra coils connected to form an auxiliary load circuit and deriving the feedback voltage from this.

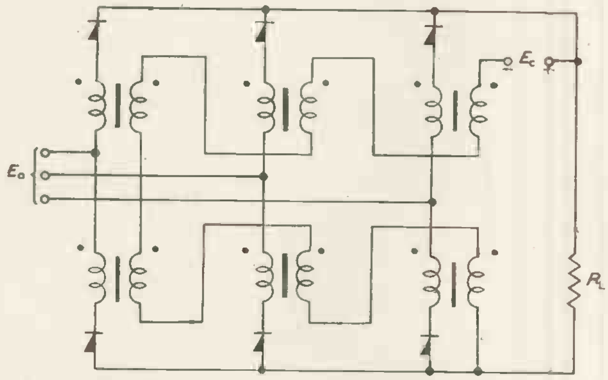


Fig. 6. Three-phase amplifier with negative voltage feedback

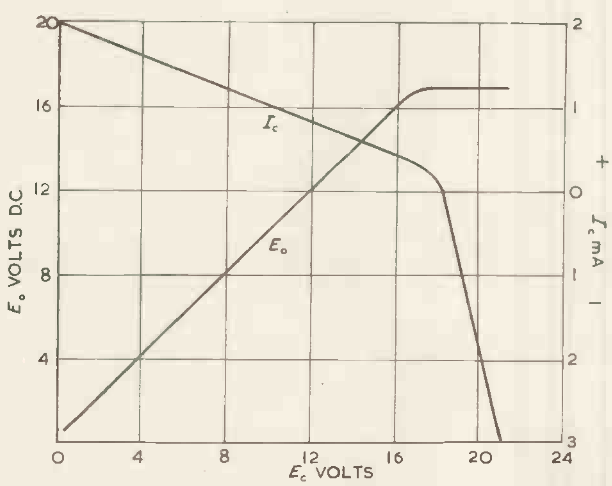


Fig. 7. Characteristics of three-phase amplifier with negative voltage feedback (H.C.R. cores)

The elimination of rectifiers from the control circuit serves a further important function in that it permits the control magnetizing current to assume any magnitude and direction necessary to support the flux conditions imposed by the applied voltages. As a consequence the amplifier maintains its essential features even when employing substantially non-remanent core materials. Fig. 7 shows the measured characteristics obtained from the circuit arrangement of Fig. 6 and employing laminated H.C.R. reactors. A duplicate experiment carried out but employing Mumetal cores showed that the form of the voltage transfer characteristic remained unchanged, but the control current reversed sign at a much lower value of control voltage. These results are important because, in common with the proposals of House they point the way to the use of lower grade core materials. It must be pointed out, however, that a rectangular loop material is generally to be preferred since the control current assumes a very simple form and the amplifier is amenable to the employment of M.M.F. bias and positive feedback schemes aimed at further increasing the gain.

The Three-Phase Amplifier with Positive Current Feedback

Elementary analysis of the half-wave amplifier suggests that if the power gain is raised by increasing the current gain, the half cycle response should be preserved. Such a proposition is put forth by Scorgie[7] and involves the use of positive current feedback. This scheme has been tried experimentally with the three-phase circuits, and appreciable increases in steady state gain observed. Fig. 8(a) shows the equivalent circuit of an arrangement wherein the load voltage is coupled back to the control circuit through a high-valued coupling resistance and the control source is shunted with a constant current. Fig. 8(b) shows the idealized construction yielding

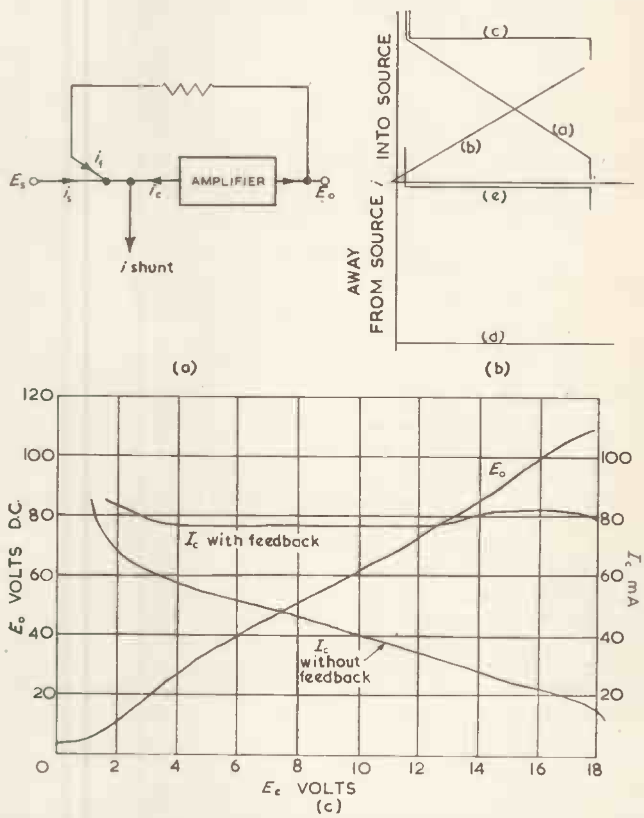


Fig. 8(a). Equivalent circuit of amplifier with positive current feedback. (b). Idealized construction showing effect of positive current feedback. (c). Measured characteristics of three-phase amplifier with positive current feedback

infinite power gain. Curve (a) is the typical form taken by the control current, and curve (b) the feedback current obtained from the load, and adjusted by the coupling resistor to have equal but opposite slope. Curve (c) is the sum of (a) and (b) and indicates the net core M.M.F. Curve (d) represents the constant current shunt. Finally, curve (e) shows the sum of (c) and (d) and is drawn displaced away from the horizontal axis only for reasons of pictorial clarity. Fig. 8(c) shows a practical characteristic obtained experimentally but without the constant current shunt. It is evident that the gain measured in terms of $\Delta I_L/\Delta I_c$ has been increased to infinity over a wide portion of the range. The amplifier used for the tests was the 400c/s unit previously mentioned. Investigation of the dynamics yielded the important result that the half-cycle response, with the current feedback was only obtainable when the signal source impedance was substantially zero. With realistic values of source impedance the control voltage is changed from its proper value in the transient condition, due to the feedback, and the half-cycle response is lost. Apart from the use as an inter-stage amplifier, where the driving source impedance can be kept reasonably small, it is thought that the technique has rather limited usefulness.

Coupling Two or More Stages

The general order of power gain obtainable from the standard full-wave three-phase amplifier is some hundreds and may reach 1 000, but hardly any more. This is usually too small for practical purposes and so a two- or three-stage design may be required. The question of coupling amplifiers of this sort (or indeed most types of high-speed amplifier) is not straightforward since the load rectifiers of the driver stage face into the control rectifiers of the power stage[4], and this prevents the proper circulation of magnetizing currents in the two amplifiers. A shunt resistance in the

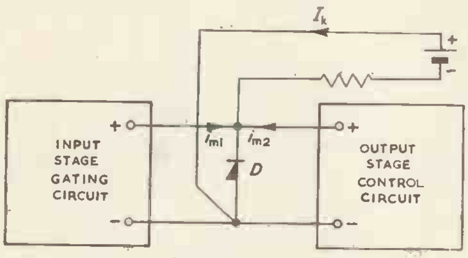


Fig. 9. Interstage coupling circuit

coupling is a possible way out of the difficulty, but is wasteful and reduces power gain. The problem may be satisfactorily solved by means of using a " constant current " coupling device. In its elementary form this consists of a rectifier supplied with current from a high impedance source : the general principle of operation may be outlined by reference to Fig. 9. Here the constant current I_K is adjusted through the diode D to a value that just corresponds to the sum of the maximum magnetizing currents of the two stages. Assuming for the moment that no reactors in the system fire in the gating periods, the magnetizing currents of the two amplifiers flow into the diode " reverse-on " without interaction. If now the driver reactors fire, the coupling diode becomes blocked, and the flow of magnetizing current in the power stage is stopped. This prevents any further change of flux level in the power stage for the rest of the half-cycle. It follows then, that the firing instant of the pre-amplifier during gating determines the flux level set in the power stage during sampling, hence the necessary coupling conditions are satisfied. If a coupling of he sort described is employed in an arrangement where the power stage feeds the armature of a D.C. motor, the scheme outlined breaks down and a considerable time lag is introduced into the system. This is because of the blocking action of the motor back E.M.F.

The mechanism involved becomes more complicated to follow when dealing with a three-phase system hence the half-wave model shown in Fig. 10 will be used for the purposes of explanation.

The mode of action is shown in Fig. 11 and treats the case where a step input is applied to the first stage demanding a reduction of motor speed from maximum to ½ maximum. The operation is shown in five epochs :

- A Steady state—motor at maximum speed.
- B Transient—step signal input, motor back E.M.F. not appreciably changed.
- C Transient—motor speed and back E.M.F. falling.
- D Transient—motor runs underspeed.
- E Steady state—motor at half maximum speed.

These divisions in time are somewhat artificial but assist in clarifying the operation.

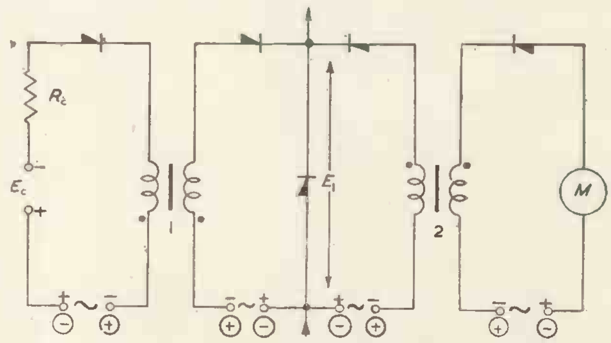


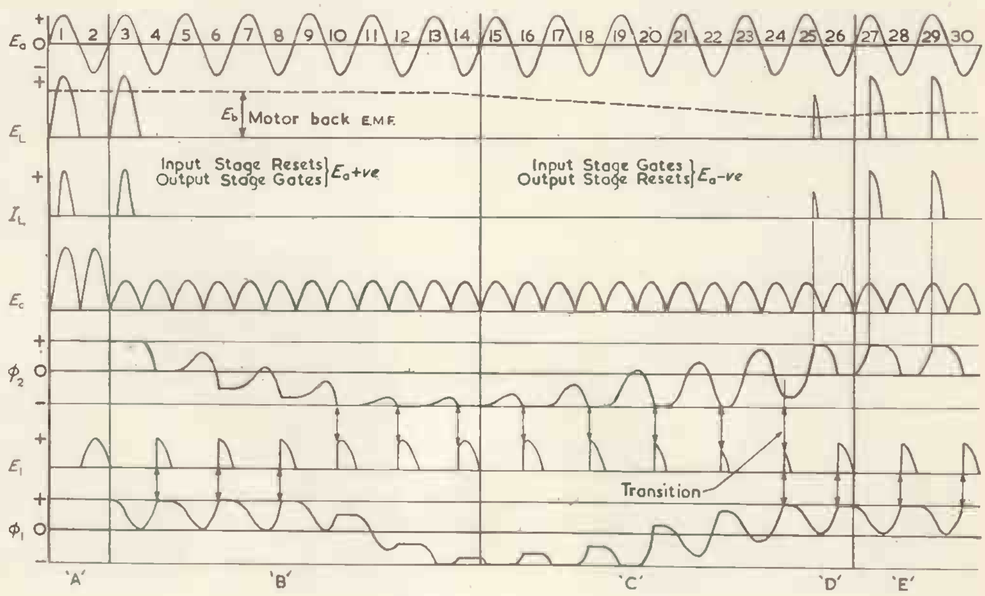
Fig. 10. Single-phase, half-wave model of coupled amplifier

In the first period A, the system is in steady state and no flux changes occur at all in the cores since both stages are delivering maximum voltage.

In period B the flux in core 2 is re-set in the normal manner during ½ wave No. 4 ; in the following gating period the flux in core 2 is not swept up to positive saturation because the voltage-time integral applied to the core is less than that applied during sampling and is given by $\int_0^{\pi} E_A.dt. - E_b/2f$.

From here until ½ wave No. 10, the flux in core 2 is " spiralled " down to negative saturation, where an important change of circuit operation takes place, i.e. the *power* stage blocks the coupling diode and freezes the flux level in the driver stage.

Fig. 11. Construction showing mechanism of " flux depression "



This condition repeats for successive cycles until the flux in the first stage is also sent to negative saturation. If the control circuit impedance of the first stage is low, large circulating currents result.

In period C it is assumed that the motor back E.M.F. is falling, consequently the voltage-time integral applied to core 2 increases during each successive gating period. The fluxes in both cores, therefore, approach progressively nearer to positive saturation and the power stage gradually relinquishes control of the coupling diode. At the end of the period, core 1 fires before core 2 can reach negative saturation so it abruptly seizes control of the coupling means. The actual transition is shown in $\frac{1}{2}$-wave No. 24.

Period D shows a transition phase with the motor underspeed and the current pulses to the armature just breaking through. This period merges into the final phase with the system in steady state at half output and the appropriate changes of flux taking place in the two cores.

The phenomenon described may be termed "flux depression" and unless counter-measures are taken, all the advantages of a high-speed amplifier drive are lost. The means adopted for overcoming this difficulty are set out

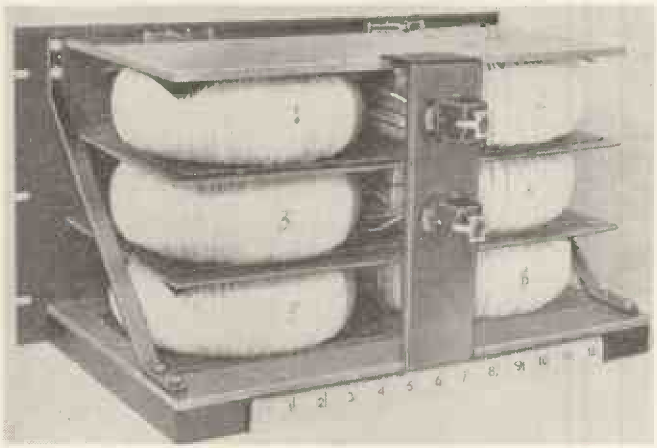


Fig. 12. Basic transductor assembly

later and are shown for a two-stage three-phase amplifier; the broad principles, however, are generally applicable.

Experimental Amplifiers

In order to check experimentally the validity of the various three-phase arrangements, and also to establish the form of the control voltage/current characteristics, a set of reactors working at 400c/s and handling an output power of about 10 watts were used. These early experiments confirmed that the circuits were basically sound. They also demonstrated the rather surprising fact that the magnitude and form of the I_C/E_C characteristics of all the configurations tested, including the single-phase full-wave circuit were substantially the same. However, in order to obtain a better evaluation of the three-phase high-speed magnetic amplifier it was decided to design a 50c/s multi-stage unit handling an output power of about 1kW and used for feeding the armature of a separately excited D.C. motor rated at $\frac{3}{4}$ h.p. The motor itself is embodied into a speed control system for driving a small 1 600c/s generator or alternatively for swinging the beam of a component test centrifuge. In the first case a constant speed is required, and in the second, a speed range of 20 : 1 is needed. Both schemes involve the use of three-stage amplifiers, the power output and driver stages being three-phase full-wave units in each case. The first stage for the alternator application is a conventional magnetic amplifier working at 1 600c/s and for the centrifuge the pre-amplifier is electronic, operating directly from three-phase rectifiers without smoothing. Of the two control systems, the centrifuge application is the more advanced at the present time being in the final stages of detailed engineering; for this reason the following sections will be concerned with this particular application only.

Practical Application of the Three-Phase Magnetic Amplifier

Power and Driver Stages

The power stage reactors employ Permalloy "F" spiral tape cores 6$\frac{1}{4}$in O.D., 5in I.D. and 1$\frac{3}{8}$in axial length, and were specially made by S.T. & C. Ltd. The core boxes were made by winding glass fabric tape impregnated with a thermosetting polyester resin around a split ring and parting the case off, inside and out, in a lathe after the resin had set. The wall thickness is $\frac{1}{8}$in and the finished item was found to be very strong. Each core, after boxing was hand wound with 2 000 turns of 18 S.W.G. wire for the load coils and 90 turns

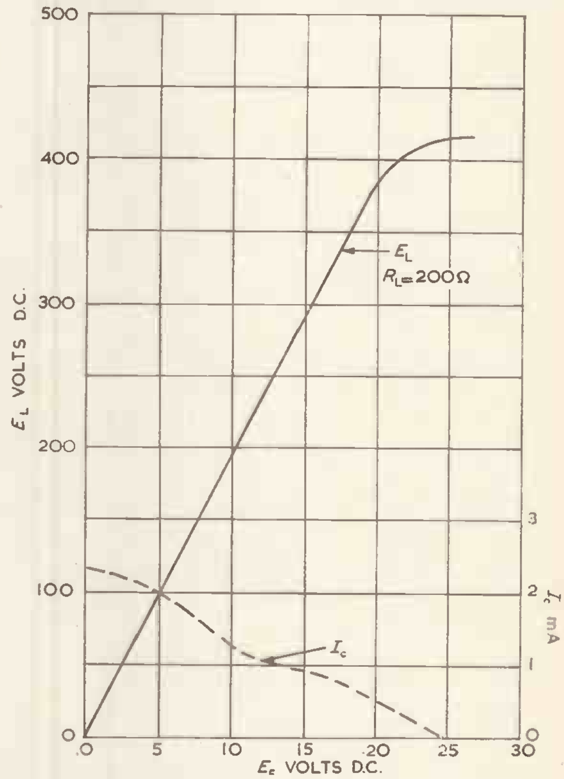


Fig. 13. Transfer characteristic of driver and power stages in cascade

of 16 S.W.G. for the control. Six units, for direct connexion to the three-phase 440V main were then mounted on a modified 19in panel to form the basic transductor assembly shown in Fig. 12. The load rectifiers comprise six pairs of Westinghouse type 13C35 units enclosed in a small oil-cooled tank 3$\frac{1}{2}$in deep and mounted adjacent to the transductor unit. The voltage transfer characteristic was found to be exceedingly linear, no discrepancies being observable when employing normal laboratory multi-range meters.

The driver stage employs $\frac{3}{8}$in stacks of 1$\frac{1}{2}$in square H.C.R. laminations and employs B.T.H. germanium junction rectifiers throughout. The turns ratio is 1$\frac{1}{4}$: 1, and the general construction follows normal transformer practice. This amplifier, together with the remainder of the circuit. is housed on a separate panel below the other two. The transfer characteristic of the two amplifiers connected in cascade is shown in Fig. 13. From this it is evident that the power gain is 16 000 (for a "one-cycle" response time), when a 200 ohm load resistor is connected to the output stage. With larger main rectifiers and an input stage based on the use of torroids instead of laminated reactors it is likely that a power gain in the region of 40 000 could be secured. It was found that the transfer characteristic remained unchanged when supplying the armature of the motor running near maximum load conditions.

The Electronic Pre-amplifier Stage

The electronic pre-amplifier stage consists of a single 12AT7 valve executing the functions of triode pre-amplifier and cathode-follower drive to the first magnetic stage. The circuit follows normal practice, except perhaps, that the grid and cathode of the cathode-follower are clamped by means of biased rectifiers to restrict the range of control voltage. The rectifier MR_3 (Fig. 14) connected to the control circuit of the first magnetic stage is a control current limiter and prevents the flow of excessive currents during the brief cathode warm-up period. Incidentally, the transfer characteristic of Fig. 13 was obtained when using the cathode-follower drive.

possible. The modification ensures that blocking of the coupling means is possible in a "forward" direction only, e.g. going from driver to power stage, but not in the opposite direction. This is accomplished by splitting the single coupling path into two parts separated by an additional rectifier and each taken to a common high voltage supply through independent current limiting resistors. On the complete circuit diagram of Fig. 14, the elements involved are MR_4 and MR_5 and resistors R_2 and R_3. It is easily seen that if the driver reactors fire both rectifiers MR_4 and MR_5 become blocked, whereas if the power stage saturates in the negative direction MR_7 prevents MR_5 from being blocked, hence the driver gating magnetizing current is not interrupted. Once again rectifiers MR_4 and MR_5 are shown dotted since their function is carried out by the rectifier groups in the two amplifiers and consequently they are not needed.

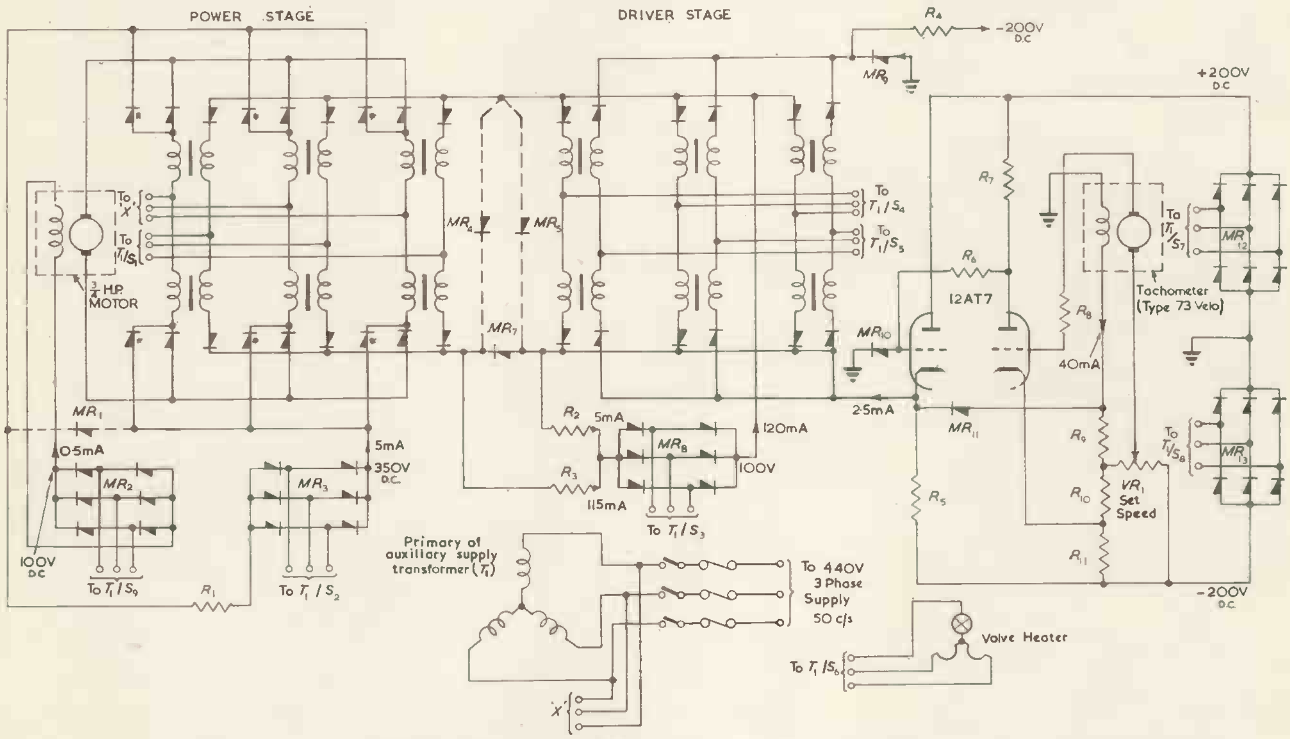


Fig. 14. Complete circuit diagram of magnetic motor control system

Anti-flux Depression and Coupling Circuits

From the considerations set out previously it is evident that if flux depression is to be avoided means must be found for permitting the proper flux changes to occur in the power stage, even when the motor back E.M.F. completely blocks the main rectifiers. In order to meet this requirement an independent low power load circuit is provided, which, feeding a non-active load is "open" during all gating periods. Since the only requirement of this auxiliary circuit is to pass the gating magnetizing current it is convenient to use a constant current source as a load element. This scheme is embodied into the circuit diagram of Fig. 14, where the rectifiers shown "starred" form the auxiliary load "routing" circuit and MR_1, R_1 and MR_3 constitute the constant current source and diode. The diode is shown dotted since in fact it need not be used at all because the routing rectifiers jointly perform the necessary function.

In addition to the scheme set out above, the interstage coupling is modified since it was found that under adverse supply conditions and when the load reactors were absorbing practically the entire applied voltage momentary blocking of the interstage coupling rectifier by the power stage was

Complete Motor Control System

As shown in the overall circuit diagram a type 73 Velodyne machine is used as a tachometer generator with its field supplied with current from the 200V D.C. line. A potentiometer connected in series with the field winding is used as a speed set control. Since the tacho output is a reasonably linear function of both speed and field excitation the arrangement is inherently self-compensating. The system is set up at the present time with a loop gain of 38db and is perfectly stable and fast in action. A detailed investigation into the dynamics of the whole system has not been attempted up to the present time but superficial tests indicate that the overall response is closely similar, if not identical to, an earlier motor control system employing a thyratron drive.

Conclusions

The work carried out to date in regard to multi-phase high-speed magnetic amplifiers must be regarded as being of an exploratory nature. Experience with the 1kW design, however, suggests that the special features of these amplifiers make them attractive propositions in some circumstances. At the present time a design study is under way, aimed at producing a 10kW amplifier using cheaper core materials: it is hoped that this may replace a high power thyratron set in an industrial type motor drive. This unit would be entirely magnetic since the tachometer envisaged for use would supply sufficient high frequency power to operate the first stage of the system. At this juncture it may be stressed that the similarity between the core units described earlier and the

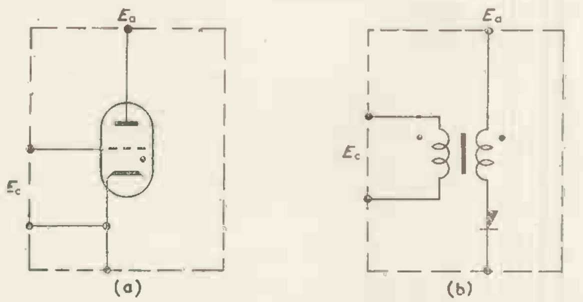


Fig. 15(a). Thyratron (b). Magnetic core unit.

thyratron valve is very striking since both are "on-off" devices and are substantially voltage controlled. The idea of ignition or firing angle, well known in thyratron practice is very readily applicable to the core units, and one may argue the equivalence of Fig. 15(a) and 15(b) with a certain amount of justification. Comparisons between magnetic amplifiers and electronic schemes have been made too frequently in the past to justify reiteration here, but at least it may be said that these new magnetic elements and ideas may well represent a challenge to a branch of engineering hitherto the exclusive field of thyratrons.

Acknowledgments

The Author wishes to express his appreciation to the de Havilland Propeller Company Ltd., for permission to publish this article and for supplying the photographic illustration.

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A Junction Transistor Amplifier for Noise Cancelling Telephones*

A noise cancelling telephone is a device to enable a telephone conversation to take place under conditions of high ambient noise such as in an aircraft engine test department or hangar. The actual microphone is so arranged that, while it is sensitive to a nearby voice it cancels out the noise which enters both from the front and the back of the device. However, such microphones are inherently insensitive and require an audio frequency amplifier to raise their output to the level of a standard telephone output.

In addition, the input to the receiver requires amplification so that it can be heard above the high local noise conditions under which these telephones are intended to operate.

The ideal amplifier for such an application should be of very small size, use little power, and be capable of operation as soon as it is switched on.

The transistor amplifier will provide these conditions and the power consumption is so small that it may even be possible to operate the entire amplifier from the microphone energizing current normally supplied for the standard carbon microphone.

A possible circuit to give the power gain required is shown in Fig. 1.

This is a basic circuit that will work well, but is not necessarily the ultimate in design since, for example, it does not include D.C. working point stabilization in the output circuit.

Again to save space and weight the input and output are not critically matched, in fact a considerable amount of gain is lost, but in return the circuit is less dependent on transistor variations with time and from one to another.

The entire circuit shown above can easily be put inside half a matchbox and will fit behind the microphone in a standard telephone handset.

Thus thanks to the small size and power consumption of a junction transistor, all the advantages of a noise cancelling microphone can be obtained without adding to the size or external appearance of the telephone.

D.C. Stabilization of Junction Transistors

Unlike valves with their high H.T. voltage, the transistor circuit which will work with 3V H.T. has to be carefully designed to ensure that the collector current through the collector load resistance does not drop so much voltage that none is left across the transistor itself.

If for the first stage of an amplifier (grounded emitter) a collector current of 0·5mA is chosen, then the increase of the back leakage current with temperature will be such that when 35° or 45° is reached it will be comparable with the original

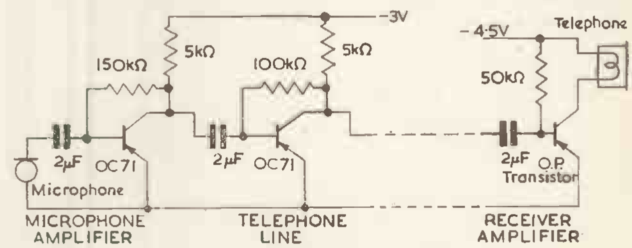

Fig. 1. Basic circuit for noise cancelling amplifier

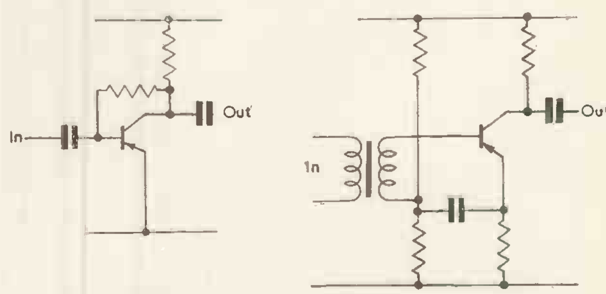

Fig. 2. Basic circuits for D.C. stabilization

collector current. This additional current through the collector load resistor will result in a reduction of voltage at the collector itself to almost zero and serious A.F. distortion will result.

Such a trouble can be overcome by providing the base current bias from a source that decreases as the back leakage current increases. This is, of course, D.C. negative feedback.

Two possible basic circuits are shown in Fig. 2.

* *Communication from Mullard Ltd.*

The Design of Triodes for U.H.F. Medium-Level Power Amplifiers

By W. E. Rowlands*, B.Sc., Ph.D., A.Inst.P.

A study is made of some of the factors involved in the design of U.H.F. *triode output amplifier valves. A brief survey is made of considerations regarding the oxide-coated cathode and the design of the grid. An expression is derived for radiant heat dissipation in the grid which allows an estimate to be made of the effects of the heat reflexion coefficients of the electrodes.*

Effects of evaporation from the cathode are investigated and relations between evaporation rate and grid growth are deduced. A procedure for estimating the effect of grid growth on valve constants and efficiency is indicated. Factors likely to cause an end of life of the valve are studied and a method is given of estimating the life as determined by grid growth.

The conflicting requirements of a valve of this type are outlined and it is shown that there is a value of cathode-grid spacing which gives an optimum power output at a given frequency.

An assessment of the possibilities at 2 000Mc/s shows that triodes are capable of a useful performance.

U.H.F. trunk radio systems require output amplifier valves of medium power level. Such valves should operate in broad-band systems, have a long life and reasonably high gain. For a valve working at 2 000Mc/s, a power output of 10W, a gain of about 10db, bandwidth of ±10Mc/s to 1db points, and a life of 10 000h are the sort of requirements in mind.

Triodes have some advantages for this application, in particular they make no great demands in the way of complex supply units and associated equipment such as magnets. On the other hand, their efficiency is limited by electron transit time effects. This means that very small inter-electrode spacings and fine-wire grids are necessary which make the valves difficult to manufacture. Furthermore, triodes tend to have rather a limited bandwidth.

A planar electrode system is usually chosen for a close-spaced triode meant to operate at about 2 000Mc/s. Compared with a cylindrical, or similar assembly, a planar grid is easier to construct and to maintain close to the cathode. Again, circular (disk) electrodes are normally used since these give rise to smaller stray capacitances. In such valves, the interelectrode spacings are much smaller than the lateral dimensions.

A relatively high electron emission density is required from the cathode and indirectly-heated oxide-coated cathodes have found almost universal application.

An attempt is made here to study some of the factors involved in the design of triodes for U.H.F. output amplifiers. No attempt is made to assess the relative merits of triodes and other valve types in this field.

The Cathode

An oxide-coating containing enough material to give a long life and applied by the normal spraying procedure has a thickness of the same order as the cathode-grid spacing. As the coating evaporates during life this spacing increases and the valve performance deteriorates. Thus, it is desirable that the coating be introduced as a thin, dense layer. One satisfactory method of doing this is to sinter fine tungsten powder on to the cathode base metal, spray the coating on this and crush it with a pressure of several ton/in^2.

Operating Temperature

The operating temperature depends on a number of factors. It is obviously essential that it should be high enough to maintain adequate emission. The actual value depends on the requirements of a particular valve and on the amount of positive ion bombardment, i.e. on the electrode potentials and the quality of the vacuum. On the other hand, it is desirable that the temperature be as low as possible to minimize the rate at which material is evaporated. Further, a low cathode temperature is of advantage in that it reduces radiant heat dissipation in the grid. With the type of cathode mentioned above, a temperature of about 720°C (oxide brightness) seems a suitable compromise.

When a desirable operating temperature has been determined, it is important that there should be little deviation from it. The electron emission and the evaporation rate are critically dependent on temperature so that small variations may be important.

The Evaporation Rate

The rate of evaporation of material from the coating is of prime importance in life considerations. Quite a number of factors are involved.

The evaporation rate probably depends on the cathode temperature. valve current density, base metal composition, the composition and treatment of the oxide-coating, the electrode potentials and the quality of the vacuum. In addition, at U.H.F. dielectric heating of the coating and bombardment by electrons returning to the cathode complicate matters.

Unfortunately there is little information on the dependence of evaporation on most of these factors. This makes it difficult to make estimates if the evaporation rate is involved and the designer is almost completely dependent on information gleaned from previous valves.

For a mixed (Ba, Sr oxides) coating on an "0" Nickel base, Woods[1] has measured the evaporation rate to be $2{\cdot}3 \times 10^{-6}$ g/cm^2/100h for a cathode temperature of 765°C (oxide brightness). He found that the rate was increased by 30 per cent when a current density of 20mA/cm^2 was drawn. The evaporated material was primarily metallic and there is some evidence[2] that it is mostly barium.

The Grid

The design and manufacture of grids for close spaced triodes has been described[3,4]. Molybdenum and tungsten wires are found to be the most suitable. For wire diameter less than 0·03mm tungsten is generally used. The wires are

* *G.E.C. Research Laboratories, Wembley.*

arranged to be in tension and it is found that grids of straight, parallel wires are preferable since the tensioning can be done in a reliable and reproducible manner.

The condition of the longest wire in the grid is the most critical, and attention will be directed on this.

If the heat dissipated from the grid by radiation is neglected and it is assumed that the dissipation is uniform along the wires then the temperature distribution along the wire is parabolic[4]. The maximum temperature at the centre of the wire is given by:

$$\theta_m - \theta_o = Ql^2/8k\pi r^2 \quad (1)$$

If the wire is stretched within its elastic limit, the strain which is just removed by a certain dissipation is:

$$\gamma = \alpha Ql^2/12k\pi r^2 \quad (2)$$

A nomogram representing these expressions for tungsten wires is shown in Fig. 1. This covers the range of parameters likely to be of interest and in its construction values of $k = 0{\cdot}4$cal/sec/cm/°C and $\alpha = 4{\cdot}4 \times 10^{-6}$/°C were used.

List of Symbols

C = valve output capacitance
d_p, d_N = grid wire diameter parallel and normal to cathode plane respectively
f = frequency in Mc/s
Δf = bandwidth in Mc/s
g_m = mutual conductance
I = intensity of heat radiation
I_a = anode current
I_Z = intensity of evaporation
i_a' = anode current swing
J = current density in A/cm²
J_o = energy conversion factor = 4·18W/cal/sec
k = thermal conductivity of grid wire in cal/sec/cm/°C
K_1, K_2, K_3 = constants
L_1 = valve life as determined by loss of emission
L_2 = valve life as determined by increase in μ
l = grid aperture diameter in cm
M_B = mass of barium in cathode in g/cm² of cathode area
P = grid dissipation due to electron bombardment in W/cm² of grid plane
P_o = power output
p = grid wire pitch in cm
Q = heat dissipation in grid wire in cal/sec/cm length of wire
Q_P, Q_R = contributions to Q due to electron bombardment and radiant heat respectively
R = load resistance
r = grid wire radius in cm
r_1 = grid wire radius corresponding to $\mu = \mu_1$, in cm
r_c, r_g, r_a = total heat reflexion coefficients of cathode, grid and anode
S_1, S_2, S_3 = cathode-grid, grid-anode and cathode-anode spacings in cm
$S_{1(opt)}$ = value of S_1 for optimum power output
V_g, V_a = grid and anode voltages
v_a' = anode voltage swing
W = radiation from cathode in W/cm²
Z = evaporation rate in g/cm²/100h
α = thermal expansion coefficient of grid wires
γ = strain in grid wire
η = efficiency
η_o = efficiency in absence of transit time limitations
θ_m = maximum wire temperature
θ_o = temperature of grid frame
μ = valve amplification factor
μ_1 = value of μ when the rated anode voltage only just draws the required anode current at zero grid voltage
ρ_g = density of material deposited on grid
ρ_e = "effective" density of evaporated material
S_1', d_p', etc. = rate of change of S_1, d_p, etc.

Some other symbols are used in the process of evaluating results, but their meaning is indicated where they occur.

Estimation of the Dissipation

The heat dissipated in the grid is derived from two main sources:

(a) radiation from the cathode and
(b) electron bombardment.

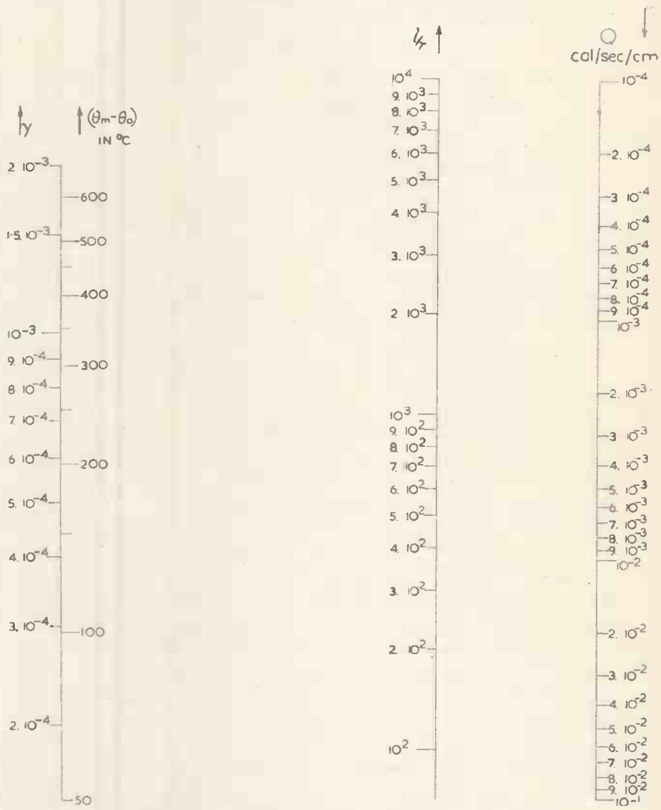

Fig. 1. Nomogram for tungsten grid wires

For a particular valve the contribution due to the latter can be fairly easily estimated and is given by

$$Q_P = Pp/J_o \quad (3)$$

Pohl[4] states that dissipation due to radiation from the cathode is usually less than 10 per cent of the total dissipation and that its contribution can be adequately represented by $2Wr/J_o$ (using our symbolism). For the type of valve under discussion, however, the radiant heat contribution is likely to be much higher than 10 per cent and a more detailed consideration is of value.

Radiant Heat Dissipation in the Grid

Consider first the radiation intercepted by a straight wire held parallel to a flat infinite radiator. The radiator 1, Fig. 2, radiates energy $W = \pi I$/unit area. The radiation from a small element of radiator area which is intercepted by a small length of wire, δl in Fig. 2, is

$$\delta q = I2r\delta l \cos\beta \cos\psi \; \delta x \delta y/z^2$$

Putting $\cos\psi = S_1/z$, $\cos\beta = \sqrt{(x^2+S_1^2)}/z$, $z^2 = x^2+y^2+S_1^2$

and integrating for x and y from $-\infty$ to $+\infty$ then the radiation intercepted by a small length of wire is $\pi W r \delta l$.

Hence the radiation intercepted per unit length of grid wire is $\pi W r$.

Consider now the state of affairs in an actual valve. Radiation leaves the cathode and some is intercepted by the grid. Some of the remainder is reflected at the anode and some of this is again intercepted by the grid. Reflexion then occurs at the cathode, so that radiation is reflected to and fro between cathode and anode. Some is intercepted by the grid in each passage across the valve, some is lost at each reflexion and some is lost at the edge of the electrode space. Furthermore, of the radiation intercepted by the grid some is absorbed and the remainder is reflected to be intercepted by neighbouring wires or to rejoin the multi-reflexion process. A complete analysis of the situation would be extremely complex, but useful results can be obtained on making some approximations.

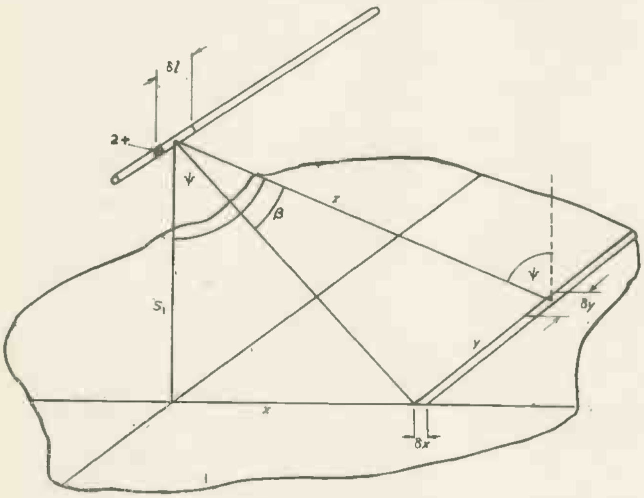

Fig. 2. Radiant heat interception

Neglect radiation originating at the grid and anode. Ignore the loss of radiation at the edge of the inter-electrode space and to counteract this also ignore the history of the radiation reflected at the grid. Further, suppose that the cathode can be treated as a simple reflecting surface.

The length of grid wire per unit area of grid plane is $1/p$. Radiant energy W/unit area leaves the cathode. An amount $\pi W r/p$ is intercepted by the grid and the remainder, $W(1-\pi r/p)$ goes on to the anode. $W r_a(1-\pi r/p)$ is reflected from the anode and of this $W r_a(1-\pi r/p)\pi r/p$ is intercepted by the grid. The remainder, $W r_a(1-\pi r/p)^2$, goes on to the cathode and so on.

Then the total radiation intercepted per unit area of grid plane is

$$\frac{\pi W r}{p}[1+r_c r_a(1-\pi r/p)^2+\ldots]+\frac{\pi W r r_a}{p}[1+r_c r_a(1-\pi r/p)^2+\ldots]$$

which reduces to $\dfrac{\pi W r(1+r_a)}{p[1-r_c r_a(1-\pi r/p)^2]}$

since $r_c r_a(1-\pi r/p)^2 < 1$

Hence the dissipation per unit length of grid wire is:

$$Q_R \simeq \frac{\pi W r\,(1-r_g)(1+r_a)}{J_o\,[1-r_c r_a(1-\pi r/p)^2]} \quad \ldots\ldots\ldots\ldots \quad (4)$$

Taking reasonable values $r_c = r_g = r_a = 0{\cdot}7$ and $r/p = 0{\cdot}1$ then $Q_R \simeq 1{\cdot}8\ Wr/J_o$. This agrees quite well with the estimate of $2Wr/J_o$ obtained following Pohl's procedure, and this approach appears adequate for most purposes.

On the other hand, expression (4) enables the effect of variation in r_c, r_g and r_a to be estimated. As an example consider a change in anode reflexion coefficient, r_a. With $r_a = 0{\cdot}9$ and other values as before $Q_R = 2{\cdot}6\ Wr/J_o$, but if $r_a = 0{\cdot}15$ then $Q_R = 1{\cdot}1\ Wr/J_o$—a significant difference.

LIMITATION ON GRID DESIGN

An increase in grid temperature has two effects. First, primary electron emission from the grid increases and, secondly, thermal expansion of the grid wires tends to occur.

It is common practice to cover the grid wires with a layer of some other metal, such as gold, to reduce the grid emission level[5]. These valves are normally used with a bias resistor and the primary grid current flows through this. If the primary grid current becomes an appreciable part of the total biasing current, the operation of the valve may become unstable. An accumulative increase in valve current leading to eventual breakdown may occur. This state of affairs is not unlikely when the bias is produced by the grid current. However, if the total valve current is used for biasing (by using a cathode resistor) this tendency to breakdown is very much reduced. If a cathode resistor is used it is unlikely that the effects of primary grid emission will become a limiting factor in grid design before other considerations have set a limit.

The tendency of the grid wires to expand is overcome by having the wires in tension. From equations (1) and (2), $\theta_m - \theta_o = 3\gamma/2\alpha$ and taking $\alpha = 4{\cdot}4 \times 10^{-6}/°C$ and $\gamma = 3 \times 10^{-3}$ then the tension is relieved when $\theta_m - \theta_o \simeq 1\,000°C$. A value of strain of 3×10^{-3} is well within the capabilities of tungsten wire. However, recrystallization of the tungsten would lead to strain relief at this temperature. Recrystallization has been observed microscopically at 1 000°C and by X-ray methods at 800°C[6]. Further, there is no information on the creep of tungsten at elevated temperatures over long periods.

It would appear then that the maximum grid temperature should not exceed about 600°C for safe operation over prolonged periods.

Bandwidth

Valves of this type are generally used in coaxial-line common-grid circuits. Tuned input and output circuits are used, but the bandwidth is almost completely determined by the output circuit.

Morton and Ryder[7] show that there is a value of electron transit angle in the grid-anode region which optimizes the gain-bandwidth product. This value gives a high μ which is undesirable from life considerations. Normally grid-anode clearances giving a transit angle considerably less than the optimum are used.

It has also been shown[8] that best bandwidth results are realized when the active and stray capacitances are approximately equal.

Now the bandwidth $\Delta f \propto 1/CR$ .... (5)

and for a given class of operation $R \propto e_a'/i_a'$ .... (6)

For optimum power output the bandwidth of a triode in a circuit of this type is usually much less than is required in trunk radio work. A value of ±5Mc/s to 1db down points was found for a valve working at 2 000Mc/s. The required bandwidth can be realized by reducing R, but in so doing there is a loss of power output.

Effects of Cathode Evaporation on Valve Geometry

Some of the material evaporated from the cathode will be deposited on the grid and anode which are relatively cool. Since the grid-anode clearance is usually several times the cathode-grid clearance and the grid pitch, the change in valve geometry due to deposition on the anode can be neglected.

If the grid temperature is not excessive there will not be any appreciable migration or re-evaporation of the deposited material.

Grid Growth

The cross-section of the wire is modified as shown in Fig. 3.

Consider first the change in wire diameter parallel to the cathode plane. To estimate this, consider a plane perpendicular to the cathode plane and tangential to the wire, see Fig. 4. Suppose the cathode plane to be infinite and let the evaporation rate be uniform.

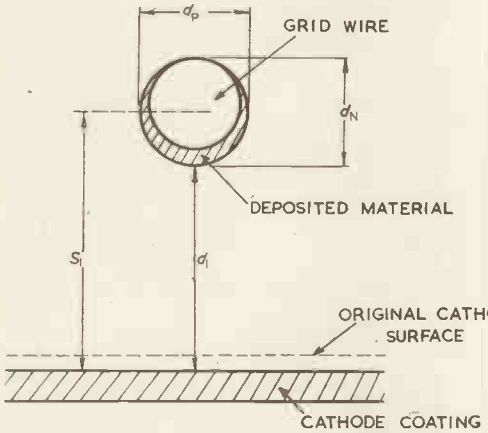

Fig. 3. Grid growth due to deposition

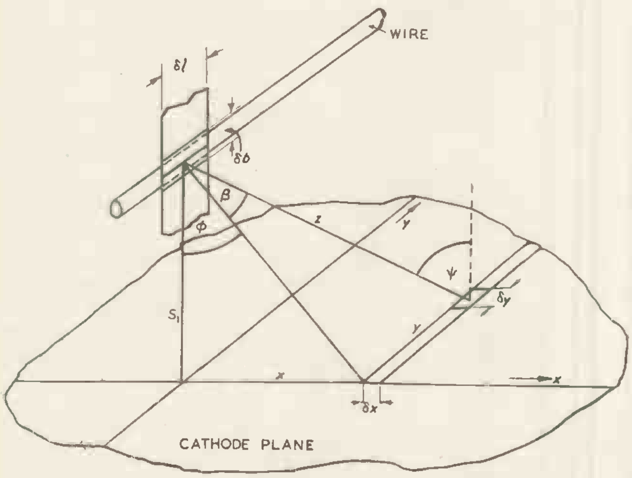

Fig. 4. Change in wire diameter parallel to cathode plane

In a high vacuum valve the cosine distribution law and the inverse square law apply to the evaporated material, hence $Z = \pi I_Z$.

A rectangular strip, $\delta l \times \delta b$, on the tangential plane intercepts an amount of material, evaporated from an element of cathode area, given by

$$\delta F_P = I_Z \delta l \delta b \sin\phi \cos\beta \; \delta x \delta y \cos\psi / z^2$$

Putting $z^2 = x^2 + y^2 + S_1^2$, $\cos\psi = S_1/z$, $\cos\beta = \sqrt{(x^2 + S_1^2)}/z$,

$$\sin\phi = x/\sqrt{(x^2 + S_1^2)}$$

and integrating for x from 0 to ∞ and y from $-\infty$ to $+\infty$

$$F_P = \tfrac{1}{2} Z \delta l \delta b$$

Hence the rate of increase of wire diameter, parallel to the cathode plane, is:

$$d_P' = Z/\rho_g \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

Next, consider the change in wire diameter in a direction perpendicular to the cathode. Consider a plane parallel to the cathode and tangential to the wire on the side nearest the cathode (see Fig. 5). A strip $\delta l \times \delta b$ on this plane, as shown, intercepts material from an element of cathode area given by

$$\delta F_N = I_Z \delta l \delta b \cos^2\psi \; u \delta u \delta\phi / z^2$$

Putting $u = d_1 \tan\psi$, $\delta u = d_1 \sec^2\psi \delta\psi$, $z = d_1 \sec\psi$ and integrating for ϕ from 0 to 2π and ψ from 0 to $\pi/2$ then

$$F_N = Z \delta l \delta b$$

There is no deposition on the point of the wire furthest from the cathode. Hence the rate of increase of wire diameter, normal to the cathode plane, is also

$$d_N' = Z/\rho_g \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

Effects on μ, g_m and Electron Transit Time

The increase in wire diameter causes an increase in the μ of the valve. This effect can be estimated from equation (7) and a standard expression for μ or the chart given by Spangenberg[9].

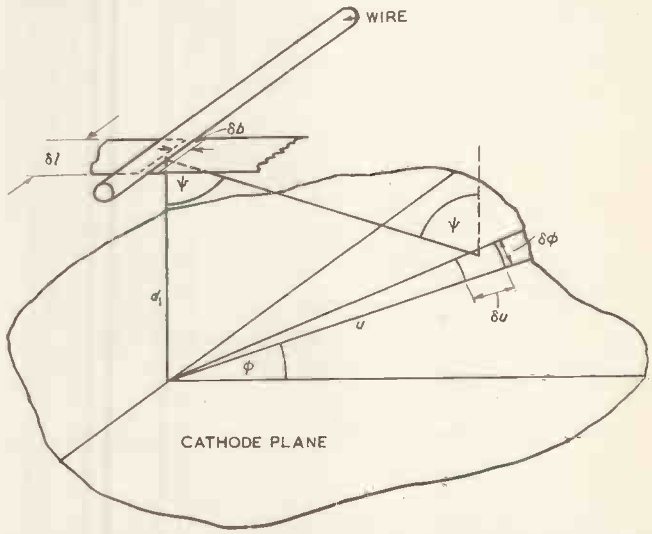

Fig. 5. Change in wire diameter in a direction perpendicular to the cathode

Growth of the grid tends to decrease the cathode-grid spacing S_1 at a rate of approximately $Z/2\rho_g$. Simultaneously the loss of material at the cathode increases S_1 at a rate of $Z/_{pe}$. Hence, there is a net rate of change of S_1 of

$$S_1' = Z(1/\rho_e - 1/2\rho_g) \quad \ldots\ldots\ldots\ldots \quad (9)$$

This change in S_1 affects the electron transit time in the cathode-grid region and may thus influence the efficiency of the valve. The change in efficiency may be estimated from Law's formula.

The change in S_1 and the change in μ cause the g_m to change. This can be estimated using a standard formula for g_m such as that given by Spangenberg[9].

Whether the loss of cathode material results in the coating shrinking by an amount corresponding to the density of the lost material or whether the coating is left with a more open structure is debatable. In any case, ρ_e can be assigned an "effective" value.

Life

Basically, the life of a valve of this type is governed by the evaporation from the cathode. This evaporation can cause the end of life in several ways.

Loss of Emission

The most obvious cause of end of life is loss of emission due to the complete loss of emissive material. In view of the evidence that there is a selective loss of barium the end of life is very near when all the barium is lost. As determined by this effect the life is

$$L_1 = 100 M_B / Z \quad \ldots\ldots\ldots\ldots \quad (10)$$

This life can be extended by using a replenisher type cathode[11,12] where a practically limitless amount of emissive material can be included. Unfortunately these cathodes operate at a higher temperature and this increases both the evaporation rate and the radiant heat dissipation in the grid.

Electrical Breakdown

Contamination of the grid and anode reduces the electrical gap insulation between the electrodes. This increases the tendency to electrical breakdown which can destroy the valve. To our knowledge, there is no information allowing this tendency to be put on a quantitative basis.

Current Limitation

The increase in μ, caused by grid growth, requires that the valve bias be reduced if the same valve current is to be maintained. No provision is usually made for positive bias. Thus, after the bias has been reduced to zero, to maintain the current, the valve output begins to drop. An "effective" end of life may be defined as occurring at this point of zero bias. After this there is a gradual deterioration in performance.

The current density can be expressed[13]

$$J = \frac{2{\cdot}335 \times 10^{-6}\,(V_g + V_a/\mu)^{3/2}}{S_1^{\,2}[1 + (S_3/S_1)^{4/3}/\mu]^{3/2}}$$

The effective end of life occurs when $V_g = 0$ and usually $(S_3/S_1)^{4/3}/\mu_1 \ll 1$ so that end of life occurs when μ is given by

$$\mu_1 = 1{\cdot}76 \times 10^{-4} V_a / S_1^{4/3} J^{2/3} \quad \ldots\ldots\ldots (11)$$

If r_1 is the grid wire radius corresponding to μ_1, the effective life is

$$L_2 = 100(r_1 - r)2\rho_g/Z \quad \ldots\ldots\ldots (12)$$

Interelectrode Short-Circuits

Grid growth causes increased heat dissipation in the grid. The radiant heat interception increases due to increased wire radius. Also, the grid must become progressively more positive because of increase in μ and this causes more electron bombardment.

Hence, a point may be reached when the maximum grid temperature exceeds the safe limit. Beyond this point a catastrophic end of life may occur due to relief of tension in the grid wires allowing an interelectrode short-circuit to develop.

This tendency should be allowed for in the design of the grid.

Valve Design Considerations

Conflicting Demands

The requirements on a valve of this type are conflicting in several respects. Consider how variation of some of the valve parameters will affect matters.

Cathode-Grid Spacing

A decrease in this would increase the efficiency and the gain. It would also increase μ_1 (equation (11)) and hence the life L_2.

Grid Pitch and Wire Size

A reduction in cathode-grid spacing implies that the grid pitch must also be reduced to maintain adequate grid control. This means that either the life L_2 would be decreased due to an increase in initial μ value, or the grid-anode spacing has to be reduced causing an increase in C and a corresponding reduction in bandwidth, or the grid-wire diameter has to be reduced implying the use of a smaller grid aperture and thus tending to reduce the power output.

Grid-Anode Spacing

There is an optimum value of this for the best gain-bandwidth product. In practice grid-anode spacings considerably smaller than the optimum are used. Increasing the spacing would improve the gain-bandwidth product, but would decrease the life due to an increase in initial μ value.

Current Density

An increase in current density would increase the efficiency and the gain. These advantages could be used to increase the power output or to reduce the electrode area. The latter has the associated advantages of reducing C and allowing a finer wire grid to be used. On the other hand, there would be a reduction in life due to increased evaporation rate from the cathode and to a lower value of μ_1.

Anode Voltage

An increase in this would increase the power output, and also the life as determined by μ_1. However, there is a tendency to shorten the life due to cathode bombardment by higher energy gas ions.

It follows from this survey that the valve design must be a compromise between several conflicting demands. The information given in the preceding sections should be of aid in reaching such a compromise.

Expression for Efficiency

Law[14] has shown that the efficiency of a triode can be represented by

$$\eta = \eta_0(1 - K_1 f S_1/\sqrt{V_a}) \quad \ldots\ldots\ldots (13)$$

and if f is in Mc/s, $K_1 = 0{\cdot}48$ for an amplifier. (This differs from the value of 1·2 given by Law since S_1 is measured in centimetres here as compared with inches in the original formula.)

Results on an experimental valve indicate that this expression gives estimates of efficiency appreciably too high for our purpose. This valve had values of $S_1 = 0{\cdot}1$mm, $V_A = 500$ volts and drive power of 2W/cm² of electrode area.

As expected, the efficiency followed a relation $\eta = \eta_0(1 - K_3 f)$. It fell from 20 per cent at 1 700Mc/s to 10 per cent at 2 300Mc/s. This gives a value of $\eta_0 = 50$ per cent and K_1 is estimated as 0·75.

Optimum Cathode-Grid Clearance

When the cathode-grid clearance is reduced, the grid pitch must also be reduced. If adequate grid control is to be maintained, $p \leqslant 1{\cdot}5\, S_1$ [15]. The grid wire radius must also be reduced if the screening factor is not to increase. Thus as the cathode-grid clearance is reduced the electrode area must also be reduced and this tends to reduce the power output from the valve.

On the other hand, a reduction in grid-cathode spacing increases the efficiency according to a relation $\eta = \eta_0(1 - K_2 S_1)$ where

$$K_2 = K_1 f/\sqrt{V_a} \quad \ldots\ldots\ldots (14)$$

Consider a valve where the anode voltage, current density, grid-anode spacing and drive power per unit area of grid plane are maintained constant. Let $p = aS_1$ and $r = bS_1$, where a and b are constants. Suppose that it is arranged that the grid operate at its maximum safe value. From equation (4), $Q_R \propto S_1$; and from equation (3), $Q_P \propto S_1$. Hence $Q \propto S_1$. To maintain the maximum grid temperature constant, $Ql^2/r^2 =$ constant, i.e. $l^2 \propto S_1$.

Now the electrode area $\propto l^2$, so that the valve current $I_A \propto S_1$.

The power output $P_o = \eta V_a I_a$, hence $P_o \propto S_1(1 - K_2 S_1)$.

This shows that there is a maximum power output for a value of S_1 given by

$$S_{1(opt)} = 1/2K_2 \quad \ldots\ldots\ldots (15)$$

Furthermore $\Delta f \propto 1/CR$. If C is taken as the active grid-anode capacitance then $C \propto l^2$ approximately. Furthermore, since the μ is kept constant v_a' is approximately constant and $i_a' \propto l^2$. Roughly then, the bandwidth is independent of S_1.

Thus $S_{1(opt)}$ gives the value of cathode-grid spacing which gives an optimum power output-bandwidth product.

POSSIBILITIES AT 2 000Mc/s

An attempt is now made to make an assessment of the possibilities at 2 000Mc/s. Some of the estimates are very tentative due to the lack of information of some of the relevant parameters.

The following are thought to be reasonable estimates of some of the factors $V_a = 500$V, $J = 200$mA/cm², $W = 5$W/cm², $P = 0{\cdot}7$W/cm², $Z = 2 \times 10^{-5}$g/cm²/100h, $\rho_g = 1$g/cm³, and $M_B = 3 \times 10^{-3}$g/cm². The following relationships are typical: $p = 1{\cdot}5S_1$, $r = 0{\cdot}1p$, and $S_2 = 5S_1$.

K_1 is estimated as 0·75 from equation (14), $K_2 = 67$ and, from equation (15), $S_{1(opt)} = 0{\cdot}07$mm.

Then $p = 0{\cdot}10$mm and $r = 0{\cdot}01$mm.

From equation (13), with $S_1 = 0{\cdot}07$mm the efficiency at 2 000Mc/s is estimated as 25 per cent.

It would be a nice compromise to arrange that the grid temperature should reach its maximum safe value at about the same time that the effective end of life due to current limitation occurs. This latter point occurs when $\mu_1 = 190$, see equation (11), and from Spangenberg's chart[9] the grid wire radius is then $r_1 = 0{\cdot}02$mm.

Suppose at this stage that about one-third of the drive power is dissipated in the grid, then from equation (3) $Q_P \simeq 2 \times 10^{-3}$cal/sec/cm. Again $Q_R \simeq 2Wr/J_o$ so that $Q_R \simeq 5 \times 10^{-3}$. Hence $Q \simeq 7 \times 10^{-3}$cal/sec/cm.

It is required that $\theta_m = 600$°C so that $\theta_m - \theta_o = 500$°C. From the nomogram, Fig. 1, this is so when $l/r = 8{\cdot}5 \times 10^2$ giving $l = 8{\cdot}5 \times 10^{-1}$cm. This gives the electrode area $\simeq 0{\cdot}57$cm² and valve current $\simeq$ 115mA.

Hence anode input power $\simeq$ 57 watts and power output $\simeq$ 14 watts.

The life as determined by complete loss of emission is approximately 15 000h, from equation (10). Current limitation would set an "effective" end of life at about 10 000h, see equation (12).

Very approximately then a triode operating at 2 000Mc/s with an efficiency of 25 per cent, a power output of 14 watts, a gain of 11db and a life of 10 000h seems possible. The bandwidth of such a valve would almost certainly be inadequate and the power output and gain would have to be reduced to realize the desired bandwidth.

Conclusion

Triodes have a useful performance in trunk radio applications at 2 000Mc/s. An efficiency of 25 per cent, a power output of 9-10 watts at a bandwidth of ±10Mc/s, a gain of about 10db and a life of 10 000h, are about the best figures obtainable on the above argument.

A limited improvement on the above figures could probably be realized. The reflexion coefficients of the electrodes could be adjusted to reduce radiant heat grid dissipation. Also the limit on electrode area, set by grid temperature, could be extended by changing from a circular grid aperture. This would tend to increase the stray capacitances so that the advantage is somewhat doubtful.

A cathode capable of a higher current density at a reduced evaporation rate and reduced heat radiation, i.e. increased J/ZW factor, might enable a significant improvement to be made.

Acknowledgment

In conclusion, the author desires to tender his acknowledgment to the M.O. Valve Co. Ltd on whose behalf the work described in this article was carried out.

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A 15MeV Linear Accelerator

For Medical Use

A 15 million electron-volt linear accelerator designed and constructed by the Mullard Research Laboratories has recently been installed at St. Bartholomew's Hospital College. The machine is of the travelling-wave type, that is, it accelerates electrons continuously by subjecting them to a more or less constant electric field as they travel down the whole length of the machine. It is similar in many respects to the linear accelerator installed at A.E.R.E. Harwell in 1952,[1] but gives a somewhat higher energy and has its controls designed specifically for medical use.

Background

Travelling-wave linear accelerators produce electrons of high energy, which have a number of applications in pure and applied physics, industry and medicine. When these electrons are stopped by a target constructed of a heavy metal such as platinum, they produce hard X-rays and these find application in radiotherapy.

In conventional X-ray tubes the electrons are made to travel down an evacuated tube under the influence of a high voltage applied to the tube. The shortest wavelength (and hence their penetrating power) of the X-rays produced depends on the accelerating voltage. Thus electrons which have been accelerated by 1 million volts are said to have an energy of 1 million electron-volts (1MeV) and the X-radiation produced is described as 1MeV X-radiation.

In practice X-ray machines operating up to 250kV can be made fairly easily, and many such units are in use in hospitals. The problem of providing efficient insulation for higher voltages in X-ray tubes is, however, a difficult one and for operating voltages above about 2 million volts the size and cost of the equipment become prohibitive. Since X-rays of higher energies are valuable in the treatment of deep-seated tumours, there is a need for machines which produce high-energy electrons without the direct application of very high voltages.

One broad group of accelerators uses a magnetic field to constrain particles to move in a circular or spiral orbit. Machines in this class are the betatron and synchrotron for electrons, and the synchroton and cyclotron for heavy particles. Such machines usually give only a small beam current so that the X-ray output is not very great.

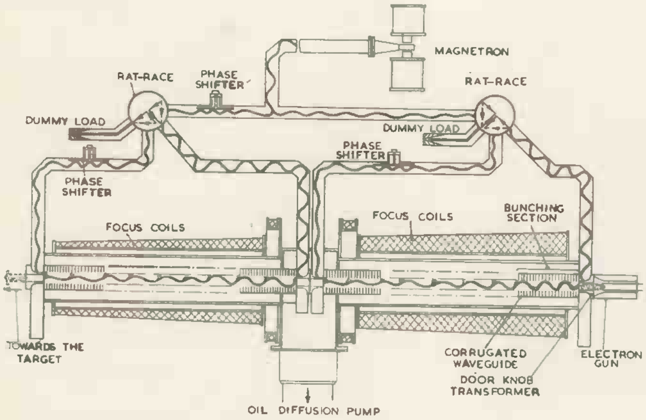
Methods of accelerating particles to high energies by means of comparatively low voltages were suggested by Wideroe in 1928 and by Professor E. T. S. Walton in 1929. These systems are known as standing-wave linear accelerators.

Radio-frequency voltages are applied at various points along a long evacuated tube. Particles come under the influence of the electric field produced by these voltages at times when the field is of such a polarity as to accelerate them.

Early linear accelerators were capable of accelerating heavy ions, but the acceleration of electrons to velocities approaching that of light was not feasible until high powers at microwave frequencies became available. This was provided by the cavity magnetron whose development was stimulated by wartime radar requirements.

This machine, like most present-day machines, is of the travelling-wave type. This means that instead of the particles receiving a large number of small accelerations, they are accelerated continuously in a more or less constant electric field down the whole length of the machine. The electric field is produced by passing the R.F. power down a type of circular waveguide known as a "corrugated waveguide". The electrons pass down the centre of this waveguide whose dimensions are arranged so that the electric field pattern travels at the same velocity as the electrons.

Like the 15MeV Harwell machine, the accelerator is divided into two halves, each taking half of the available magnetron power. The two halves receive the electron beam in series. Each half consists of 3 metres of corrugated guide.



Arrangement of the linear accelerator.

The energy provided by the first half of the machine is about 6·5MeV.

Since the accelerating field is still appreciable at the end of each 3-metre section, there will be considerable R.F. power still available at these points. This is used profitably by means of a recirculation system in each half, so that the input power to each section of corrugated waveguide is the sum of half the magnetron power and the unused power remaining after 3 metres.[2] The recirculation power is added to the generated power in a special form of "rat-race" waveguide ring circuit.

Electrons are injected into the first section of corrugated waveguide from an electron gun. This has pulses of about 50kV applied to it so that electrons with a velocity of 0·4c will pass into the waveguide where they form into close bunches. These bunches will travel in a part of the moving electric field such that electrons tending to lead the bunch will receive less acceleration while electrons tending to lag behind will receive more acceleration.

The R.F. field in the corrugated waveguide has a component which tends to produce defocusing of the electron beam. This defocusing is prevented by the application of a magnetic field by means of focus coils surrounding the vacuum envelope which contains the corrugated waveguide.

The closer the spacing of the corrugations the greater will be the power losses in the copper surfaces. On the other hand too few corrugations reduce the accelerating efficiency due to axial variations of the electric field.

At the input end of the corrugated waveguide the electron velocity is 0·4c, so that a guide wavelength of 4cm, which corresponds to a phase velocity of 0·4c, will keep wave and electrons in step. The corrugation pitch is here made 1cm, corresponding to 4 irises per wavelength. At the point in the waveguide where the electron velocity and therefore the guide wavelength, have doubled, the pitch is changed to 2cm between irises. This means that the attenuation of the corrugated waveguide due to losses in the walls is much less than if a uniform spacing had been adopted for the whole length. A quarter-wave transformer matches the wave impedances of the two pitches of waveguide.

As described by Bareford and Kelliher, the corrugated waveguide is made up of a large number of cells machined from copper pressings and soldered together using the copper-silver entectic. These cells were mostly machined to dimensions intended to give the R.F. wave a velocity equal to the predicted electron beam velocity. At four places, correcting sections were inserted to correct for any errors in wave velocity due to machining tolerances.

The 15MeV Harwell accelerator, like other machines constructed about the same time, gave electron energies somewhat less than the design figures indicated[1,3]. In the 15MeV St. Bartholomew's macnine the corrugated waveguide in the second 3-metre section was redesigned to give a somewhat greater electric field in the axial direction. This resulted in the final electron energy being up to the original design figure.

The accelerator itself is divided into six units, each containing a one-metre length of corrugated waveguide, mounted on trolleys running on a system of rails. The electron gun section is also trolley-mounted. The miscellaneous supplies are divided into conveniently small units mounted in racks and

A general view of the linear accelerator.

connected by flexible cables fitted with plugs and sockets. Ease of servicing was constantly borne in mind during the design of the whole machine.

Use of the Machine

The machine is installed at the Physics Department of St. Bartholomew's Hospital College, where it will be operated under the direction of the Professor of Physics. The machine is being acquired by the hospital for the treatment of patients, but during the first few years it will be used entirely for research work in fundamental investigations into action of radiations on living tissue and into the best methods of treatment of malignant tumours.

St. Bartholomew's Hospital was the first to enter the field of high-voltage therapy by installing a 1 million volt X-ray machine in 1937. This machine is still in operation in the radio-therapy department of the hospital. The new linear accelerator is the largest and most powerful machine of its kind to be used for medical purposes anywhere in the world.

In addition to its use as a source of high-energy X-rays, the St. Bartholomew's accelerator is so designed as to make it possible to use the beam of 15MeV electrons directly for treatment. The therapeutic value of high-energy electrons has only recently been demonstrated, and has many advantages over high-energy X-rays in the treatment of certain types of tumour. A third use of the accelerator will be as a source of fast neutrons. The effect of these on living tissue has yet to be investigated. The neutrons are produced when electrons are made to hit a uranium target.

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A Torquemeter for Testing Gas Turbine Components

At 11 000rev/min 500kW

(Part 2)

By J. F. Field* and D. H. Towns*

A BLOCK diagram of the whole equipment is shown in Fig. 9 and a circuit diagram of the connexions of the photo-electric cells, differential amplifiers and other electronic equipment is shown in Fig. 10. These diagrams indicate that push-pull operation was adopted not only at the end from which two anti-phase voltages are required, but also at the other end. The two push-pull generators at end 1 of the torsion rod have been designated HF1A and HF1B and those at end 2 have been designated HF2A and FH2B. There are, in addition, the two generators which resolve the ambiguity of the quadrant being measured, i.e. they form the "hour hand of the clock". These generators have been designated LF1 and LF2 for ends 1 and 2 respectively.

Photo-multiplier cells were used for the generators as these have the advantage of giving a large output and therefore do not require an amplifier. American 931A's were used, the British equivalent being Mazda type 27M1. The voltage dividing network is connected to the high-tension supply through resistor R_1 and is decoupled through capacitor C_1. The anode of the photo-electric cell is connected to the high-tension supply by its load resistor R_{12}, and the output is taken from the anode through a blocking capacitor C_3.

The phasemeter and associated apparatus were located at some distance from the voltage generators, which necessitated the outputs of the generators being delivered over a run of some 80ft of screened cable. In order to eliminate phase shifts due to the cable, its capacitance to earth was kept low by using 75 ohm coaxial cable. Even then it was found that the capacitance caused a variation in waveform with frequency; and to overcome the trouble the source impedance was reduced to about 250 ohms by interposing cathode-followers V_{3a} and V_{3b} between the photocell output terminals and the cables.

The outputs from the cathode-followers are then fed to their respective differential amplifiers. Each differential amplifier consists of a twin-triode V_5, having anode loads R_{37} and R_{38} and a common cathode load R_{34}. Bias for both triodes is obtained from R_{33}, which is decoupled by C_{11}. The voltage of the grid of each triode is held at the correct D.C. value relative to its cathode through resistors VR_5 and R_{32} and resistors VR_6 and R_{35} respectively, and the grids are decoupled from R_{34} and from each other by C_9 and C_{10}. The outputs from the two anode circuits are applied via blocking capacitors C_{12} and C_{13} to voltage dividers consisting of resistors R_{39} and R_{41} and resistors R_{40} and VR_7. With the component values shown in Fig. 10 the gain with anti-phase inputs is 3·53 and with in-phase inputs is 0·07, a ratio of 50·5:1. While a valve of higher mutual conductance, together with suitably selected anode and cathode loads, would undoubtedly have given a ratio of the two gains of 200 or more (see appendix), the value achieved with the 6SN7 valve was considered satisfactory and had the advantage of keeping the valve types standard throughout.

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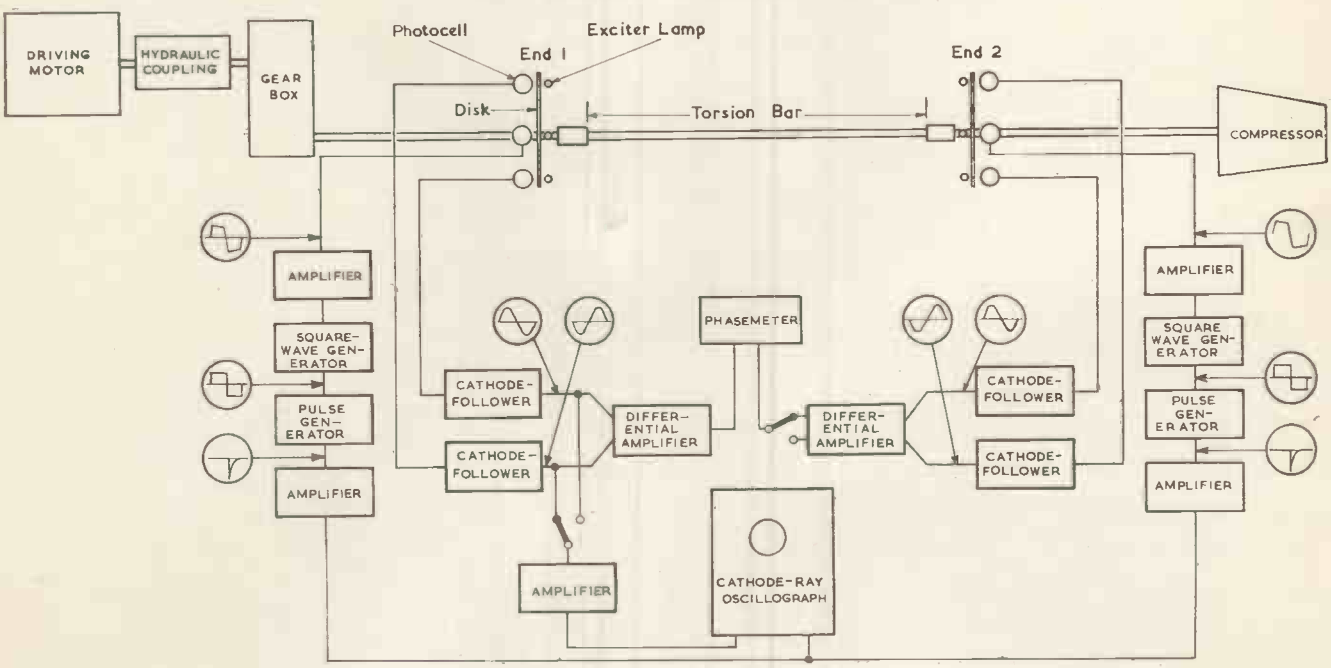
Fig. 9. Arrangement of the electronic equipment



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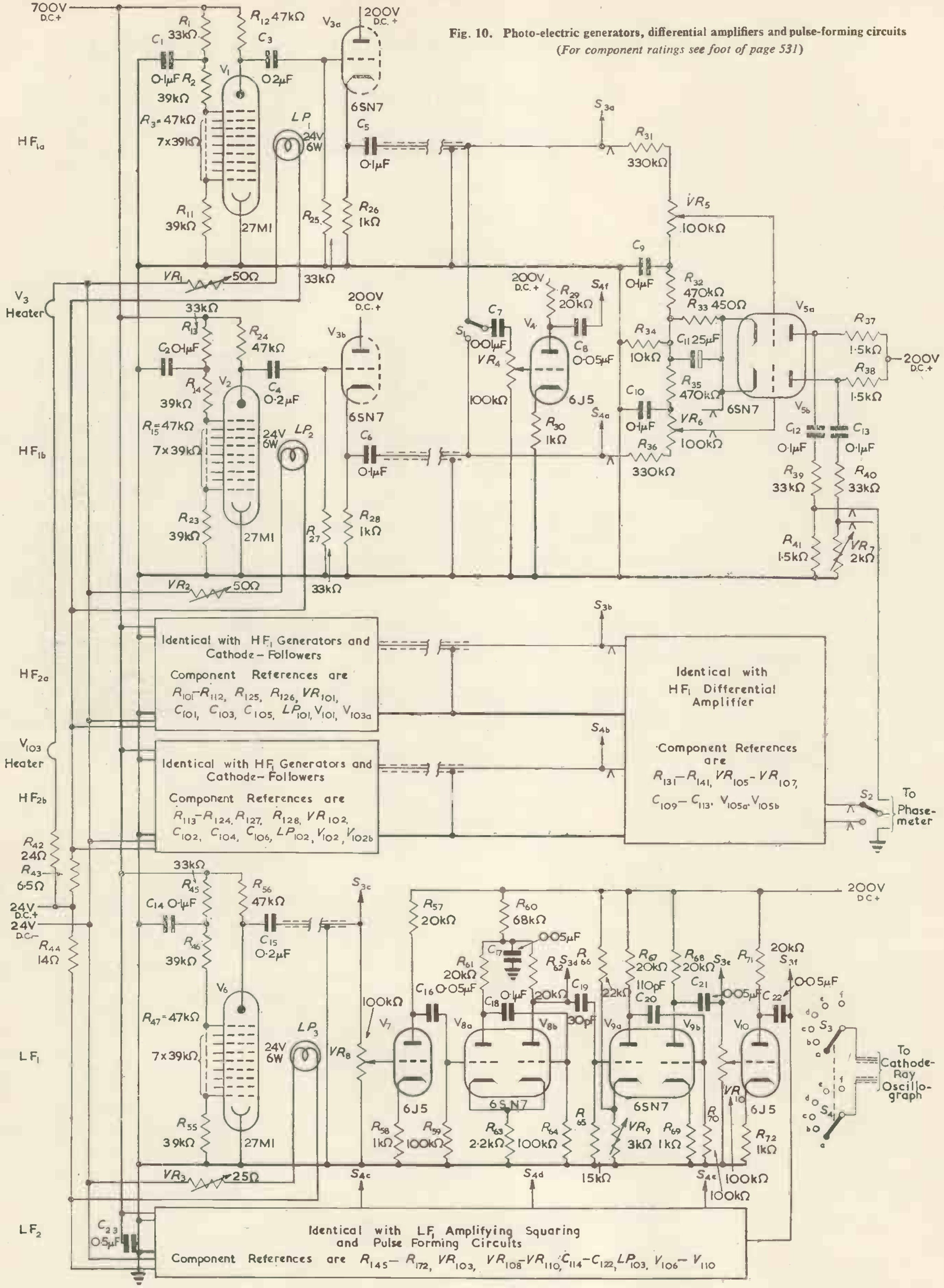
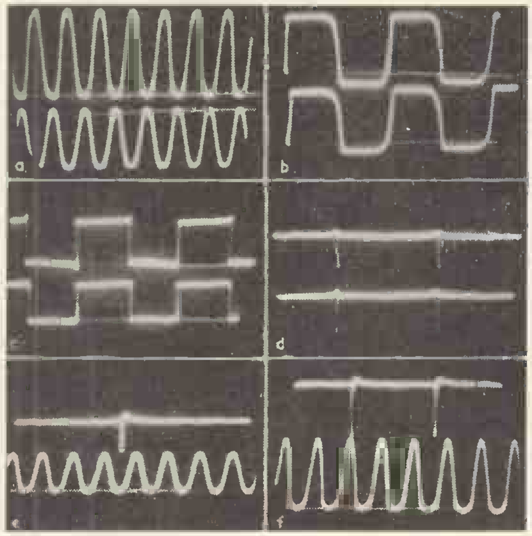


Fig. 10. Photo-electric generators, differential amplifiers and pulse-forming circuits
(For component ratings see foot of page 531)

The two L.F. generators, being associated with the twelve rectangular notches round the periphery of the disk, produce a voltage which has 1/10th of the frequency of that produced by the H.F. generators and has an almost square waveform. There are no cathode-followers associated with the L.F. generators, partly on account of the lower frequency and partly because waveform distortion is not as important as in the case of the H.F. circuits. The output from each generator is first fed via a gain control VR_8 to an amplifier valve V_7. This is followed by a multivibrator, V_{8a} and V_{8b}, which is locked by the signal from V_7. The output of the multivibrator is taken from the anode of V_{8b} and consists of a signal having a waveform which is extremely steep sided and is, therefore, ideal for generating pulses. This signal is applied to a differentiating circuit consisting of C_{19} and R_{65} in series, which produces at the grid of triode V_{9a} a series of alternate positive and negative-going pulses. This triode is biased to cut-off point by means of R_{66} and VR_9, so that only negative-going pulses appear at its anode. These pulses are then amplified in V_{9b} and V_{10}, two stages of amplification being used in order to preserve the negative-going character of the pulses. It should be noted that the coupling capacitor C_{20} has a value which is sufficiently small to give appreciable attenuation of any mains ripple content introduced in the preceding stages. The pulses so derived from the L.F. generators at the two ends of the torsion rod are then combined and fed to one trace of a double-beam cathode-ray oscillograph. One of the two voltages from the H.F. generators at end 1 is fed to the other trace of the oscillograph by a buffer amplifier V_4. This amplifier can be selected through switch S_1 to either of the two inputs to the differential amplifier for end 1. The output of the buffer amplifier can be varied by means of gain control VR_4, which is connected to switch S_1 through C_7, the purpose of the latter capacitor being to prevent VR_4 being connected across C_9 or C_{10} and disturbing the D.C. conditions in the grid circuits of the differential amplifier. The display on the cathode-ray tube takes the form of one H.F. trace, about 5 cycles in length, to which the time-base is locked, and two L.F. pulses which are arranged to be coincident at zero twist. The L.F. pulse which is derived from the same end of the torsion rod as the H.F. trace appears as if locked to the H.F. trace, but the pulse derived from the other end of the torsion rod moves across the screen as the rod twists, and indicates on the H.F. trace the quadrant being measured. Since it will already be known from the position of the milliammeter reversing switch and the input phase-shifting switch whether the quadrant being measured is in the series 1st. 5th, 9th . . ., 2nd, 6th, 10th . . ., 3rd, 7th, 11th . . ., or 4th, 8th, 12th . . ., no great accuracy is required from the cathode-ray oscillograph display and the associated circuits in order to resolve the ambiguity relating to a particular series.

The photocell exciter lamps are supplied from a rectifier having a battery floated across its terminals. An A.C. source of supply for the lamps cannot be used as this would give rise to a ripple in the light intensity which would appear in the photo-cell output. The battery serves to smooth the rectifier output and absorbs variations in output due to sudden changes in the mains voltage. It consists of four 6V 150Ah commercial vehicle batteries which were on hand, although doubtless a battery of smaller capacity would have been equally effective. The lamps are connected to the rectifier and battery through two fixed voltage-dropping resistors, one for the four H.F. generator exciter lamps and the other for the two L.F. generator exciter lamps. In addition each lamp has an individual adjustable dropping-resistor which controls the intensity of the light on the cell. These adjustable resistors comprise the means for varying the photocell output and are also capable of controlling the waveform of the output to a certain extent.



(a) Outputs from pair of H.F. *generators at one end of torquemeter. (b) Outputs from the two* L.F. *generators. (c)* L.F. *generator outputs after squaring. (d) Pulses derived from square waveforms. (e) Normal display. Top trace shows* L.F. *pulses coincident, corresponding to zero torque. (f) Normal display. Top trace shows* L.F. *pulses separated by approximately* 900° *due to torque applied to torsion rod.*

The cathode-followers associated with the H.F. generators are located close to the photo-electric generators and the valve heaters were, therefore, connected in series and operated through a dropping-resistor from the same D.C. supply as is used for the exciter lamps in order to avoid the additional cabling which would be required for an A.C. supply to the heaters.

The double-beam oscillograph is connected through a six-position selector switch so that the waveforms at various points in the circuit can be examined while setting up the apparatus before a run is made, and can be checked during the course of a run. The following are the waveforms which can be displayed for the various switch positions:

Switch Position	*Waveform*
1	HF1A and HF1B generator outputs
2	HF2A and HF2B generator outputs
3	LF1 and LF2 generator outputs
4	LF1 and LF2 square waves
5	LF1 and LF2 pulses
6	Normal display.

COMPONENT RATINGS FOR FIG 10

RESISTOR TOLERANCES

R_{37} and R_{38} are a matched pair. Also R_{137} and R_{138}. All other resistors — ±20%

CAPACITOR WORKING VOLTS AND TOLERANCES

C_{11} and C_{111} 25V. C_{23} 1 000V. All other capacitors 350V. ±20%

RESISTOR RATINGS

½W	1W	2W	4W	7W
R_{29}	R_{34}	R_{44}	R_{43}	R_{42}
R_{57}	R_{134}	R_{66}		
R_{60}		R_{166}		
R_{67}				
R_{68}				
R_{157}				
R_{160}				
R_{167}				
R_{168}				

All other resistors—¼W

VALVE TYPES

27M1 OR 931A	6SN7	½6SN7 OR 6J5
V_1	V_3	V_4
V_2	V_5	V_7
V_6	V_8	V_{10}
V_{101}	V_9	V_{107}
V_{102}	V_{103}	V_{110}
V_{106}	V_{105}	
	V_{108}	
	V_{109}	

Photo-multiplier cell and lamphouse assembly

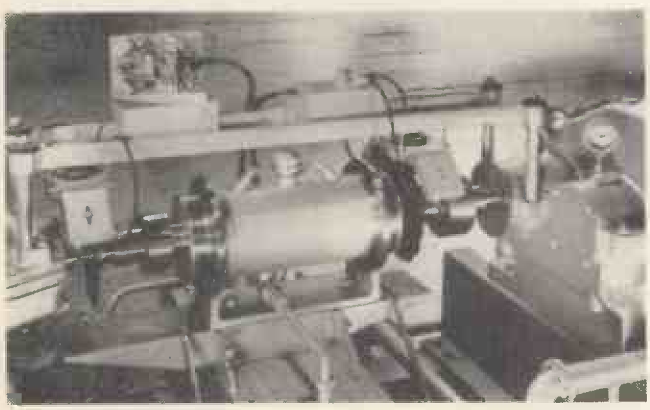
Various methods of mounting the photocells were tried out. Originally they were fixed directly to the adjacent portions of the compressor and gearbox, but the inevitable high frequency vibration at 11 000 rev/min, notwithstanding the good balance, caused the electrodes to disintegrate. This problem was resolved by mounting the whole photocell assembly upon a fairly heavy angle iron frame, which was supported from the main machinery by resilient rubber sleeve bushes. All the photocells were capable of adjustment in the peripheral and radial direction by means of fine thread jack screws with lock nuts. It was found most convenient to lock the lamphouses solidly on to the photocell housings.

The photo-electric cells were purchased complete in their housings together with the divider network R_2 to R_{11} (Fig. 10). Originally they were used as noise generators for jamming enemy radar, and as such embodied a small exciter lamp and its holder. Both the lamp and the holder were removed and a hole drilled in the housing opposite where the lamp filament had been, so that the cell could be excited from an external light source.

The cathode-follower units are shown in Fig. 10. It was necessary to mount these as closely as possible to the photocells in order that the capacitance of the cable carrying the signal from the photocell to the cathode-follower should be kept to a minimum. The photo-electric generators and the cathode-follower units were connected to a distribution box by means of plugs and sockets so that the whole of the beam carrying them could be readily removed at any time. In addition the box contains terminal blocks for cable connexion, the dropping-resistors for lamps and valve heaters, the lamp adjusting resistors, and capacitor C_{23} which it was found necessary to add in order to suppress mains hum picked up on the E.H.T. supply lead.

The photocells are very susceptible to changes in the E.H.T. voltage, appreciable variations in output taking place for comparatively small variations in voltage. It is therefore preferable that a stabilized power pack be used.

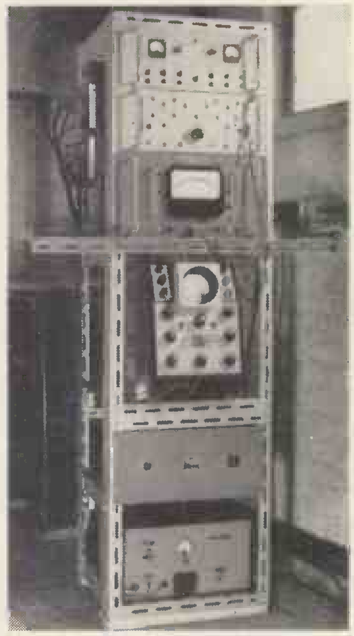
Torque meter assembly with cathode-follower unit



In the present instance an existing unstabilized power pack was employed in preference to buying or constructing a new one, and as a result it was found necessary to make periodic checks of the inputs to the differential amplifiers and, as necessary, to readjust them.

The first stage of setting the equipment to work is to ensure that the optical system is collimated, otherwise there will be loss of output and, in the case of the H.F. generators, distortion of waveform. The lampholders should be so positioned in the lamphouses that the lamp filament is presented edgewise to the photocell and the lamphouse itself should be turned so that the hole in it through which the light passes is directly opposite the hole in the photocell housing.

In the case of the L.F. generators nothing further is required, apart from positioning the disks on the shaft so that the light beams from both photocells are interrupted simultaneously. This will ensure that the L.F. marker pulses are superimposed one on the other at the zero-torque condition.



The torquemeter indicator panel

In the case of the H.F. generators it is necessary to check the H.F. voltage output for magnitude and waveform. This must be done with the torsion bar and its generator disks spinning so as to produce a voltage, but quite a low speed of rotation, say $1/10^{th}$ to $1/5^{th}$ full speed, is adequate for the purpose. The waveforms of the output voltages from each H.F. generator are then observed in turn. If any maladjustment exists, it is most likely to consist of misalignment of the holes in the lamphouse and photocell housing in relation to the holes in the disk. The flexible mounting of the photocell beam provides a ready means of checking such misalignment, since by pushing or pulling on the brackets carrying the photocells and lamphouses it is possible to move them inwards or outwards radially. If either pushing or pulling causes a reduction in output together with a deterioration of waveform, then the photocell is in the optimum position. If on the other hand the output increases and the waveform improves when the bracket is either pushed or pulled, an indication will be given of the direction in which the photocell should be moved in order to obtain the optimum position, the necessary adjustment being made by means of the screwed rods from which the photocell brackets are suspended.

At the same time as the foregoing adjustments are being made, the variable dropping-resistors for the exciter lamps should be set. In the case of the L.F. generators the lamp brilliance need not be any greater than is required to give adequate signal strength and should not in any case be so great that the photocell anode current exceeds the maximum permissible, viz. 1mA. In the case of the H.F. generators, the aim is not only to obtain sufficient signal amplitude as in the case of the L.F. generators, but also to achieve as far as possible an output voltage which has a symmetrical waveform.

It was found that the H.F. waveform exhibited a flat one one side corresponding to minimum light on the photocell. This was assumed to be due to a finite dark period when the light from the lamp was interrupted by the space between the holes in the disk and was countered by increasing the lamp brilliance sufficiently to cause a similar flat to appear on the other side of the waveform corresponding to maximum light on the photocell, this second flat being caused by overloading of the cathode-follower.

It may occasionally be found that one of the H.F. waveforms appears lop-sided. This can usually be corrected either by turning the lamp holder in the lamphouse or by turning the lamphouse itself. In some cases it may be necessary to make both these adjustments. If the condition still persists in spite of these adjustments being made, it may be due to a distorted lamp filament, in which case the lamp should be replaced. The output of the photo-electric generators should be in the region of 2 to 5 volts, which gives adequate signal-to-noise ratio at the input terminals of the equipment on the rack.

The next stage in setting up consists of adjusting the L.F. pulse-forming circuits, again with the torsion rod and disks spinning slowly. The mark-to-space ratio of the output waveform of V_{8b} is to a certain extent dependent on the amplitude of the signal at the grid of V_{8a}. VR_8 is therefore adjusted so that this mark-to-space ratio is about unity. VR_9 is then adjusted so that the cathode bias of V_{9a} is just sufficient to suppress completely pulses which are negative-going at the grid of the valve without at the same time effecting appreciable reduction in the amplitude of pulses which are positive-going at the grid of the valve. Finally, VR_{10} is set so that the pulse amplitude is about 10 volts at the anode of V_{10}.

If the disks have been positioned correctly on the shaft the pulses derived from the two ends should be coincident, but a small separation of the pulses can often be eliminated by altering the mark-to-space ratio of the output waveform of one or both multivibrators by means of VR_8 and VR_{108}. If, on the other hand, it is desired to eliminate a larger separation between the pulses, this can be done only by repositioning the disks on the shaft.

There now remains only the adjustment of the inputs to and outputs from the differential amplifier. The inputs to the amplifier require to be adjusted in respect of both phase and magnitude. Adjustment of magnitude is obtained through potentiometers VR_5 and VR_6. These are set so that the two inputs are equal and of such magnitude as to give approximately 70mV output. Adjustment of phase is done by shifting the photocells on their mounting brackets in a direction tangential to the disk, the aim being to arrange that the two inputs have a phase difference of 180°. The existence of 180° phase difference can be checked by connecting the phasemeter leads to the inputs of the differential amplifier through suitable voltage dividers. A simpler procedure is merely to display the input waveform on the cathode-ray oscillograph by turning the selector switch to the appropriate position, when the phase relationship between the waveforms can be observed. Since any minor departure from the anti-phase condition will be corrected by the amplifier itself, this method was found to be sufficiently accurate in practice, provided two further checks were applied. The first of these consists of examining the waveform of the voltage developed across R_{34} by means of a wander-lead from the oscillograph to the U-point connected to the cathodes of V_5, when any excessive departure from the anti-phase condition will be evident in the form of a fairly large voltage of fundamental frequency across R_{34}. Secondly, the torsion rod can be loaded so that milliammeter has full scale deflexion in the unshunted condition, corresponding to a phase displacement of either 90° or 270°. If S_2 is then operated, the milliammeter should have exactly full scale deflexion in the opposite sense if the amplifier outputs have a phase difference of 180°.

Finally, VR_7 should be set so that the outputs from the amplifiers are equal. If the phasemeter leads are now connected to the outputs from the differential amplifiers and the torsion rod is spun unloaded at low speed, a reading will be obtained on the phasemeter. This reading can be logged as the reading corresponding to zero torque or alternatively it can be reduced to zero either by means of the backing-off current or by a tangential repositioning of the H.F. photocells at one end relative to those at the other end. In doing the latter it will be necessary to check that the 180° phase displacement between the outputs of the photocells is preserved, although with the screw adjustment provided, this does not present any great difficulty if both screws are adjusted by the same amount when moving the cells.

Errors due to variation of the input voltage to the phasemeter must be avoided. The main causes of variation in the input voltage are changes in the E.H.T., H.T., and L.T. D.C. supplies and changes in frequency.

It has been found that the torsion rod used in this torquemeter has a very high degree of elastic stability and the instrument has maintained its calibration even after prolonged intervals between calibration tests, which, for the research work under discussion, had to be performed immediately after each test.

Acknowledgments

In conclusion, acknowledgment is made of the valuable assistance rendered by Messrs. R. L. Gudgeon, S. A. Soutter and S. M. G. Geercke, who together with the authors formed the small team responsible for the design and development of this instrument.

Appendix

Note on the Differential Amplifier Design

It can be shown that the gain of the differential amplifier is:

$$\frac{\mu R_a}{R_a + r_a} \text{ with anti-phase inputs}$$

and:

$$\frac{\mu R_a}{R_a + r_a + 2R_k(\mu + 1)} \text{ with in-phase inputs.}$$

where μ = valve amplification factor

r_a = valve impedance

R_a = anode load

R_k = cathode load

The in-phase inputs are, therefore, attenuated in relation to the anti-phase outputs by a factor:

$$a = \frac{\mu R_a}{R_a + r_a} \times \frac{R_a + r_a + 2R_k(\mu + 1)}{\mu R_a}$$

$$= 1 + \frac{2R_k(\mu + 1)}{R_a + r_a}$$

In general, if a is to be large

$$\frac{R_k(\mu + 1)}{R_a + r_a} \text{ will be} \gg 1$$

and also μ will be $\gg 1$

so that an approximate expression for the attenuation is:

$$a = \frac{2\mu R_k}{R_a + r_a}$$

$$= \frac{2g_m r_a R_k}{R_a + r_a}$$

where g_m = valve mutual conductance.

Again if α is to be large, the denominator $(R_a + r_a)$ should be small. In general, it will be possible to make R_a small compared with r_a, so that the expression for the attenuation further simplifies to:

$$\alpha = 2g_m R_k$$

The amplifier gains are then:

$g_m R_a$ for anti-phase inputs

$R_a/2R_k$ for in-phase inputs.

The design requirements of the amplifier can then be summarized as follows:

(1) For maximum relative attenuation, the mutual conductance of the valve and the cathode load should both be large.

(2) In order that the actual gain of the amplifier and the relative attenuation shall both be independent of the valve impedance, the latter should be large compared with the anode load.

(3) For maximum gain in the amplifier, the anode load should be as large as is consistent with it remaining small compared with the valve impedance as stipulated in (2) above.

A Coaxial Standing-Wave Indicator

For Frequencies near 10 000Mc/s

By F. A. Benson*, M.Eng., Ph.D., A.M.I.E.E., M.I.R.E., and G. V. G. Lusher*, B.Eng., Ph.D.

The design of a precision coaxial standing-wave indicator is described and some measurements with the instrument are discussed. Complete information is given for the construction of the instrument and attention is drawn to difficulties which are likely to be encountered during manufacture. The variations in the characteristics of silicon crystals suitable for use in the detector are also dealt with because they may easily lead to serious errors.

In measuring the attenuation of coaxial cables at frequencies near 10 000Mc/s the authors first made a large number of measurements with a standing-wave indicator in a run of rectangular waveguide. It was found to be more convenient, however, to make such measurements in a system composed wholly of cables and coaxial lines. To do this it was necessary to produce a coaxial standing-wave indicator. Very little information can be found in the literature on such an instrument although the paper by Hirst and Hogg[1] and the book by Barlow and Cullen[2] are both extremely helpful. These authors are mainly concerned with indicators for rectangular waveguide propagating the normal H_{01} mode but many of their remarks apply equally well to indicators for either circular waveguide or air-filled coaxial line. Details are given in this article of an instrument which has proved to be reliable.

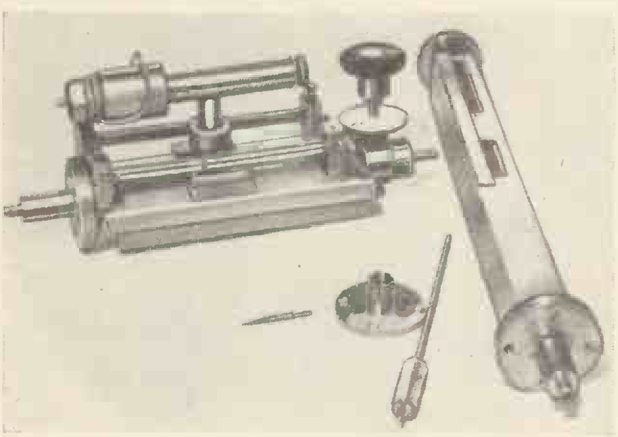


Fig. 1. Construction of the indicator

* *The University of Sheffield.*

The principles of its design are discussed and some tests on it are described. Care has been taken to describe fully the methods adopted for overcoming manufacturing difficulties which arise in producing an instrument of this type.

Repeated calibrations of the instrument with its crystal detector have shown clearly that there are pronounced changes in the characteristics of silicon crystals with time of operation which can lead to large errors if unnoticed. Some interesting facts are, therefore, included regarding the characteristics of these crystals.

Design Principles

The construction of the indicator which is shown in the photograph of Fig. 1, is best dealt with by considering three parts separately; first the coaxial line itself, secondly the probe and detector unit and finally the connexion to the external system.

The Coaxial Lane

The dimensions of the line are shown in Fig. 2. To ensure a constant impedance throughout the entire length of the line the diameter of the inner and outer conductors were machined to ± 0·0005in. The inner conductor was ground from stock silver-steel rod and the outer conductor was bored from square-section brass and finish reamed with a parallel reamer. The characteristic impedance of the line is near 70 ohms as most of the cables employed at frequencies near 10 000Mc/s, including those being tested, have a nominal impedance of this value (e.g. Uniradio 21, 32 and 43). The outside diameter of the line is the same as that of cable type Uniradio 21.

The design of the support for the centre conductor at the cable-sample end presents some difficulties and is rather important because the field pattern must not be disturbed at this point. A distrene spider (shown in Fig. 3) was used in which there is very little dielectric round the centre conductor where the field is strongest. The method of support at the other end is unimportant and a cylinder of distrene was employed as can be seen from Fig. 4.

The slot through which the probe projects is accurately parallel to the common axis of the inner and outer part of the line.

It has been standard practice in the past to silver plate waveguide components of this type but measurements made by one of the authors[3] recently on electroplated waveguides have shown that plated surfaces are generally rougher than drawn ones. Plating may not give the improvements expected and it has not been used in the present case.

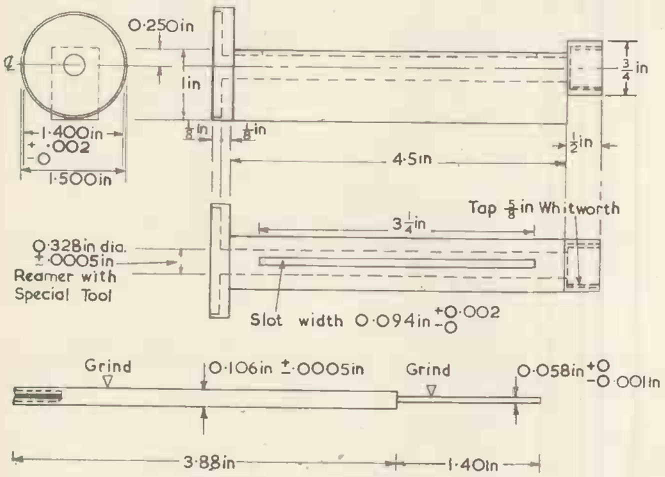


Fig. 2. Dimensions of the line and inner conductor

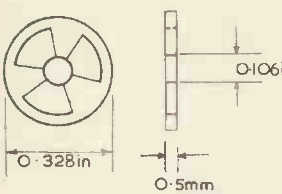


Fig. 3. Distrene spider

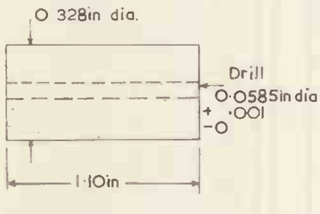


Fig. 4. Distrene cylinder

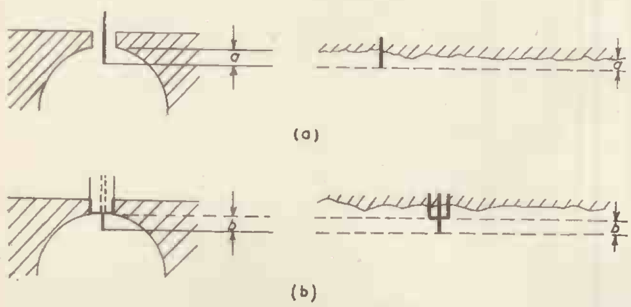


Fig. 5. Effect of shielded probe

The Probe and Detector Unit

A "shielded-probe" type of detector was used in which the "shield" takes the form of a tube round the probe and moves with it so that the effective length of the probe, i.e., the distance the probe projects beyond the shield, remains constant throughout the longitudinal movement of the probe carriage. This ensures that the amount of pick-up is independent of minor irregularities on the surface of the transmission line, as illustrated by Fig. 5. Fig. 5(a) shows an unshielded probe the distance *a* varying about a mean value depending on the inner surface of the line. As *a* is quite small, to avoid disturbing the field in the line, the variations in the amount of pick-up may be considerable. In Fig. 5(b), however, the distance *b*, which is the distance the probe projects beyond the shield, remains constant throughout the longitudinal movement of the probe carriage and the amount of pick-up is independent of any irregularities in the surface of the line.

Drawings of the probe carriage and its transit mechanism are shown in Figs. 6 and 7. The carriage has six degrees of freedom. By imposing five constraints on it there remains only freedom of motion parallel to the slot in the line. Figs. 6 and 7 also show how the constraints are

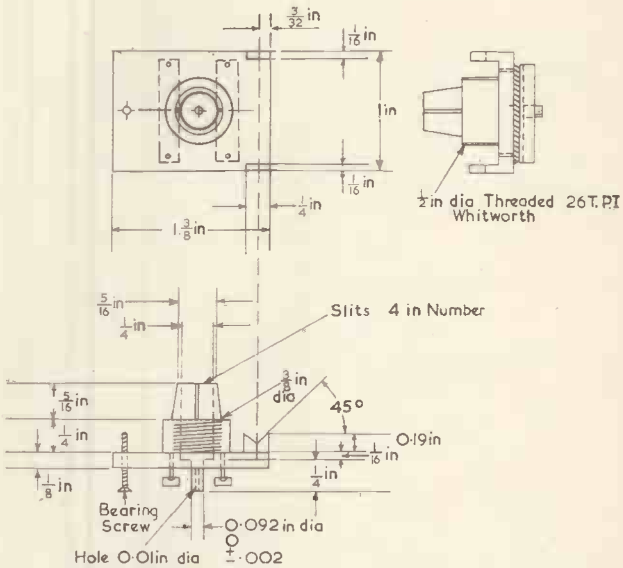


Fig. 6. Probe carriage and transit mechanism

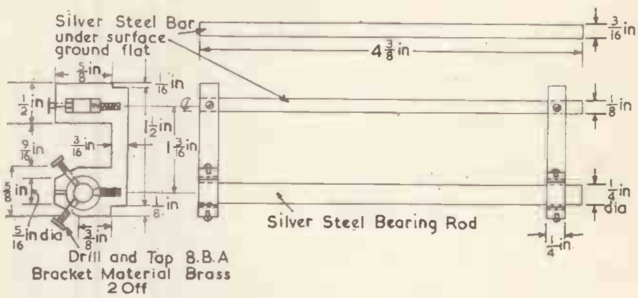


Fig. 7. Guide rods and brackets

imposed. Two bars of ground silver steel, one circular and the other rectangular, are mounted on the line and are adjusted to be parallel with its inside surface. The probe carriage has two V bearings which are forced into contact with the circular bar in two places, thus imposing four constraints and allow the carriage to move round the bar and parallel to it. The former motion is eliminated by the contact on the rectangular bar which has an accurately-finished bearing surface. When the axis of the circular bar and the bearing surface of the rectangular bar are parallel to the axis of the line the probe carriage will move so that the tip of the probe is always the same distance from the centre conductor. The carriage is kept in good mechanical and electrical contact with the bars by springs interposed between the carriage and the main contact shoes.

The main drive is by a thread of silk-nylon clamped at the centre of mass and friction of the probe carriage.

The resulting motion is smooth with negligible backlash; a metric vernier scale accurately determines the probe position.

The detector unit consists of a crystal fitted in a tuned concentric line as shown in Fig. 8. The output from the crystal is connected to a sensitive microammeter.

The Connexion to the External System

The method of supporting the inner conductor has already been described. If a cable of type Uniradio 21 (which has the same diameter as the line) is connected directly to the standing-wave indicator there is a discontinuity due to the change of dielectric and of the inner-conductor diameter. This discontinuity is always present when using the indicator with polythene-filled cables. The type of connector used by the authors is shown in Fig. 9. The outer diameter is the same as cable type Uniradio 21 and the slotted hole in the centre conductor is an interference fit with the inner conductor of this cable. When smaller cables are used an air taper a half-wavelength long is fitted as shown in Fig. 10. The outer conductor of the coupling was machined with a tapered "D" bit, shown in Fig. 11, which was ground and hardened from silver steel. The inner conductors were machined from brass rod and drilled to take the respective cables, the final joint was a sweated soldered one.

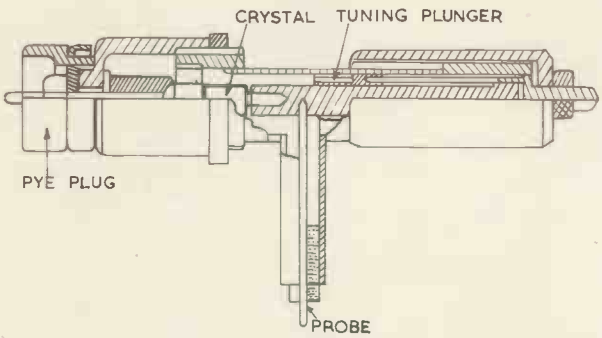


Fig. 8. Crystal detector

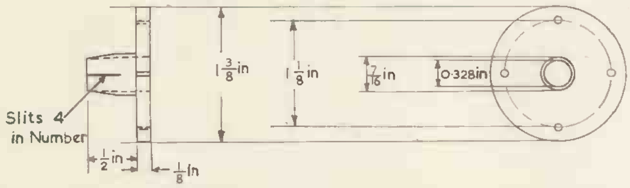


Fig. 9. Connector for cable Uniradio 21
(*Material—brass, silver plated*)

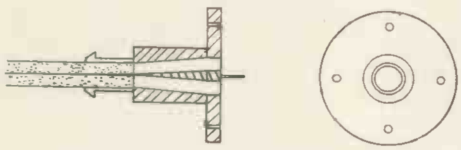


Fig. 10. Tapered connector for Uniradio 32 and 43

Fig. 11. D-bit
(*Silver steel ground to correct taper*)

Measurements with the Instrument

The instrument was first tested by short-circuiting the end of it and taking readings of the field pattern at frequent intervals along the line. In this way the effects of any eccentricity can be observed as differences in the maximum readings of the detector. The result of this test is illustrated by Fig. 12. It will be seen that the error is negligible. It should be noted that this error depends very much on the accuracy with which the distrene spider and cylinder are made. This test on the instrument also makes apparent any errors caused by a probe intrusion which is large enough to disturb the field. The probe can be regarded as an impedance in parallel with the line and will, therefore, have a different effect on the pick-up at different points on the line. If the probe impedance is purely resistive the only result will be a reduction in the amplitude of the maximum reading. If, however, as seems likely, the impedance of the probe has a reactive component the effect produced will depend on the sign of the line impedance at the point of probe intrusion. It follows that the error will not be symmetrical about a maximum or minimum value and in a really bad case the typical result of Fig. 13 can be obtained.

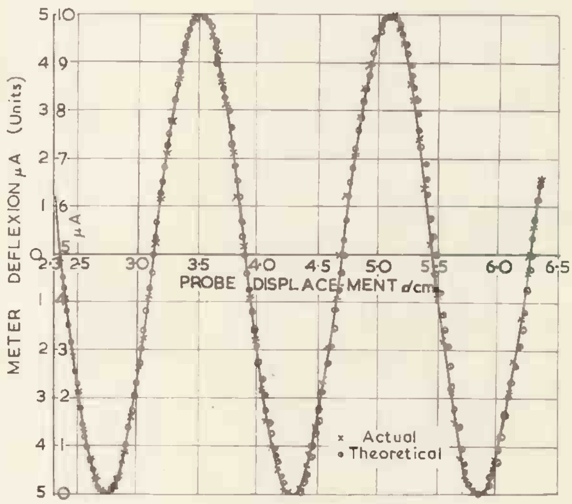


Fig. 12. Meter deflexion/probe displacement

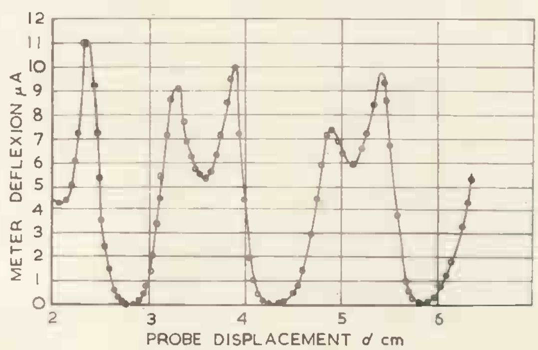


Fig. 13. Meter deflexion/probe displacement

The instrument has been used with success to measure the attenuation of a number of coaxial cables. The block diagram of Fig. 14 shows the arrangement. The impedance presented by the test length of coaxial cable which is backed-off by the tuning piston are measured by the standing-wave indicator. For each position of the piston, three readings of the standing-wave indicator are taken at

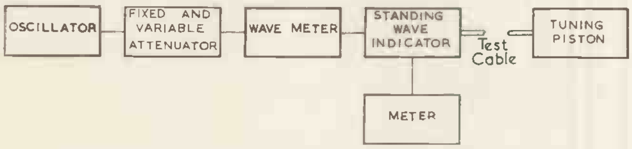


Fig. 14. Arrangement of equipment for cable attenuation tests

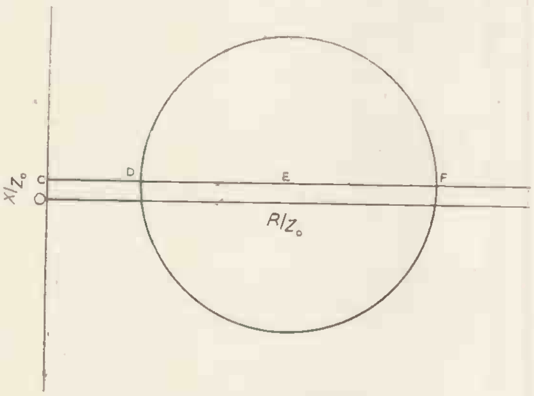


Fig. 15. Impedance circle

$$\alpha l = 8.686 \tanh^{-1} \sqrt{(CD/CF)}$$

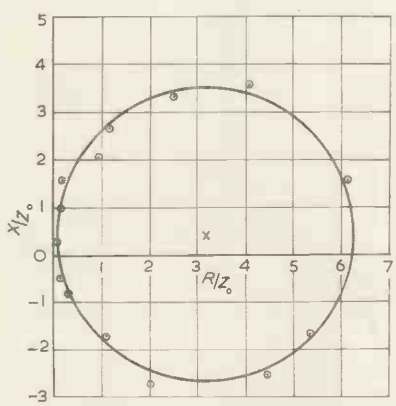


Fig. 16. Typical plot for 6 inch length of Uniradio 43

$$\alpha l = 8.686 \tanh^{-1} \sqrt{(0.15/6.30)}$$
$$= 8.686 \times 0.1553$$
$$= 1.35 \text{ db at room temperature}$$

$\lambda_g/8$ intervals about a node position. If these are V_A, V_B, V_C, from the calibration curve, points on an impedance circle can be obtained by drawing arcs of radii V_A/V_C, and $\sqrt{2}V_B/V_C$ and centres 0, and -1 respectively. The attenuation αl is given by $8{\cdot}686 \tanh^{-1} \sqrt{(CD/CF)}$ from Fig. 15.

Normally about 12 points are required for a circle, a typical result for a 6 inch length of cable type Uniradio 43 is shown in Fig. 16. The location of the points on this diagram confirm that the instrument described is quite reliable.

Crystal Calibration

The method of calibration uses the sinusoidal distribution of the field when the standing-wave indicator is closed by a short-circuiting disk. First a node is accurately located and λ_g is determined; then the carriage position is adjusted to give convenient meter deflexions. A series of meter readings and probe positions is noted and a calibration curve can then be plotted of meter reading against $\sin^2\theta$ where θ represents an electrical angle of $(360° \times x/\lambda_g)$; x being the distance of the probe from the node. Repeated calibrations of the instrument with its crystal have shown that large changes occur in the characteristics of crystals during use but not during storage. Such variations may lead to considerable errors if allowed to pass unnoticed. A large number of unused crystals, both British and American, have been tested, some immediately on receipt from the manufacturer and others having been stored for

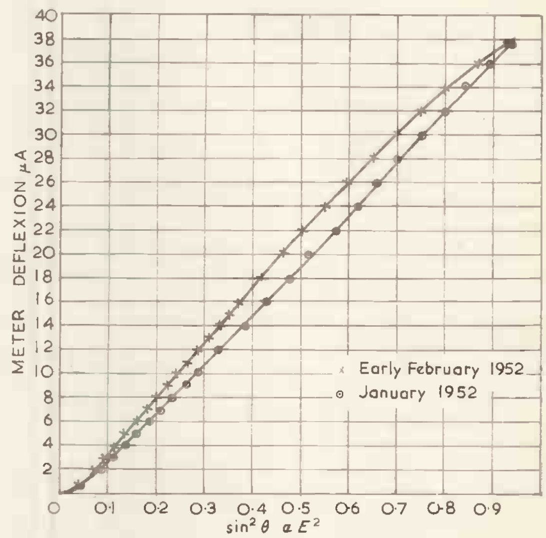


Fig. 17. Calibration of crystal detector
(*Sylvania type* CHS-IN21)

varying periods from 1 to 8 years in the laboratory. Crystals of a given type give almost precisely the same calibration curves. Variations in the characteristics of a particular crystal during use in a standing-wave indicator over a period of 1 month are shown by the typical curves of Fig. 17.

Acknowledgments

The authors wish to thank Mr. O. I. Butler for facilities afforded in the Laboratories of the Department of Electrical Engineering at the University of Sheffield. They are also grateful to Mr. J. Thomas, Manager of the Engineering Workshop, and to the late Mr. A. B. Rawlinson who constructed the instrument and made several helpful suggestions during its design.

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Electrical Breakdown in Waveguides at 3000Mc/s

By J. W. Sutherland*, M.A., A.M.I.E.E.

This article describes details of high power testing techniques in waveguides. Results on waveguides filled with gases other than air are given. The power at which breakdown takes place is related to gas pressure.

ONE of the serious problems in high power radio equipments operating at centimetric wavelengths is the electrical breakdown in the waveguide system.

The power handling capacity of an air-filled waveguide propagating the H_{01} mode is 416 kilowatts per square centimetre of the cross-sectional area of the waveguide, assuming a value of 2.9×10^6 volts per metre as a breakdown figure for air at centimetre wavelengths[1]. Therefore, in size 10 waveguide, which has an internal section of 1·34in × 2·84in, the theoretical breakdown figure is 10·2 megawatts. If, however, there is any discontinuity in the rectangular section of the waveguide, then there will be local electric fields set up of greater magnitude than the uniform field. The power carrying capacity of the waveguide is therefore reduced by the presence of the discontinuity. The magnitude of the electric field set up by a local discontinuity is extremely difficult to calculate, but it can be shown that the electric field in the proximity of a spherical surface is inversely proportional to a function of the radius of the surface. It is therefore desirable in a high power system to avoid surfaces with small radii (i.e. sharp corners, edges, points, etc.), and all corners and bends should be "radiused", i.e. finished with as large a radius as possible. The object of the work reported here was to develop methods of testing high power waveguide elements and to investigate methods of improving their power handling capacity by filling with gases other than air.

Testing Techniques

A flexible test arrangement was required for breakdown testing, and it was decided to build as simple and robust a modulator as possible. The result, which was designed to give maximum peak power for minimum size and circuit complication, was an A.C. resonant charging modulator operating directly from 50c/s mains, capable of giving modulator peak pulse power up to 6mW. The circuit is shown in outline in Fig. 1. A conventional blocking oscillator, synchronized to the mains frequency, was used to "fire" a circuit containing a trigatron. The output of the trigatron was applied to the igniter of the ignitron, which formed the main switch of the modulator, and discharged the artificial line through the pulse transformer. The modulator cubicle also incorporated a power unit to supply the electro-magnet for the magnetron.

The power output of the magnetron was measured with a direct reading thermistor bridge. The waveguide mount containing the bead thermistor was coupled to the main waveguide by a directional feed with a known coupling figure. A preset attenuator was also introduced into the side arm, to adjust the signal level at the thermistor. The thermistor bridge measured mean power, but since the duty cycle of the modulator was known quite accurately, the peak power could be deduced simply. The value of the preset attenuator was adjusted to make the bridge direct reading in peak power, in order to simplify the taking of readings. The bridge was of the type which is balanced with no incident power, and then rebalanced with the power entering the thermistor mount. The rebalancing is done by bleeding off some of the bridge current through a network containing a meter. The network is such that the meter reads directly in power[2].

The breakdown tests were carried out in a special

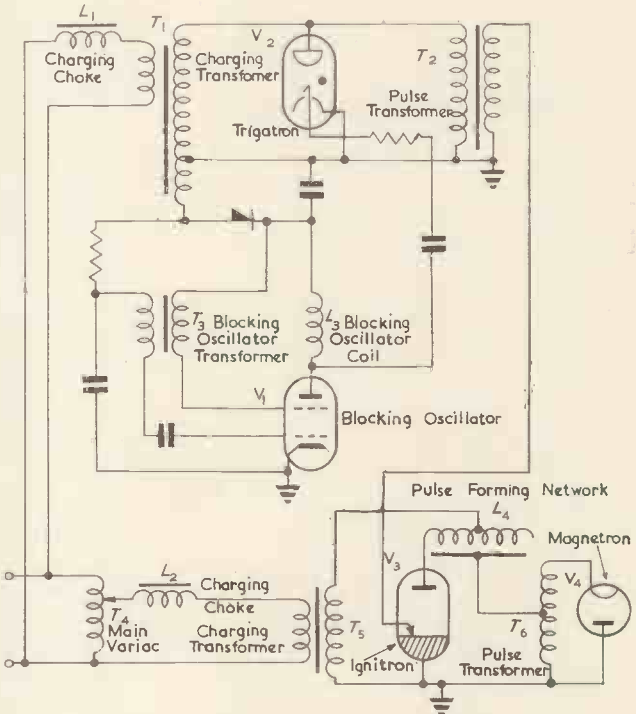


Fig. 1. Outline circuit of 50p/s test modulator

section of waveguide shown diagrammatically in Fig. 2. The section was designed to put a known gap across the waveguide and at the same time present the minimum mismatch. It consisted of a waveguide 8in long, in which the narrow dimension tapered from 1·34in at each end to ¼in in the middle. Each taper was half a wavelength long, and the ¼in gap was maintained for a distance of a quarter of a wavelength. Certain tests were also performed using a similar waveguide section with a ½in gap. To seal off the test section from the rest of the waveguide run for evacuation and gas filling purposes, 6in diameter disks of 0·020in thick mica were used in conjunction with a rubber sealing ring seated in a choke flange. An H-plane corner was introduced into the waveguide run, with a small hole drilled in the vertex of the corner, to facilitate visual inspection of the inside of the test section during breakdown. A diagram of a typical test arrangement is shown in Fig. 3, and the photograph shows the apparatus used for this investigation.

* *Formerly Metropolitan-Vickers Electrical Co. Ltd.*

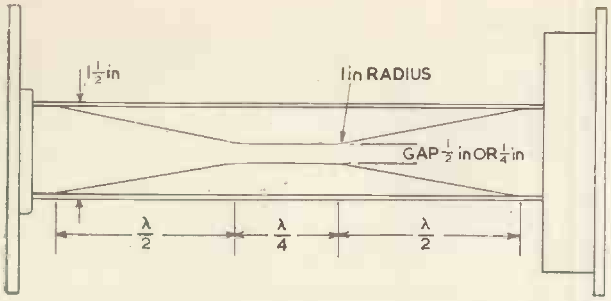


Fig. 2. Test piece for waveguide breakdown

It was important to assess the effect of pulse repetition frequency on the electrical breakdown since most radar equipments operate with a P.R.F. between 500 and 1 500p/s and the tests were to be carried out at 50p/s. A modulator was modified in such a way that the P.R.F. could be varied continuously from 50 to 750p/s. It was found that the initial "ticking" associated with the onset of the breakdown took place at a power level which was independent of P.R.F. However, the power at which a maintained discharge took place decreased very rapidly with increase of P.R.F. Therefore the tests with the 50p/s modulator were all related to the commencement of the "ticking" noise which indicated breakdown on occasional pulses.

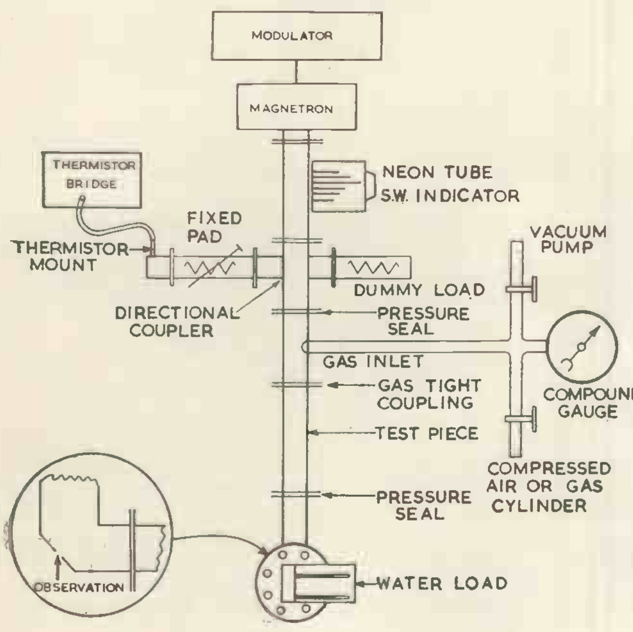
References 3 and 4 describe similar work in this field.

Tests with Air-filled Waveguide

Breakdown tests were carried out using the waveguide test section described above. The air pressure at which the breakdown started was measured for a series of power levels from the modulator. Each reading was repeated several times, and sufficient time was allowed between observations for the recombination of the ionized air. The results were remarkably consistent, and it was not considered necessary to initiate ionization by irradiation or other means. The air pressure was reduced by a small vacuum pump, and increased by a lead from the compressed air line. The manifold system included a compound pressure gauge.

The results are shown in Fig. 4. The full line shows the observed reading in each case, and the broken line is plotted by assuming that the power at which breakdown occurs is proportional to the square of the pressure, and extrapolating from one of the experimental points. The fairly close agreement between observed and calculated points indicates that Paschens law (gas pressure is inversely proportional to the voltage at which breakdown occurs) which was postulated for D.C. potentials, at least approximately holds good at centimetre wavelengths.

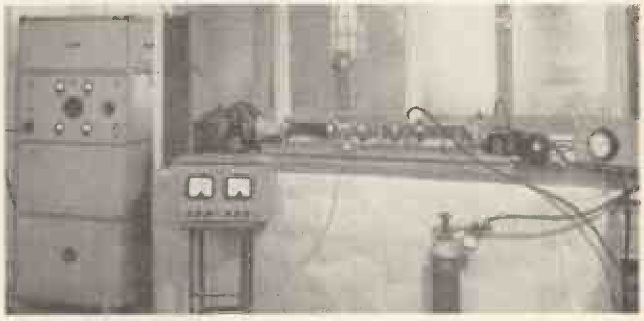
Fig. 3. Typical test arrangement



Tests with Other Gases

Sulphur hexafluoride and difluorodichloromethane (known as "Arcton 6") were used as gas fillings.

Sulphur hexafluoride is a colourless gas without odour or taste. It is completely inert, and in itself non-toxic. It may contain up to 100 parts per million of impurities as supplied commercially, and these may give rise to hydrogen fluoride on hydrolysis. It is, however, safe to use in conditions of reasonable ventilation. Its properties are described more fully in reference 5.



The equipment used for the breakdown tests

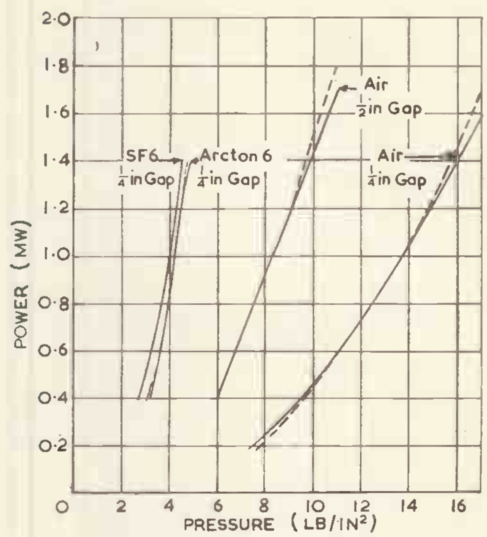


Fig. 4. Results of breakdown tests

"Arcton 6" is a colourless gas with a faint smell, it is totally inert and non-toxic and is not decomposed under normal conditions. Reference 6 gives more details of this compound.

Both are supplied in cylinders in liquid form. The relative cost of these gases is important when their suitability for incorporation into equipments is considered. The approximate cost (February, 1954) of sulphur hexafluoride is £3 to £4 per pound according to quantity, and of "Arcton 6," 4s. 6d. to 6s. 8d. per pound according to quantity.

To conduct the breakdown tests, the test section was totally evacuated, the gas admitted, re-evacuated, refilled and so on several times to ensure the purity of the gas. After each breakdown the section was evacuated and refilled. The results of the tests are recorded in Fig. 4.

It appears from these results that the relative power handling capacity of waveguides filled with air, "Arcton 6" and sulphur hexafluoride is approximately $1:9:10\frac{1}{2}$.

There is no simple explanation of the nature of the discharge, and the mechanism of energy transfer between gas molecules at velocities approaching those corresponding to the ionization potential is so complex that it is incapable of treatment by simple mathematical reasoning (see reference 7).

In conclusion, it appears that "Arcton 6"—a cheap, non-toxic, freely available gas—can increase the power-handling capacity of a waveguide system by a factor of 8 or 9.

Acknowledgments

The author acknowledges the work of his colleagues, particularly in the design of the test modulator and thermistor bridge, and also the technical information provided by Imperial Chemical Industries Ltd. Acknowledgment is also made to Metropolitan-Vickers Electrical Co. Ltd., for permission to publish this article.

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R.F. Cable Characteristics Measured with a Q-Meter

By Jacob Shekel*

The conventional methods of measuring velocity of propagation, characteristic impedance and attenuation are discussed. A method is described of measuring the frequency of half-wave resonance on a Q-meter in such a way that the other parameters can be easily computed.

THE following three characteristics, which are of interest in specifying or testing cables for R.F. applications, will be discussed in this article:

1. Velocity of propagation.
2. Characteristic impedance Z_o, or its inverse, the characteristic admittance Y_o.
3. Attenuation.

There are a number of methods of measuring these characteristics, of which the following are those mostly in use:

1. Velocity of propagation is usually measured by finding the frequency at which a section of the cable exhibits resonance (half-wave or quarter-wave).

2. The capacitance per unit length is measured on a bridge, usually at an audio-frequency. This measurement and the velocity of propagation are sufficient to determine Z_o. An alternative method is to find a terminating impedance that will be transformed in a 1:1 ratio along a cable section of arbitrary length.

3. Attenuation is measured by substituting a calibrated variable attenuator for a section of the cable in a transmission system. The cable section has to be equipped with matched connectors at both ends.

In the method outlined in this article, the frequency of half-wave resonance of a section of the cable is measured on a Q-meter; but this same measurement is conducted in a way that yields enough information to compute the two other characteristics. The obvious advantages of this method are that a single measuring instrument is used, and all characteristics are determined from measurements executed at the same frequency (or, more precisely, in a small frequency range). The cable section is easy to prepare; it is open-circuited (i.e., cut off flush) at one end and any convenient and suitable connector, not necessarily a matched one, is mounted on the other end. It seems to the writer that the method is easily applicable to production testing, by setting limits on the quantities directly measurable on the Q-meter.

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The Method

The cable to be tested has a characteristic admittance Y_o mhos (assumed real), and a propagation constant $\gamma = \alpha + j\beta$ (α in nepers/metre, β in radians/metre). The Q-meter is used to measure the input admittance of a section l metres long, whose end is open-circuited. This input admittance, in the general case, is:

$$\begin{aligned} Y &= Y_o \tanh \gamma l \\ &= Y_o \tanh (\alpha + j\beta)l \\ &= Y_o \frac{\tanh \alpha l + j \tan \beta l}{1 + j \tanh \alpha l \tan \beta l} \end{aligned}$$

Let G and B denote the real and imaginary parts, respectively, of Y, then:

$$G = Y_o \tanh \alpha l \, \frac{1 + \tan^2\beta l}{1 + \tanh^2\alpha l \tan^2\beta l} \quad \ldots\ldots (1)$$

$$B = Y_o \tan \beta l \, \frac{1 - \tanh^2\alpha l}{1 + \tanh^2\alpha l \tan^2\beta l} \quad \ldots\ldots (2)$$

At the frequency of half-wave resonance, f_o,

$$\pi = \beta l = 2\pi f_o l / v$$

where v is the velocity of propagation, in metres/second;

$$v = 2 f_o l \quad \ldots\ldots (3)$$

The input admittance at this frequency is real,

$$G = Y_o \tanh \alpha l \quad \ldots\ldots (4)$$

Another relation between Y_o and α is obtained from the off-resonance behaviour of B. Let $f = f_o + \delta f$,

$$\begin{aligned} \beta l &= 2\pi/v \ (f_o + \delta f)\, l = \pi + 2\pi l/v \ \delta f \\ \tan \beta l &= \tan 2\pi l/v \ \delta f \\ &= \tan 2\pi l / 2 f_o l \ \delta f \\ &= \tan \pi \ \delta f / f_o \end{aligned}$$

If the deviation from the centre frequency f_o is small enough, the angle may be substituted for its tangent. The error of this approximation, as a function of the relative frequency deviation $\delta f/f_o$, is shown in Fig. 1. It should be borne in mind that neither the measurements made on a Q-meter, nor the usually required cable specifications, are of an accuracy greater than 1 per cent.

Another approximation will be made by writing 1 for the denominator in equation (2). The validity of this approximation is dependent on cable attenuation as well as on the frequency deviation from resonance. Fig. 2 shows the attenuation of a half-wave section of three types of coaxial cables. (The attenuation of RG-21/U is intentionally achieved by a centre conductor of nichrome, so that this attenuation is already much higher than the values usually encountered in cables for regular R.F. applications). Fig. 3 shows the error of the approximation as a function of attenuation and frequency deviation, and it is evident that the approximation is valid even under the most extreme conditions that may be expected in practice.

Returning now to the off-frequency behaviour of B, and applying both approximations,

$$B = Y_o (1 - \tanh^2 \alpha l) \pi \ \delta f / f_o.$$

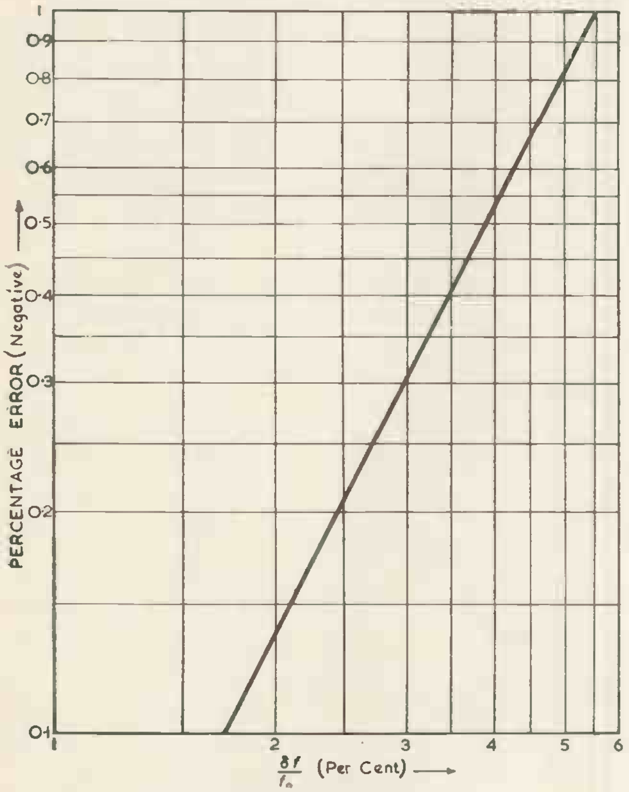


Fig. 1. Error of the approximation $\tan \beta l \simeq \pi \ \delta f/f_o$

The Q-meter measures susceptance as an equivalent capacitance

$$C = B/2\pi f$$
$$= Y_o (1 - \tanh^2 \alpha l) \ \delta f / 2 f f_o$$
$$C = Y_o (1 - \tanh^2 \alpha l) \frac{f - f_o}{2 f f_o}$$

Equations (4) and (5) may be solved for Y_o and α.

Measurement Procedure

A section, whose length corresponds (approximately) to half a wavelength at the test frequency, is cut off the cable to be tested. (The wavelength in cables with polyethylene dielectric is about 2/3 of that in air dielectric or free space propagation). One end of the section is cut off flush, and the other end is prepared for connexion to the capacitor terminals of the Q-meter. An auxiliary coil, that will resonate with about 40 to 80pF at the same frequency, is connected to the coil terminals.

Measurements are made at two frequencies, f_1 and f_2, respectively above and below the frequency of half-wave

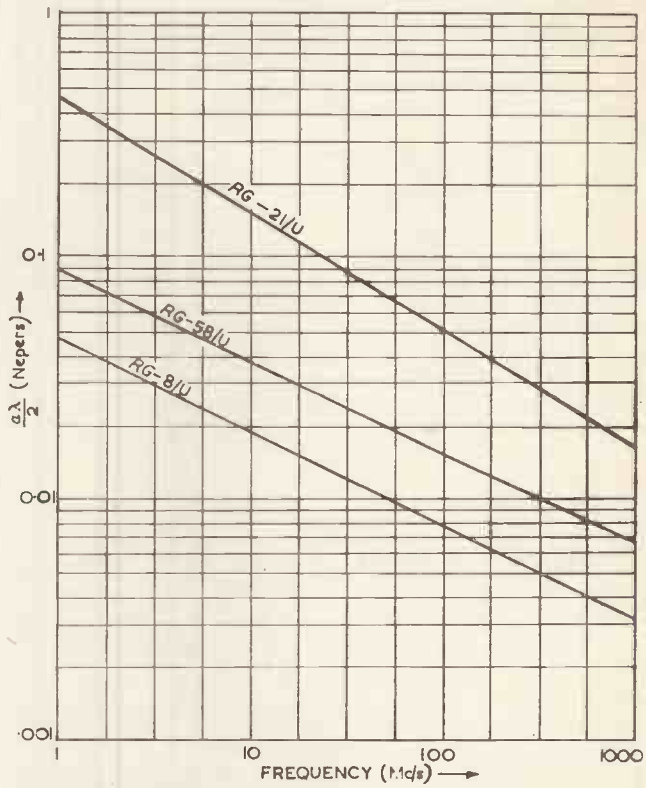
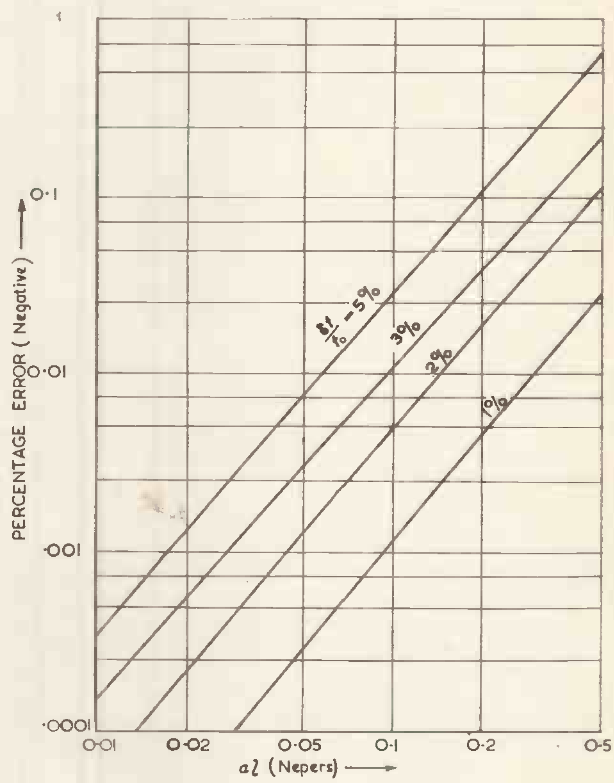


Fig. 2. Attenuation of half-wavelength sections of three types of coaxial cable

Fig. 3. Error of the approximation $1 + \tanh^2 \alpha l \tan^2 \beta l \simeq 1$



resonance f_0. At each frequency, four magnitudes are read:

CpF—the capacitance resonating with the coil, with the cable disconnected.

Q — of the capacitor-coil combination.

C'pF—the capacitance required to return to resonance after the cable has been connected. This value is read on the "increment" dial of the instrument.

Q′ — of the capacitor-coil-cable combination.

The resonance frequency f_0 is then computed by interpolation as the frequency for which $C' = 0$. (C', the negative of the cable equivalent capacitance equation (5), is negative at f_1 and positive at f_2). The values of C, Q and Q′ at f_0 are found by interpolation. The change in Q and Q′ will be very small, if any. The input admittance G

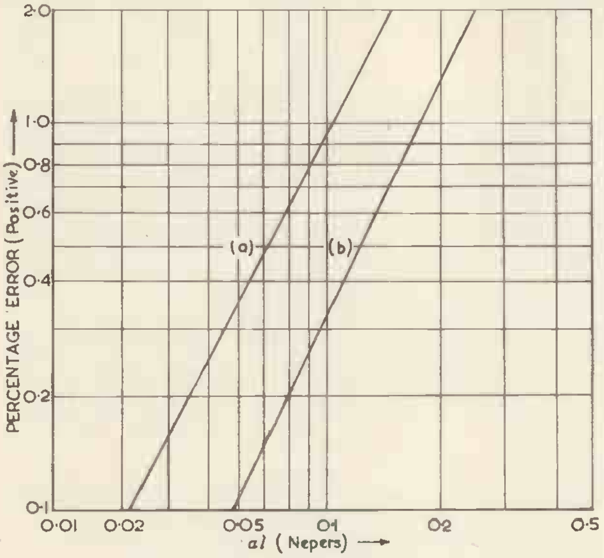


Fig. 4. Errors of the approximations

(a) $1 - \tanh^2 \alpha l \simeq 1$

(b) $\tanh \alpha l \simeq \alpha l$

(at f_0) is then computed, as explained in the instruction manual of the Q-meter.

From equation (5), the capacitance increment C' is:

$$C' = -C = Y_0(-\tanh^2 \alpha l)\ 1/2f_0\ (f_0/f - 1),$$

so that the measured values are:

$$C_1' = Y_0(1 - \tanh^2\alpha l)\ 1/2f_0\ (f_0/f_1 - 1)$$

$$C_2' = Y_0(1 - \tanh^2\alpha l)\ 1/2f_0\ (f_0/f_2 - 1)$$

$$\Delta C = C_2' - C_1' = Y_0(1 - \tanh^2\alpha l)\ 1/2f_0\ (f_0/f_2 - f_0/f_1)$$

$$= Y_0(1 - \tanh^2\alpha l)\ \tfrac{1}{2}\ \frac{f_1 - f_2}{f_1 f_2}$$

Denoting $f_1 - f_2$ by Δf,

$$Y_0(1 - \tanh^2\alpha l) = 2\ f_1 f_2\ \Delta C/\Delta f \quad \ldots\ldots (6)$$

Equations (3), (4) and (6) are used to derive v, Y_0 and α.

It should be noted that in equation (6) use is made of a frequency difference. In order to measure it with the required precision, it is advisable to chose f_1 and f_2 as multiples of 1Mc/s, and to set the Q-meter to the desired frequencies with a heterodyne frequency meter that utilizes harmonics of a 1Mc/s crystal oscillator.

Further simplification of equations (4) and (6) is possible if the attenuation is below certain limits. First, $\tanh \alpha l$ may be approximated by αl, and at even lower attenuations $(1 - \tanh^2\alpha l)$ may simply be approximated by 1. Fig. 4 shows the errors of these approximations as a function of αl, and reference to Fig. 2 gives the expected range of values for αl. Generally speaking, for a given cable, the higher the test frequency the better are the approximations involving αl.

End Effects

At the junction between the cable section and the terminals of the Q-meter, a certain end effect may be expected. The effective length of the section may be somewhat different from the physically measured length, and an additional susceptance is effectively shunted across the cable input admittance. These end effects are very difficult to compute, but it is possible to conduct the measurements so as to minimize them. The following procedure was followed to make the measurements for the illustrative example at the conclusion of this article, and the results, when compared with the cable specifications, prove that the precautions were sufficient.

The end of the cable to be connected to the Q-meter was fitted with a coaxial plug (Amphenol type 83-1SPN), and a coaxial socket (Amphenol type 83-1R) was connected to the terminals of the Q-meter. The measurements of C and Q were carried out with an identical plug screwed into the socket; this plug was removed and the cable connected for measuring C' and Q′. The length l was measured from the open end of the cable to the place where the dielectric was removed for mounting the plug.

Illustrative Example

Cable to be tested: Amphenol cable, type RG-8/U.

Test instrument: Q-meter, Boonton Radio Corporation, type 160-A.

Length of sample: $l = 2{\cdot}14$ metres.

Measurements:

	f_1	f_2
	47·0Mc/s	46·0Mc/s
C	52·0pF	54·6pF
Q	270	270
C'	−3·12pF	+1·33pF
Q′	67	67

By interpolation, $C' = 0$ at $f_0 = 46{\cdot}3$Mc/s

$$v = 2f_0 l$$

$$= 1{\cdot}98 \times 10^8 \text{ metres/second},$$

or, the "velocity factor" of the cable is:

$$\frac{1{\cdot}98 \times 10^8}{3 \times 10^8} = 66 \text{ per cent.}$$

$$\Delta f = 1\text{Mc/s}$$

$$\Delta C = 4{\cdot}45\text{pF}$$

$$Y_0 = 2f_1f_2\ \Delta C/\Delta f$$

$$= 19{\cdot}2 \times 10^{-3} \text{ mho},$$

$$Z_0 = 52{\cdot}1 \text{ ohms}$$

At the frequency $f_0 = 46{\cdot}3$Mc/s, by interpolation,

$$C = 53{\cdot}8\text{pF}$$

$$Q = 270$$

$$Q' = 67$$

which gives the input impedance of the half-wave section as:

$$G = 0{\cdot}175 \times 10^{-3} \text{ mho}$$

$$\alpha l = G/Y_0 = 0{\cdot}00911 \text{ nepers}$$

(this result shows that the approximations made in deriving Y_0 are justified).

$$\alpha = 0{\cdot}00426 \text{ nepers/metre}$$

$$\alpha = 0{\cdot}037 \text{ db/metre.}$$

Acknowledgment

The method outlined in this article was developed by the writer during his employment as Visiting Lecturer at the Electronics and Telecommunication Laboratory, Israel Institute of Technology, Haifa.

A Pulse-Interval Meter

for Measuring Pulse Repetition Frequency

(Part 2)

By A. M. Andrew*, B.Sc., and T. D. M. Roberts*, B.Sc., Ph.D.

Power Unit

This is shown in Fig. 8. Transformer T_2 and rectifier V_{32} provide a D.C. supply which is smoothed and then stabilized by the series-parallel valve voltage stabilizer incorporating V_{33}, V_{34} and V_{35}. This gives a stabilized output of 150 volts, of which the positive side is earthed. A voltage-doubler rectifier V_{36} gives a further 20 volts or so by rectifying a 6·3 volt heater supply, and provides a line at -170 volts.

Transformer T_1 and rectifier V_{27} provide a D.C. supply which is stabilized to give outputs at 350 and at 250 volts. The stabilizers giving these outputs use the -150 volt supply as reference voltage.

The high-voltage winding of T_1 has a current rating of 135mA. A current rating of about 80mA would be sufficient, but this transformer is used because it is a standard type which is commercially available.

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Setting-up Procedure

The supply voltages are first adjusted to their correct values, starting with the -150 volt supply, since this is used as a reference voltage.

To set VR_1 the polarity selector switch is turned to its middle test position. A sine wave of, say, 50c/s is applied to the input of the instrument. VR_1 is adjusted so that the instrument triggers on the smallest possible amplitude of sine wave. The instrument should then be equally sensitive, for a given setting of the gain control, to positive pulses when in the positive pulse position and to negative pulses when in the negative pulse position.

To set VR_2 a high-resistance D.C. voltmeter is connected between V_{23} cathode and earth. With no input to the pulse-interval meter, the voltmeter reads a few volts in excess of 250 volts. A pulse or sine wave input of frequency 100 per second is then applied. VR_2 is adjusted so that the voltmeter reads 50 volts.

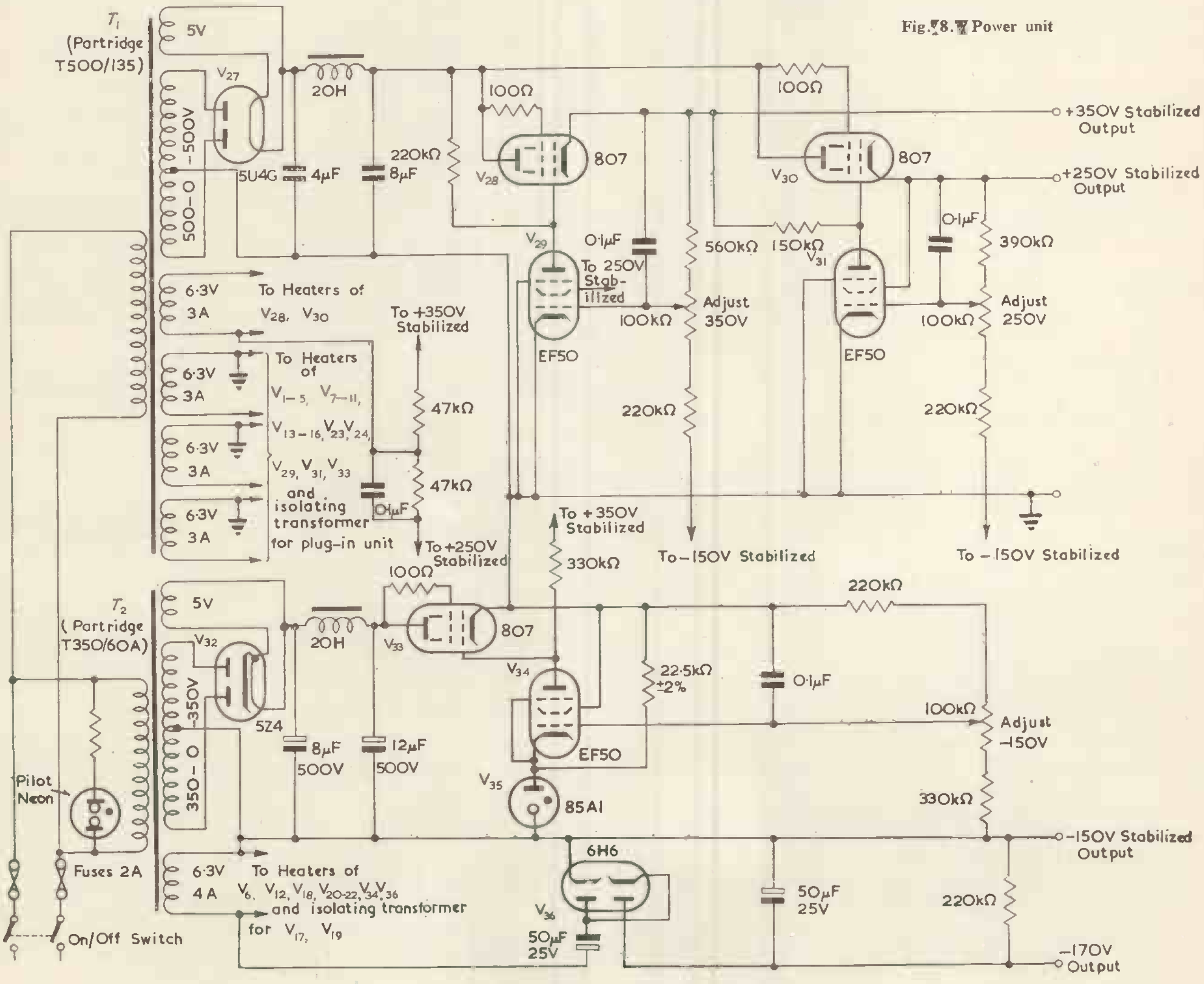


Fig. 8. Power unit

Selection of Valve for V_{18}

Both sections of the double triode V_{18} should have a sharp cut-off, for otherwise the current which still flows when the section is meant to be cut off will produce a voltage across R_7 or R_8. With the D.C. voltmeter connected between V_{23} cathode and earth, and with no input to the pulse-interval meter, there should be no observable change in voltage reading when V_{18} is pulled out. The voltage with V_{18} in place should be read several times, with some input being applied to the pulse-interval meter between readings, to make sure that the test is satisfactory whichever way the changeover circuit happens to stop.

Different valves should be tried until a satisfactory one is found. Of 11 valves type 6SL7 (various makes) which were tried, two were found to be satisfactory. Of 11 Mullard valves type ECC35, ten were found to be satisfactory. The ECC35 is therefore preferred in this position.

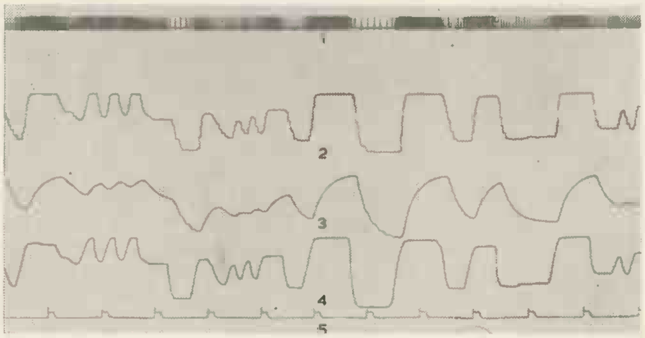


Fig. 9. Comparison of the response of the pulse-interval meter with that of a counting-rate meter to a stream of pulses of rapidly-changing P.R.F. *Trace 1: Pulses from monitor output of pulse-interval meter. Trace 2: Output of pulse-interval meter. Trace 3: Output of counting-rate meter. Trace 4: Position of frequency control of pulse generator. Trace 5: Timing marks at 1-second intervals*

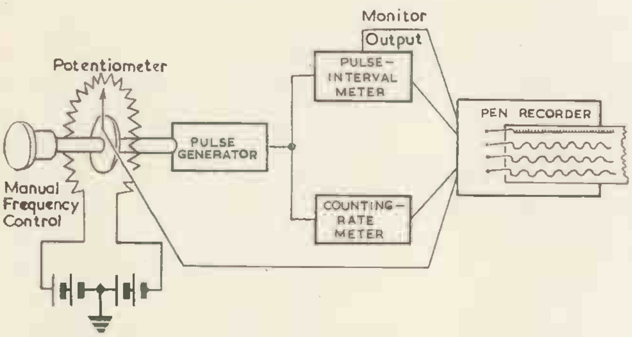


Fig. 10. Arrangement used for making comparisons as shown in Fig. 9

Performance

The instrument has proved satisfactory in use. We have not measured the long-term stability of calibration, but any drift which may occur is certainly not enough to be troublesome. The superiority of the instrument over a counting-rate meter is illustrated in Fig. 9, a piece of record taken with the arrangement shown in Fig. 10 as the frequency control of the pulse generator was twiddled. It will be seen that traces 2 and 4 agree fairly closely, but trace 3 is more sluggish. Fig. 11 shows a record of the varying repetition frequency of the impulses from a sense-organ in the cat knee-joint while the knee was rhythmically flexed and extended by a mechanical device (Boyd and Roberts, not yet published). The relation between P.R.F. and the position of the joint is very clearly brought out by the trace obtained with the pulse-interval meter.

Acknowledgments

The pulse-interval meter was constructed using components purchased by a grant from the Carnegie Trust for the Universities of Scotland. The development of the instrument was facilitated by the careful workmanship of Mr. J. J. Brown.

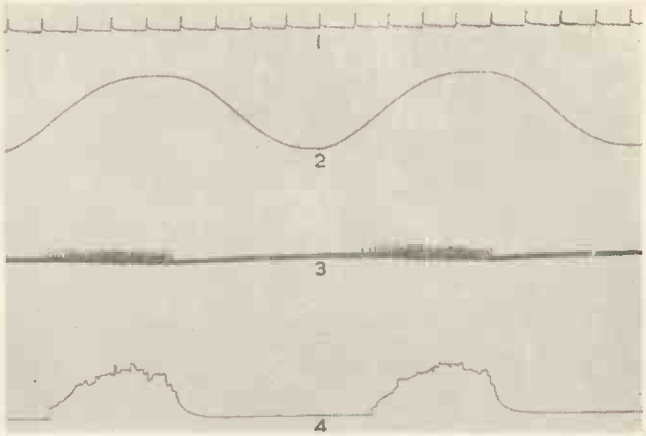


Fig. 11. Record of the response of a sense-organ in the cat knee-joint to mechanical stimulation. *The knee was rhythmically flexed and extended by a mechanical device. Trace 1: Timing marks at ½-second intervals. Trace 2: Record of angular position of the knee-joint. Trace 3: Pulses from the monitor output of the pulse-interval meter (50 c/s interference also appears on this trace owing to a fault in the amplifier driving the pen). Trace 4: Record of pulse repetition frequency given by the pulse-interval meter*

Appendix A

Design of the Charging Network for the Single-diode Approximation

Values for the resistors in Fig. 12(a) have to be found which will ensure that the capacitor charges approximately according to equation (2), after reaching $V = 50$ at time $t = 1/100$, where V is the voltage across the capacitor.

Consider first Fig. 12(b), where the capacitor C is initially charged to V_A, and the switch S is opened to initiate the discharge at time $t = 1/100$. Let $V_B < V_A$, so that at first the capacitor discharges through both R_A and R_B, but when the voltage across the capacitor becomes less than V_B the discharge is through R_A only. The discharge will be made to conform as nearly as possible to the equation:

$$V = V_A/100t \text{ for } t > 1/100 \qquad (4)$$

The discharge from $V = V_A$ to $V = V_B$ is represented

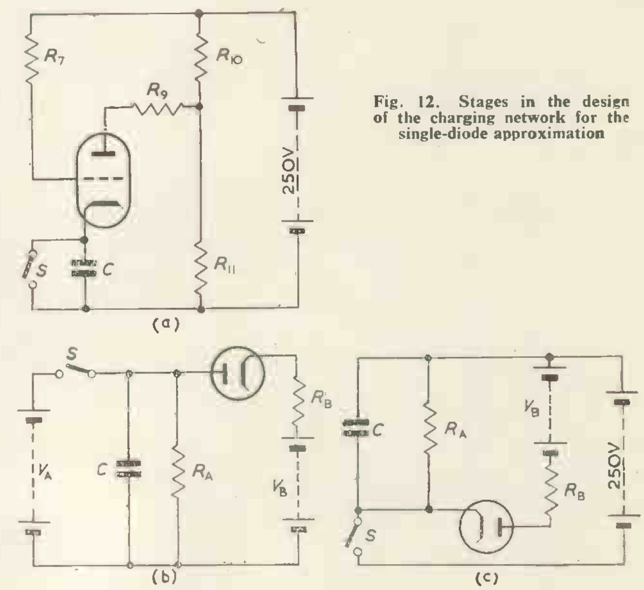


Fig. 12. Stages in the design of the charging network for the single-diode approximation

by:

$$V = V_B\frac{R_A}{R_A + R_B} + \left(V_A - V_B\frac{R_A}{R_A + R_B}\right)\exp-(t-\tau)/R'C \quad \ldots\ldots\ldots\ldots (5)$$

where $R' = R_A R_B/(R_A + R_B)$ and $\tau = 1/100$.

Let t_1 be the time (which can be calculated from equation (5)) at which $V = V_B$.

Then the discharge from $V = V_B$ towards $V = 0$ is represented by:

$$V = V_B \exp-(t - t_1)/R_A C \quad \ldots\ldots\ldots\ldots (6)$$

It has been found by trial and error that if

$V_B = 0{\cdot}2475\, V_A$, $CR_A = 0{\cdot}0825$sec, and $CR_B = 0{\cdot}01175$sec . . (7)

the composite curve represented by equations (5) and (6) is very close to the curve represented by equation (4). The curves are shown in Fig. 13.

The value of $0{\cdot}05\mu$F has been chosen for the timing capacitors. Putting $C = 0{\cdot}05\mu$F in equation (7) gives $R_A = 1{\cdot}65$MΩ and $R_B = 235$kΩ.

Now consider Fig. 12(c). In this the circuit of Fig. 12(b) is inverted. The capacitor is charged to 250 volts when S is closed. Suppose S is opened at a time earlier than t = 1/100 such that $V = 200$ at time $t = 1/100$. Then $V_A = 200$ volts.

If $V_B = 200 \times 0{\cdot}2475 = 49{\cdot}5$ volts, and R_A and R_B are as above, the voltage across the capacitor is given approximately by (see equation (4)):

$$V = 200/100t = 2/t \text{ for } t > 1/100 \quad \ldots\ldots (8)$$

Now consider Fig. 12(a) once more. Fig. 12(a) differs from Fig. 12(c) in the following respects:—

(a) The diode is replaced by a triode, and R_7, corresponding to R_A, is connected to the grid. A 6SN7 triode with zero grid bias behaves roughly as a resistance of 10kΩ.

(b) R_B and the 49·5 volt battery have been replaced by a resistor network. For the two circuits to be equivalent we must have:

$$R_B = 235\text{k}\Omega = 10\text{k}\Omega + R_9 + \frac{R_{10}R_{11}}{R_{10} + R_{11}} \quad \ldots\ldots (9)$$

and

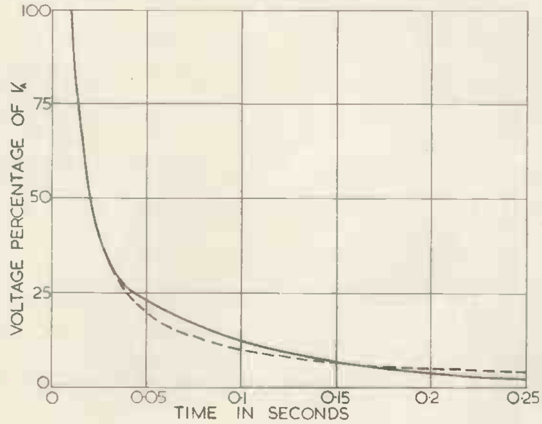
$$\frac{R_{10}}{R_{10} + R_{11}} = 49{\cdot}5/250 \quad \ldots\ldots\ldots\ldots (10)$$

The 10kΩ term is included in equation (9) to allow for the resistance of the triode valve.

(c) The plate of the capacitor which goes to the positive side of the 250 volt supply in Fig. 12(c) goes to the negative side in Fig. 12(a). Since both sides of the supply are at fixed potentials, this alteration does not affect the working of the circuit, but the voltage across the capacitor, instead of being given approximately by equation (8) is now given approximately by:

$$V = 250 - 2/t \text{ for } t > 1/100 \quad \ldots\ldots\ldots (11)$$

Fig. 13. Comparison of the decay of voltage given by the circuit of Fig. 12(b) (solid line) with the ideal curve represented by equation (4) (broken line)



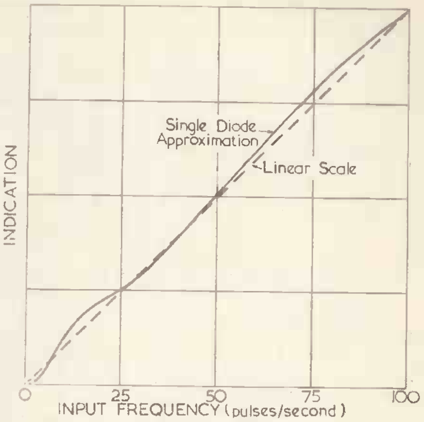


Fig. 14(a). Calculated calibration curve for instrument employing the single-diode approximation

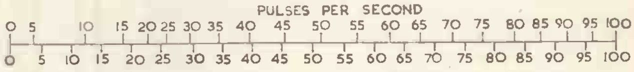


Fig. 14(b). Upper scale: Frequency scale of instrument employing the single-diode approximation. Lower scale: Linear scale for comparison

Since equation (11) is equivalent to equation (2), the desired result has been achieved.

R_{11} has been given the value 470kΩ since this is an approved value. By substituting in equations (9) and (10), and solving for R_9 and R_{10},

$$R_9 = 132\text{k}\Omega,\ R_{10} = 116\text{k}\Omega.$$

Also $R_7 = R_8$ (Fig. 5) $= 1{\cdot}65$MΩ, for both R_7 and R_8 correspond to R_A.

These resistance values are used in the practical circuit (Fig. 5).

Fig. 14(a) shows a calculated calibration curve for an instrument employing the single-diode approximation to the desired charging law and conforming to equations (7). Fig. 14(b) shows the form of the frequency scale, with a linear scale for comparison. These figures are drawn from calculated values. The experimental calibration curve and scale are very similar to those shown.

Appendix B

Design of the Charging Network for the Five-diode Approximation

Values for the resistors in Fig. 7(c) have to be found which will ensure that the capacitor charges approximately according to equation (2), after reaching $V = 50$ at time $t = 1/100$, where V is the voltage across the capacitor.

Consider first Fig. 15(a), where the capacitor C is initially charged to 200 volts and the switch S is opened to initiate the discharge at time $t = 1/100$. The discharge has to conform as nearly as possible to the equation:

$$V = 2/t \text{ for } t > 1/100 \quad \ldots\ldots\ldots\ldots (8)$$

The current leaving the capacitor if equation (8) is obeyed is given by:

$$i = -C\, dV/dt = 2C/t^2$$

i.e.

$$i = CV^2/2 \quad \ldots\ldots\ldots\ldots\ldots\ldots (12)$$

The part of Fig. 15(a) to the right of the broken line has to be designed so that the current and voltage are related approximately by equation (12).

The graph of i against V for an arrangement of five diodes, resistors and batteries as shown in Fig. 15(a) is made up of six straight lines. Fig. 16 shows an approximation, consisting of six straight lines, to the curve represented by equation (12). The sections are numbered from the origin outward and the co-ordinates of the "corners" are shown. Values can be assigned to the resistors and battery voltages in Fig. 15(a) such that the relationship between i and V corresponds to this six-line approximation.

H

If the battery voltages are made to correspond to the voltage co-ordinates of the "corners" of the approximate curve, so that $V_a = 22{\cdot}5$, $V_b = 54$, $V_c = 91$, $V_d = 132$ and $V_e = 173$, the gradient of section 1 of the curve corresponds to the conductance of R_o, the gradient of section 2 to the conductance of R_o and R_a in parallel, and so on. To determine the resistance values it is convenient to tabulate the quantities shown in Table 1.

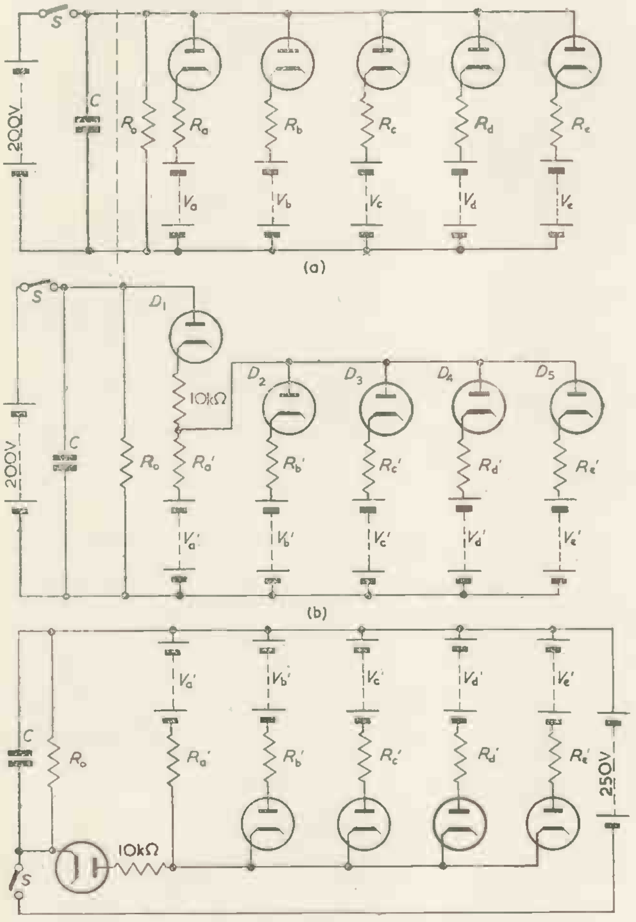


Fig. 15. Stages in the design of the charging network for the five-diode approximation

TABLE 1

SECTION OF CURVE	GRADIENT = CONDUCTANCE	INCREMENT IN CONDUCTANCE	RESISTANCE VALUE	RESISTOR
1	$0{\cdot}444 \times 10^{-6}$	$0{\cdot}444 \times 10^{-6}$	2·25MΩ	R_o
2	1·90 ,,	1·456 ,,	687kΩ	R_a
3	3·65 ,,	1·75 ,,	571kΩ	R_b
4	5·48 ,,	1·83 ,,	546kΩ	R_c
5	7·68 ,,	2·20 ,,	454kΩ	R_d
6	9·45 ,,	1·77 ,,	564kΩ	R_e

With the above values of battery voltage and resistance the circuit of Fig. 15(a) will give a discharge conforming closely to equation (8).

Before deriving values for the practical circuit of Fig. 7(c), however, an additional complication must be introduced. In Fig. 7(c) the four diodes are in series with the triode section, which has an internal resistance of about 10kΩ. Instead of Fig. 15(a), therefore, Fig. 15(b) must be considered.

By an extension of the method used to find values for Fig. 15(a), values have been found for the resistances and battery voltages in Fig. 15(b). They are as follows:

$R_a' = 676\text{k}\Omega$, $R_b' = 546\text{k}\Omega$, $R_c' = 507\text{k}\Omega$, $R_d' = 390\text{k}\Omega$, $R_e' = 478\text{k}\Omega$.

$V_a' = 22{\cdot}5$, $V_b' = 53{\cdot}5$, $V_c' = 89{\cdot}2$, $V_d' = 128$, $V_e' = 167$.

That these values give a discharge conforming to equation (8) can be shown as follows:

Consider the stages in the discharge in the reverse order. For $V < 22{\cdot}5$ none of the diodes conducts, so the conductance $= 1/R_o = 1/(2{\cdot}25 \times 10^6) = 0{\cdot}444 \times 10^{-6}$

When $V = 22{\cdot}5$, diode D_1 starts to conduct.

For $22{\cdot}5 < V < 54$, diode D_1 conducts, so conductance of circuit

$$= 1/R_o + 1/(R_a' + 10\text{k}\Omega) = 1.90 \times 10^{-6}$$

When $V = 54$, diode D_2 starts to conduct, for its anode potential

$$= (54 - V_a')\frac{R_a'}{10\text{k}\Omega + R_a'} + V_a'$$

$$= (54 - 22{\cdot}5)\ 676/686 + 22{\cdot}5 = 53{\cdot}5$$

= its cathode potential.

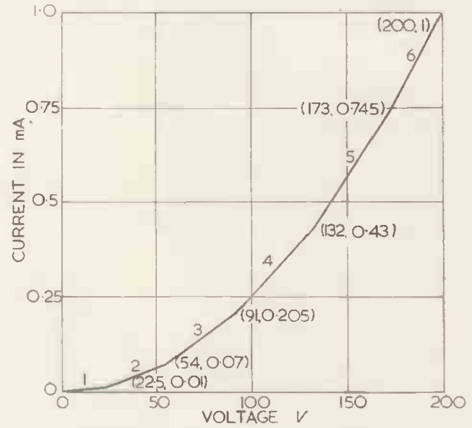


Fig. 16. Approximation, consisting of six straight lines, to the curve represented by equation (12). *The approximation is so close that in a small figure it is not possible to show the exact curve for comparison on the same diagram owing to the thickness of the printed lines*

For $54 < V < 91$, diodes D_1 and D_2 conduct, so conductance of circuit

$$= 1/R_o + \frac{1}{10\text{k}\Omega + R_a'R_b'/(R_a' + R_b')} = 3{\cdot}65 \times 10^{-6}$$

Proceeding in this way it can be shown that the diodes commence conducting in turn at the voltages corresponding to the "corners" in Fig. 16, and the conductance of the network of resistors in which current is flowing corresponds in each voltage interval to the value in column 2 of Table 1. Hence the circuit of Fig. 15(b) is equivalent to that of Fig. 15(a) and will give a capacitor discharge conforming closely to equation (8).

The circuit of Fig. 15(b) is now inverted to give Fig. 15(c), in which the switch S is opened at a time earlier than $t = 1/100$ such that $V = 200$ at time $t = 1/100$. Then the practical circuit of Fig. 7(c) is obtained by transferring one side of the capacitor from the +250 volt line to the −250 volt line, introducing the triode valve, and replacing each resistor and battery by a network of two or three resistors, as was done in deriving Fig. 12(a) from Fig. 12(c). The calculated values for the resistors are shown in Fig. 6. The practical circuit with these values then determines a charging process conforming closely to equation (2).

While operating the pulse-interval meter with the plug-in unit of Fig. 6, the voltage at V_{23} cathode has been plotted against the frequency of the incoming pulses. The graph does not differ significantly from a straight line, the deviations from strict linearity being about one volt at most, in a range of 250 volts.

Appendix C

Obtaining a Non-linear Frequency Scale

The plug-in unit shown in Fig. 17(a) gives a useful non-linear frequency scale. When this unit is plugged in, the circuit of part of the timing units becomes that of Fig. 17(b). The capacitor in either timing unit is simply charged through two resistors in parallel, each of value 2MΩ. With this plug-in unit it is not possible to set VR_2 so that the potential of V_{23} cathode is 50 volts for an input frequency of 100 per sec. If, instead, VR_2 is set so that V_{23} cathode is at 15 volts for this frequency, the frequency calibration curve and scale are as shown in Figs. 18(a) and (b). The circuit of Fig. 17(b) is preferred to those of Fig. 17(c) and (d) for a reason which is dealt with in Appendix D.

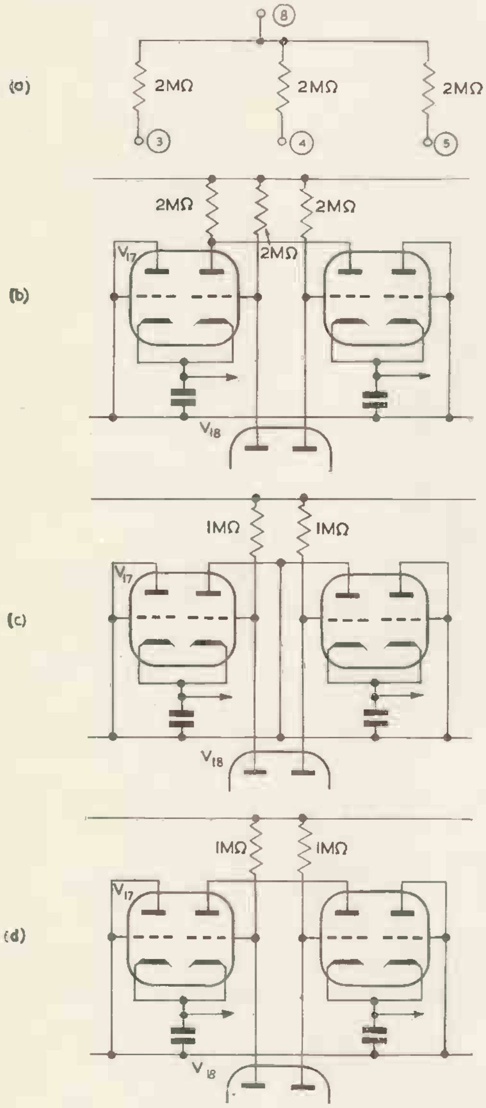


Fig. 17(a). Plug-in unit for non-linear frequency scale

(b). Part of timing units, incorporating plug-in unit of Fig. 17(a)

(c) and (d). Unsuitable alternatives to circuit of Fig. 17(b) (*see text*)

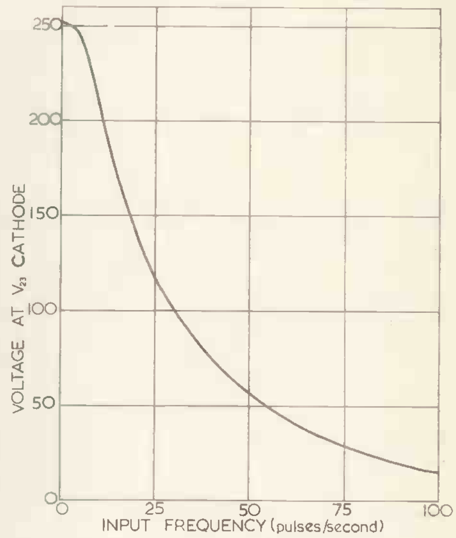


Fig. 18(a). Calibration curve for instrument with plug-in unit of Fig. 17(a)

(b). Frequency scale for instrument with plug-in unit of Fig. 17(a)

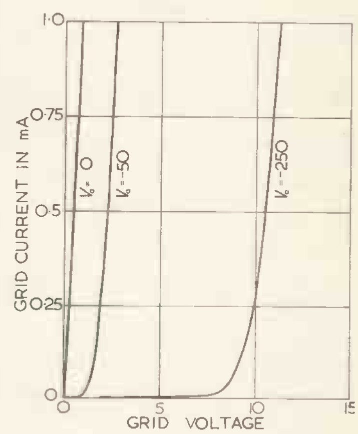


Fig. 19. Grid current characteristics of 6SN7 triode

Appendix D

The Inverted Triode

When the pulse-interval meter is operated with any of the plug-in units of Figs. 5, 6 or 17(a), the cathode potential of V_{23} corresponding to zero input frequency is one or two volts more positive than the 250 volt line. When it incorporates the circuit of Fig. 17(c) or (d), the potential corresponding to zero frequency is three or four volts more negative than the 250 volt line.

The explanation appears to lie in the fact, pointed out by Terman[6] that the anode potential of a triode valve affects the grid current. In Fig. 17(c), when the cathode potential of V_{17} approaches 250, the anode is about 250 volts negative with respect to the cathode. As can be seen from Fig. 19, when the anode is 250 volts negative the grid current falls to a very low value for values of grid-cathode voltage less than about eight. In the circuit of Fig. 17(c) the capacitor is charged by grid current only. If there is a small amount of leakage from the capacitor, perhaps owing to grid current in V_{23} or to anode current in V_{21}, the charging process will stop when the grid current in V_{17b} falls to a low value.

In the circuit of Fig. 17(d) the potential of V_{17b} and V_{19a} anodes is not known, but the results obtained are similar to those obtained with the circuit of Fig. 17(c).

It appears to be advisable, in order that the frequency scale obtained using the network should be calculable, to design charging networks so that the anodes of V_{17b} and V_{19a} never become highly negative with respect to their cathodes during the charging process. The plug-in unit of Fig. 17(a) conforms to this recommendation.

REFERENCE

6. Terman, F. E. The Inverted Vacuum Tube, A Voltage-Reducing Power Amplifier. *Proc. Inst. Radio Engrs. 16*, 447 (1928).

The Transistor as a D.C. Amplifier for use in Microwave Measurements

By C. F. Davidson*, A.M.I.E.E.

A D.C. *amplifier using a junction type transistor is described which enables the sensitivity of a crystal rectifier used for microwave power measurements to be increased. The application of bias to the crystal rectifier causes a further increase in sensitivity.*

IN microwave measurement work it is frequently necessary to measure low-level signals; examples of this occur in wavemeters, standing-wave indicators, power monitors, etc. If the microwave signal can be pulsed at an audio frequency no difficulty arises in the detection of low-level signals since a silicon crystal valve rectifier followed by an audio frequency amplifier-detector can be used. If it is not convenient to pulse the microwave signal then a superheterodyne receiver may be employed; this has the advantage of sensitivity but the equipment is relatively complex and a local oscillator has to be tuned. For the measurement of C.W. microwave powers greater than about $1\mu W$ a silicon crystal rectifier followed by a galvanometer is adequate.

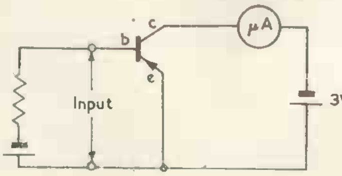
Use of a D.C. Amplifier to Improve the Sensitivity of the Crystal Rectifier-Galvanometer Arrangement

The use of a silicon crystal rectifier and galvanometer suffers from the disadvantage that the sensitivity is limited by that of the galvanometer and also while galvanometers are satisfactory for use in the laboratory they are not robust enough for field use or for installation in equipment. The use of a D.C. amplifier with a gain of say 30 to 50 times would improve sensitivity and enable the galvanometer to be replaced by a more robust micro-ammeter. A single stage valve type D.C. amplifier is not very satisfactory. Suppose the rectified crystal current is $1\mu A$, the internal resistance of the crystal rectifier at this current level is about $5k\Omega$ so that the internal E.M.F. is 5mV. If this voltage is applied to the grid of a valve with a mutual conductance of 10mA/V the change in anode current is $50\mu A$. This change in anode current is only a small fraction of the total anode current, a backing-off circuit must therefore be used for the indicating meter and a very small change in anode current, due to thermal effects or a change in one or more of the voltages applied to the valve, will produce a serious error in the meter reading. The use of negative feedback in the amplifier reduces the gain while the addition of more stages with negative feedback is a difficult design problem.

The Transistor as a D.C. Current Amplifier

A junction type transistor used with the emitter earthed can produce appreciable current amplification up to perhaps 50 times. The input impedance is low, a few thousand ohms, and the output impedance is high. The basic circuit is shown in Fig. 1. Such an amplifier is obviously ideal for amplifying the current from a silicon crystal rectifier; it has the added advantage that the only power required can be supplied by a three-volt battery.

Fig. 1. The basic circuit



* *Engineering Dept., G.P.O.*

Fig. 2. gives a curve of collector current against input current for a junction type transistor, Mullard prototype OC12, used in the circuit of Fig. 1; it is seen that the current gain is about 35 times.

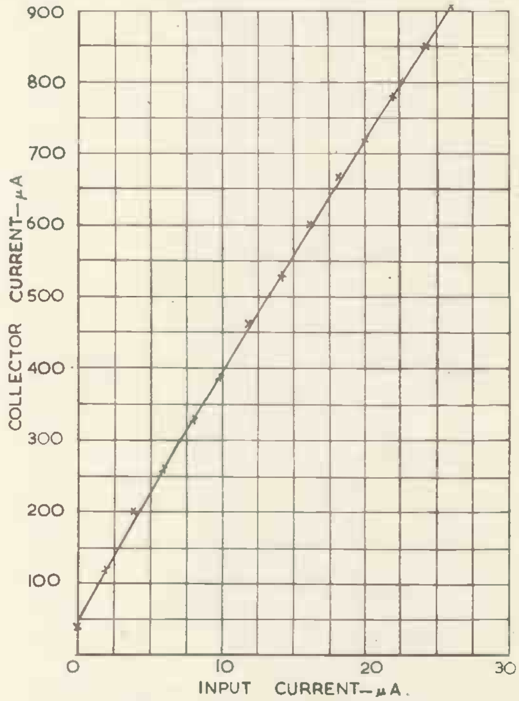


Fig. 2. Collector current/input current for OC12 (emitter earthed)

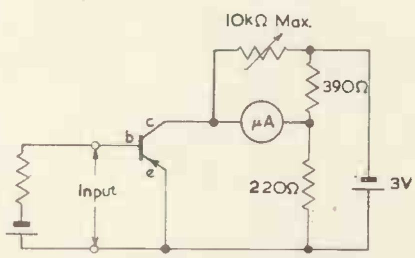


Fig. 3. Backing-off standing current

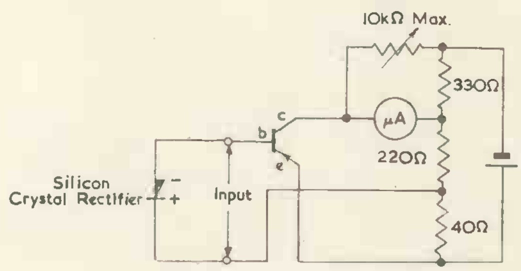


Fig. 4. Method of biasing crystal rectifier

The basic circuit suffers from the defect that with no input current there is a standing collector current. With the transistor used this was about $40\mu A$, but the circuit may be modified as shown in Fig. 3 so that this standing collector current is backed-off.

When the transistor is used to amplify the current from a silicon crystal rectifier the transistor circuit biases the crystal rectifier and reduces its sensitivity. This is readily

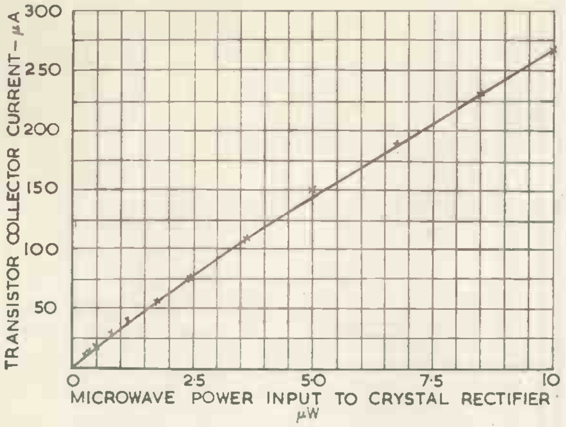


Fig. 5. Collector current/input power to crystal rectifier

overcome by putting a bias in the opposite direction on to the crystal rectifier, an excess of bias improving the crystal sensitivity. Fig. 4 shows how this can be accomplished.

Fig. 5 shows the manner in which the collector current of the transistor varies with the microwave power applied to the crystal rectifier when used in the circuit of Fig. 4. It is seen that the sensitivity of the arrangement is about $30\mu A/\mu W$. The sensitivity of the crystal rectifier used by itself without bias is $0{\cdot}4\mu A/\mu W$; the effect of the transistor and bias on the crystal rectifier is to increase the sensitivity by a factor of 75.

Conclusions

A junction type transistor D.C. amplifier has been described and a circuit given which will improve the sensitivity of a crystal rectifier used for microwave power measurements by a factor of 75. The use of such a circuit enables the galvanometer, usually used for such measurements, to be replaced by a microammeter. As the impedance of the collector circuit of the transistor is high it follows that the resistance of the microammeter can be several hundred ohms enabling a sensitive and robust meter to be used.

Acknowledgment

The writer is indebted to the Engineer-in-Chief, G.P.O. Engineering Department, for permission to publish the above note.

An Improved Reactance Valve Circuit*

The control of the frequency of an oscillator by means of a reactance valve is a practice that is well known. It is also known that the operation of the simple reactance valve circuit may be improved by using two reactance valves connected in series across the oscillator tank circuit. In this arrangement voltages derived from opposite ends of the tank circuit (and therefore 180° out of phase with

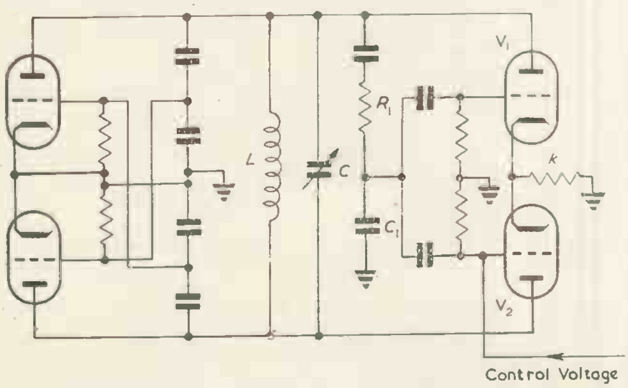


Fig. 1. Method of coupling by common cathode resistor

each other) are applied respectively to the anodes of the reactance valves, and a voltage of 90° out of phase with one of these is applied to both grids. The two valves simulate reactances of opposite kind, that is to say one will appear as an inductance in the circuit and the other as a capacitance. This form of circuit has the advantage that the presence of the reactance valves does not affect the uncontrolled frequency of the oscillator since the simulated reactances produced by the valves cancel out. Other advantages are that the oscillator output voltage remains much more constant with respect to the frequency control voltage and with respect to supply voltage variations. However, in spite of these useful qualities there has been little tendency to use the circuit since the control voltage must be applied to the two valves in opposite phase and in most practical cases this cannot easily be done.

It has been found that this disadvantage may be simply overcome without the need to add unduly to the number of components in the circuit by coupling the two valves together by a common cathode resistor. In this form of connexion any volage applied to the grid of one valve is also transferred in opposite phase to the other valve.

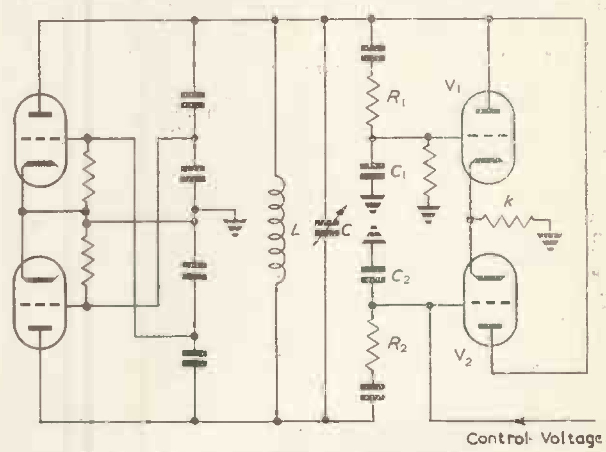


Fig. 2. An alternative arrangement

Two arrangements using the idea are shown in Figs. 1 and 2. In both figures a push-pull form of oscillator is shown in which the tank circuit is denoted by L and C. In Fig. 1 the oscillations are applied in opposite phase to the anodes of the reactance valves V_1 and V_2 which are coupled together by means of the common cathode resistor k, and the network R_1C_1 provides the 90° phase shift for the voltage to be applied to the grids. The frequency control voltage is applied to the grid of valve V_2. In Fig. 2, which shows a simple rearrangement of the circuit of Fig. 1, the anodes of the reactance valves are connected together and in consequence separate networks R_1C_1 and R_2C_2 are employed to provide the quadrature voltages on the grids.

* *Communication from the Telefunken Company, via E.M.I. Ltd.*

TRIDAC

(A Three Dimensional Analogue Computor)

The continuous development of radar and gunnery systems, guided missiles, and high speed aircraft for both civil and military purposes creates a stream of mathematical problems of ever-increasing number and complexity and to solve some of the most difficult of these problems the Ministry of Supply has recently brought into operation at the Royal Aircraft Establishment a new electronic-hydraulic calculating machine or "simulator" called TRIDAC. The name is derived from "Three Dimensional Analogue Computor", and a particular feature is that the machine can deal with problems in which a radar beam or an aeroplane flying under automatic control moves with complete freedom in any direction in space.

TRIDAC, the biggest calculating machine in this country, occupies 6 000 square feet of floor space, uses 8 000 valves, and under peak conditions absorbs about 650kW.

The computor is intended to serve as a research aid for the applied physicist, aerodynamicist and engineer and the basic data available for most of the problems to be studied will rarely be known to better than 1 per cent so that solutions with an error somewhat less than 1 per cent are adequate. In TRIDAC each individual mathematical operation is carried out with an error not greater than 0·1 per cent of full scale. The fact that an analogue machine of the TRIDAC type provides a working model of the system greatly assists the research worker to understand the system and as such it may be regarded as an "aid to physics" or as an "aid to engineering."

The basic analogue quantity in TRIDAC is a direct voltage and the basic element in the majority of the computing units is a high gain D.C. amplifier with input and feedback impedances arranged so as to give an overall transfer function corresponding to addition or integration. To avoid errors due to drift in the amplifiers each amplifier is stabilized by association with a separate drift-stabilizing A.C. amplifier of either the mechanical-chopper or magnetic-modulator type. The gain of the basic D.C. amplifier is of the order of 60 000 and the drift stabilizers can confine the drift to within a few millivolts. To facilitate servicing all the amplifiers and similar units are built in unit form, i.e., on small chassis of approximately 12in by 8in by 3·5in, which slide into arrays of pigeon holes in the cabinets. There are 44 cabinets containing in all about 2 000 units of which approximately 600 are D.C. amplifiers, 350 are mechanical chopper stabilizers and 250 are magnetic modulator stabilizers.

A major section of the computor is a group of nine electrically controlled hydraulic servo motors. Each servo is provided with its own 35 h.p. electric motor and pump delivering oil at 2 000lb/sq.in to the main hydraulic valves. A common low pressure pump supplies oil at 250lb/sq.in to the pilot valves. These high performance servos drive potentiometers. There are about 224 such potentiometers, some of which are directly driven and some of which are driven through sine or cosine linkages. They may be used to multiply voltages appearing in the main electronic computation by the appropriate servo input or by the sine or cosine of the appropriate servo shaft angle, and many of them are used in "axis transformation" calculations.

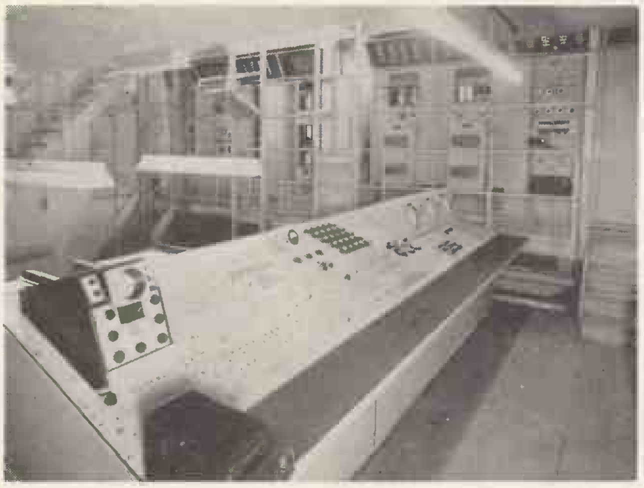
A central generator room housing rotating machinery supplies power at various voltage levels and frequencies to the entire simulator. The supply voltages are accurately stabilized in individual voltage-stabilizing units before being applied to the computing units. A monitoring system is used to detect and give warning of fault conditions in the computing units.

The way in which TRIDAC works may be illustrated by the following examples. Suppose that the forward speed of an aircraft is represented by a voltage on a scale of one volt to ten miles per hour. This voltage, and many others, can be shown on meters, so the speed of the aeroplane in the problem can be observed continuously by watching the meter, and recording devices are provided to give a permanent record when required. The voltage representing forward speed can be fed into an electronic integrator, which produces a second voltage which represents the distance covered by the aeroplane, again according to some pre-arranged scale which might be say, one volt to 100 miles. This voltage also appears on a meter. Alternatively, the voltage representing forward speed might be fed into a resolver into which is also set, either by hand or automatically, the direction in which the aeroplane is flying, say for example 20 degrees north of east. The resolver gives out two more voltages, one representing the rate at which the aeroplane is moving eastward, and the other the rate northward. These two voltages can be fed into two integrators which give output voltages representing distances covered by the aeroplane in the easterly and northerly directions, and by observing these voltages on meters the position of the aeroplane can be plotted on a map. Usually, however, events move too fast for manual plotting and an automatic plotting table is provided. This is a flat board on which runs a carriage moving east and west, and this carriage has a long rail, pointing north and south and carrying a pen which marks a sheet of paper laid on the board. The carriage is moved by the "eastings" voltage so that it travels towards the east by an amount representing the eastward travel of the aeroplane according to the scale of, say, ten miles to the inch. At the same time the pen is travelling along a rail under the influence of the "northings" voltage, so that the pen moves simultaneously east and north, just as the aeroplane does, and the pen traces the path of the aeroplane on the paper. If desired the paper could be replaced by a map of suitable scale. By using a more complicated resolver the variations in height of the aeroplane can be taken into account by feeding in an angle of dive or climb and another plotting table will give a continuous record of the height.

The forward speed of the aeroplane may not be constant, but TRIDAC can still perform the desired calculations no matter how the speed varies, and can indeed go further and calculate the speed from the engine thrust which it in turn calculates from the throttle settings and the engine characteristics, which can be set into the computor by setting a series of controls. These calculations take into account the effects of changing trim of the aeroplane as fuel is used, and the changes of aerodynamic drag and engine efficiency with changing height. Besides varying its forward speed the aeroplane may also vary its direction of flight, and TRIDAC is able to calculate these variations by performing mathematical operations which represent the angles of deflexion of the ailerons, rudder and elevator. These calculations take account of the change of air flow over a wing or control surface when the direction of the flight changes, and also take account of the changes of lift on the wings and other surfaces due to changes of height, speed, and trim of the aeroplane. If the aeroplane is being flown by an auto-pilot the characteristics of this device can be reproduced, and if the aeroplane is using its own radar set to pick up signals from the ground or from another aeroplane the magnitudes of the signals from the radar set can be calculated, allowing for the effects of vibration, interference, etc.

In the case of a fighter aeroplane chasing a bomber the motion of the bomber can be reproduced by setting into the computor a pre-arranged programme, which may include periods of straight flight, slow turns, or violent evasive manœuvres. TRIDAC will then calculate what signals the radar set in the fighter will produce, allowing for the fact that both fighter and bomber are moving at high speed. All the calculations are carried out at the same speed as the actual events occur, so that if the fighter-bomber interception takes ten minutes in actual flight then TRIDAC will do the calculation in ten minutes. While the calculations are proceeding all the quantities in the problem are available as voltages, and can be observed by means of meters, plotting tables, cathode-ray tubes, etc., so that the operator can get an excellent picture of what is going on. He can stop the calculation at any time, and he can repeat as often as he wishes, using different kinds of aircraft by setting different aerodynamic characteristics, different speeds for fighter and bomber, different starting positions, etc., so the chance of an interception in any given circumstances can be found, and the most favourable tactics can be discovered. Similar calculations can be performed for guided missile systems, including beam-riding, homing, and command-link systems. The use of TRIDAC for this purpose will reduce the number of actual missiles which need to be fired for test purposes, with a consequent saving in time and expense.

Since TRIDAC works in the same time scale as the real system, actual equipment, such as the motor which moves the ailerons in a powered-control aeroplane, can be included in the operation of the computor. A voltage in TRIDAC which represents the input signal to the aileron motor is brought out and fed into the actual motor which may move either a real aileron or a dummy. The aileron carries a potentiometer which provides a voltage proportional to the angle through which the



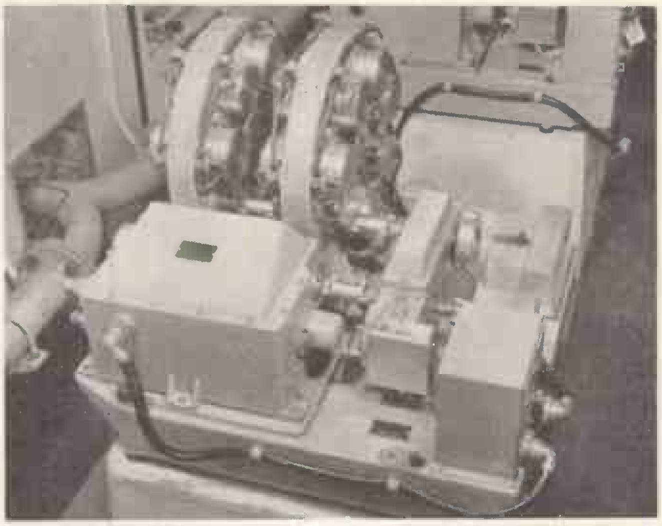
A general view of TRIDAC, showing the control desk with the three-dimensional display at the right-hand end

aileron turns, and this voltage is passed back into the computor for the calculation of rate of roll, etc. This procedure allows measurements to be made of the effect of using different kinds of aileron motors. The pilot of the aeroplane can also be included when required. He is provided with "displays" of radar and instrument information, operated by signal voltages from TRIDAC, and he operates controls of the normal type which drive potentiometers to give another set of voltages to be fed back into TRIDAC.

One of the central parts of the problems which TRIDAC has to handle is that of "axis transformation", which involves complex geometrical and trigonometrical calculations. In the fighter bomber problem the fighter-borne radar dish normally points ahead, but when the bomber comes into range the dish must be turned, say, 20 degrees to the right and then tilted, say, 15 degrees downward. These angles are measured with reference to the frame of the fighter, and if the fighter turns or dives these angles will change, because the auto-follow action of the radar keeps the dish pointing continuously at the bomber. In many TRIDAC calculations it is necessary to know the direction in which the radar dish is pointing, in terms of an angle from the true north and a depression or elevation relative to the true horizontal. To do this TRIDAC takes the angles of the dish as measured with reference to the frame of the fighter, and also takes the two angles which represent the direction in which the nose of the fighter is pointing, i.e. the bearing relative to true north and the angle of climb or dive relative to the true horizontal. This second pair of angles is available from another part of the machine which is calculating the flight path of the fighter. From the two pairs of angles TRIDAC is able to find the direction in which the dish is pointing, referred to true north and horizontal. This calculation proceeds continuously, since the position and attitude of the fighter and the position of the bomber are changing continuously.

Before any calculations can begin on TRIDAC a large

Swash-plate sine-cosine mechanism.



Hydraulic servo driving a number of slave linear potentiometers.

amount of information has to be supplied. In the case of numbers which will not change as the problem proceeds, such as the size of an aeroplane wing, or relation between the rudder deflexion and the rudder-bar deflexion, there are up to 2 000 potentiometers which can be set before calculation begins. For known variations, such as the change of speed of sound with height, specially shaped cams and potentiometers are provided. Adjustable electronic apparatus using sets of biased diodes are used for other non-linear variations.

TRIDAC has been designed jointly by R.A.E. and Elliott Brothers (London) Limited, and it has been constructed by Elliott Brothers and installed at the R.A.E. It will be operated by R.A.E. staff, and will be available to help industry in the same way as wind tunnels and other major items of research equipment.

LETTERS TO THE EDITOR

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

A Count-Rate Meter Circuit

DEAR SIR,—Mr. E. W. Pulsford's article on an ingenious method of correction for dead time losses in count-rate meters (page 356, August issue) omits a point which seems to extend the application and utility of the system. As is well known, the diode pump and D.C. amplifier metering system is not absolutely linear owing to the finite amplifier gain. Using Mr. Pulsford's symbols the signal voltage at the anode of V_8 in the absence of compensation can be expressed in the form

$$v = AVn \cdot \frac{R_t C_f/(\beta A + 1)}{1 + nR_t C_f/(\beta A + 1)} \quad \cdots (1)$$

where A is the amplifier gain without feedback and βA is the loop gain across R_t. It is of interest to note that, for given values of R_t, V, A and maximum output voltage, the non-linearity expressed by this equation is independent of β, since C_f must be made proportional to $(\beta A + 1)$. If, while maintaining the full feedback connexion of C_f, the right-hand end of R_t is earthed, β is then zero and the circuit becomes a type of Miller integrator; this form, however, suffers from poor stability of gain and zero.

Writing $T = R_t C_f/(\beta A + 1)$ and $n = N/(1 + Nt)$ we have

$$v = AVN \cdot \frac{T}{1 + N(t + T)} \quad \ldots\ldots (2)$$

This equation shows that the inherent non-linearity of the diode pump and D.C. amplifier system is equivalent to increasing the dead time t by the quantity T. Mr. Pulsford's circuit can therefore be adjusted to eliminate this source of inaccuracy without any modification.

The percentage departure from non-linearity, due to both dead time and diode pump, is approximately $-100N(t+T)$ per cent. In the circuit described T is probably about $0{\cdot}1t$. The fact that both sources of inaccuracy can be exactly eliminated by the same circuit allows additional freedom in the design of the diode pump metering system.

Yours faithfully,

B. D. CORBETT,
Dept. of Clinical Research,
University College Hospital Medical School.

The Author replies:

DEAR SIR,—I am indebted to Mr. B. D. Corbett for pointing out that the circuit of the count-rate meter which corrects for dead-time losses, is also capable of correcting the inherent non-linearity of the scale, and I am in full agreement with his analysis.

There is a simple way of calculating the linearizing voltage to be fed back to the diode pump driving wave. When the signal voltage at the anode of V_5 is v, the movement of the input grid from its initial voltage is $-v/A$, where A is the gain of V_5, without feedback. This grid voltage shift is effectively subtracted from the amplitude of the driving wave, and so to maintain the effective amplitude constant, v/A must be added to it, which can be done by the appropriate adjustment of the circuit, as Mr. Corbett notes.

It is doubtful if it is worth the extra complication in a simple count-rate meter, but if a dead-time compensating instrument is being set up, it might be worth going to the trouble of setting in this linearizing feedback before proceeding to adjust for the dead-time compensation. In the experimental model the departure from linearity, as read on a first grade meter, was quite small, and this is the common experience with this type of rate-measuring circuit.

A further extension of the utility of this system is believed to exist, although the analysis is not yet complete. In the counting of random-in-time pulses with a count-rate meter not having any intentional resolving time, some losses always occur because of the incomplete charge or discharge of the diode pump feed capacitor when the pulse spacings are of the same order of magnitude as the charging time-constant (C_1R_4). Our analysis so far leads us to believe that the instrument behaves as though it has an effective resolving time equal to this time-constant, although none is intentionally provided. In the cases where a resolving time is provided it is lengthened by this time-constant, and therefore losses from both these causes may be compensated by the method of the article. (Reverting to the instrument described, having an intentional resolving time of 10μsec, and a charging time of 2μsec, it might be more accurate to adjust for an effective resolving time of 12μsec). Should this prove to be the case, then more accurate count-rate meters for dealing with the very fast count-rates from scintillation counters could be designed economically. It should be stressed that the remarks of this paragraph refer only to the counting of random-in-time pulses.

In conclusion. I must apologize for a small error in Fig. 3, page 357, where a high potential terminal of R_4 should be connected to the cathode of V_3, and not to the junction of R_6 and R_7.

Yours faithfully,

E. W. PULSFORD,
A.E.R.E., Harwell.

Output Impedance of Anode-Follower Type Circuits

DEAR SIR,—I find that some text books quote the output impedance of the circuit of Fig. 1 as being $Z_o = 2/g_m$ or to be some value which to a first approximation does not include R. Below is a simple but complete analysis which shows the output impedance to be proportional to R.

Let $\beta = \dfrac{R_1}{R_1 + R_2}$

Let A be the gain of the circuit from the grid to the anode when the output

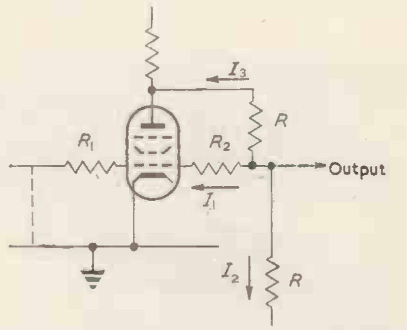


Fig. 1. Circuit for calculating output impedance of anode-follower

point is shorted to earth and R between the anode and output is replaced by $\dfrac{A\beta \cdot R}{A\beta + 1}$ (which is approximately equal to R).

If the output voltage is changed by ΔV the current will be split between R_2 and the two R's, I_1 and I_2 are given by $\dfrac{\Delta V}{R_1 + R_2}$ and $\Delta V/R$ respectively.

The current I_3 is given by:

$$\Delta V/R + \frac{AR_1}{(R_1 + R_2)} \ \Delta V/R$$

adding the currents and dividing by ΔV gives:

$$1/Z_o = \frac{1}{R_1 + R_2} + 1/R + 1/R + \frac{AR_1}{(R_1 + R_2)R}$$

giving:

$$Z_o = \frac{RR_1}{\beta R + R_1(2 + A\beta)}$$

for all practical purposes. βR can be neglected compared with $R_1(2 + A\beta)$ giving:

$$Z_o \simeq \frac{R}{2 + A\beta}$$

and again $A\beta$ is always very much greater than 2, giving:

$$Z_o \simeq R/A\beta$$

This result does not agree with that very often quoted.

Yours faithfully,

D. McDonnell,
Weybridge, Surrey.

The Design of High Efficiency Radio Frequency E.H.T. Supplies

Dear Sir,—In view of the correspondence upon this subject published in the November issue, the author feels that a detailed calculation of one of the original examples given in the article would clarify the design process considerably. In the simple case where E.H.T. only is required, and no heaters have to be supplied, the calculations are perfectly straightforward. To indicate the influence of other windings upon the design, a detailed calculation of example 2 in its corrected form follows:

The required supplies are 6·6V D.C. at 2mA, 4V 2A for C.R.T. heaters, and two floating supplies of 300V D.C. at 10mA.

Here the A.C. load on the E.H.T. coil due to the D.C. load is $6{\cdot}6/4 = 1{\cdot}65\text{M}\Omega$. This would indicate the choice, from Table 2 of the 1 600 turn coil, but in the present case the loads on the other windings are reflected across the E.H.T. coil by the transformer, and reduce the Q factor as seen by the oscillator valve. These reflected loads lead to a choice of the 800 turn coil, when the reflected A.C. loads can be found by means of voltage ratios.

Using the 800 turn coil we have volts per turn $= \dfrac{6\,600}{800\sqrt{2}}$ R.M.S. $= 5{\cdot}82$. Thus the two heater windings will each have 1 turn. In order to produce 4V for the C.R.T.s, a small auto-transformer wound on a miniature Ferroxcube core was used, to avoid undue power loss. For the rectifier heater, a dropping resistor was used, the valve requiring 1·4V at 0·14A.

The reflected loads are then as follows:

From rectifier heater $\dfrac{5{\cdot}82}{0{\cdot}14}\left(\dfrac{6\,600}{5{\cdot}8\sqrt{2}}\right)^2 \times 10^{-6} = 26{\cdot}7\text{M}\Omega$

From C.R.T. heater $4/2\left(\dfrac{6\,600}{4\sqrt{2}}\right)^2 \times 10^{-6} = 2{\cdot}72\text{M}\Omega$

From each 300V winding $15\,000\left(\dfrac{6\,600}{300}\right)^2 \times 10^{-6} = 7{\cdot}26\text{M}\Omega$

The effective A.C. shunt is therefore $0{\cdot}777\text{M}\Omega$.

Thus with the 800 turn coil we have

$$Q = \frac{0{\cdot}777}{0{\cdot}069} = 11{\cdot}25,\quad \eta = \frac{100 \times 43{\cdot}3}{43{\cdot}3 + 1{\cdot}65} = 96{\cdot}5 \text{ per cent.}$$

The remaining data originally given for this example are correct.

Yours faithfully,

J. Barron,
University of Cambridge,
Department of Physics.

Multi-stage Amplifier Output Impedance

Dear Sir,—The ratio of the net amplification to the parallel output impedance of a simple amplifier, both with and without voltage feedback is a constant. When this result is applied to the case of a multi-stage amplifier with over-all voltage feedback, a simple expression relating the parallel output impedance to the amplifier constants is obtained.

The stage gain A, of a simple amplifier can be expressed by the relationship: $A = \mu Z_L/(r_a + Z_L)$ where μ and r_a refer to the valve constants, and Z_L is the load impedance. As the mutual conductance of the valve $g_m = \mu/r_a$ an alternative expression can be obtained which states that $A = g_m r_a Z_L/(r_a + Z_L)$. By definition the parallel output impedance $Z_p = r_a Z_L/(r_a + Z_L)$ so that:

$$A/Z_p = g_m \quad \ldots\ldots\ldots\ldots\ (1)$$

(Parallel output impedance Z_p is used to mean the impedance between the output terminals with the load still connected. The normal output impedance Z_o is always measured with the load disconnected.

The relationship connecting the two output impedances is obviously:

$$1/Z_p = 1/Z_o + 1/Z_L$$

where Z_L is the load impedance.

N.B.—The expressions developed hold only for voltage feedback and not for circuits using current feedback.)

When feedback is employed, so that a fraction β of the output voltage is returned to the input, the net gain:

$$A' = A/(1 - \beta A) \quad \ldots\ldots\ (2)$$

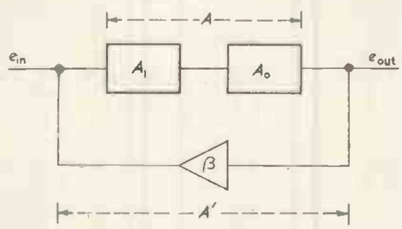


Fig. 1. Multi-stage amplifier with voltage feedback

and the new parallel output impedance:

$$Z_p' = Z_p/(1 - \beta A) \quad \ldots\ldots\ldots\ (3)$$

so that dividing equation (2) by equation (3) we have:

$$A'/Z_p' = A/Z_p = g_m \quad \ldots\ldots\ldots\ (4)$$

Hence the ratio of the net gain to the parallel output impedance is equal to the mutual conductance of the valve. Equations (2) and (3) are dependent upon the nature of the feedback, and β will be negative for negative voltage feedback, but the term involving β does not appear in equation (4), so that this result is independent of the nature of the feedback employed.

The result becomes of interest when applied to a multi-stage amplifier. If we split the amplifier into two sections; one, from the input to the grid of the last stage, having a gain of A_1, and the other, the output stage, having a gain of A_o, then the total amplification $A = A_1A_o$ and the parallel output impedance $Z_p = A_o/g_m$ where g_m applies to the last stage only.

If a voltage feedback is connected, as shown in Fig. 1, the total gain of the amplifier is now A', and by comparison with equation (2) and (3) it can be seen that:

$$A' = A/(1-\beta A) = A_1A_o/(1-\beta A_1A_o) \quad \ldots\ldots\ldots\ldots\ (5)$$

and that the new parallel output impedance is:

$$Z_p' = Z_p/(1-\beta A) = Z_p/(1-\beta A_1A_o) \quad \ldots\ldots\ldots\ldots\ (6)$$

Dividing equation (6) by equation (7) it can be seen that $Z_p'/A' = \dfrac{Z_p}{(A_1A_o)}$ and since in this case $Z_p/A_o = 1/g_m$, if g_m is the mutual conductance of the last stage, therefore $Z_p'/A' = 1/g_m \,.\, 1/A_1$ or the parallel output impedance:

$$Z_p' = 1/g_m \,.\, A'/A_1 \quad \ldots\ldots\ (7)$$

In the more useful case where the voltage feedback is negative, A' is constant if the relationship $\beta A_1A_o \gg 1$, which means that the parallel output impedance of the multi-stage amplifier will then be independent of the load in the last stage, as A_o does not enter into the expression.

If the multi-stage amplifier is reduced to a single stage, then $A_1 = 1$ and the parallel output impedance reduces to the value for the simple case in equation (4).

A simple example is the cathode-follower which has a gain very nearly equal to 1, and by means of equation (7) the parallel output impedance can be seen immediately to be $1/g_m$ which is, of course, the well-known value.

More precisely, the gain is equal to $g_m/(1/R_k + 1/r_a + g_m)$, where R_k is the cathode load, and the parallel output impedance can quickly be recognized as $1/(1/R_k + 1/r_a + g_m)$.

The design conditions for an amplifier with a gain of 1 000 and a parallel output impedance of 1 ohm can be easily calculated from equation (7) by treating A_1 as the unknown. If we chose a valve with a mutual conductance of 5mA/V for the last stage then:

$$A_1 = 1/g \,.\, A'/Z_p' = \frac{1\,000}{5 \times 10^{-3}} = 2 \times 10^5$$

and

$$\beta = -1/A' = \frac{-1}{1\,000} = -10^{-3}$$

The condition that $\beta A_1A_o \gg 1$ is satisfied if $A_o > 1/20$, so that the load of the last stage can have almost any value. In practice it would be chosen so that the current flowing through the valve gave the desired value of g_m.

Reduction of the output impedance by these means only reaches a limit when the amplifier becomes unstable due to phase changes in the feedback network.

Yours faithfully,

J. B. Earnshaw,
Physics Department,
Auckland University College,
New Zealand.

ELECTRONIC EQUIPMENT

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

Strobodynamic Balancing Machine

(*Illustrated below*)

TWO models of a balance testing machine of special interest to the electric motor, aircraft and motor car industries are announced by E.M.I. Engineering Development Ltd.

The equipment is simple and positive in use. The rotor to be balanced is supported in the open bearings of two freely-suspended cradles and belt driven by an electric motor. By means of special circuits, the unbalance produces an electrical signal which is arranged to flash a stroboscopic lamp which illuminates the workpiece under test. It thus appears stationary, with the position of the unbalance indicated against a pointer;

at the same time, the signal actuates a meter to indicate the amount of unbalance.

The Type 1 machine takes loads up to 30lb, with a maximum diameter of 11in and maximum length between bearings of 14in. Balancing speed range is 1 000 to 2 500rev/min.

The Type 2 (illustrated) machine has a similar speed range, but the other maxima are: weight of load 7lb; diameter 4in; length between bearings $8\frac{1}{4}$in.

E.M.I. Engineering Development Ltd,
Wells,
Somerset.

Silvered Mica Capacitors

(*Illustrated above right*)

INTENDED primarily for applications where long-term stability is essential, the Johnson Matthey range of new design capacitors takes advantage of increased electrode areas and lower voltage rating to bring about a reduction in both bulk and cost.

These capacitors are rated at either 200 or 350V peak working and are of fired construction, thus completely obviating the use of eyelets. They are impregnated with a wax suitable for use over the temperature range −30 to +70°C and a coating of the same wax is used to protect them.

Three overall sizes from 1 11/16in by $1\frac{1}{8}$in to 1 1/16in by 9/16in are available and for each size there are two possible lead arrangements (see illustration). Any value of capacitance in the range 300pF to 0·25μF can be supplied with a minimum tolerance of $\pm\frac{1}{2}$ per cent.

Johnson, Matthey and Co, Ltd,
78 Hatton Garden,
London, E.C.1.

Quadrant Meter

(*Illustrated below*)

A RECENT addition to the range of electrical measuring instruments manufactured by Measuring Instruments (Pullin) Ltd. is the quadrant meter. It is ideal for use where a long scale instrument is required on equipment having a very limited interior space available as it occupies the same space behind the panel as a conventional meter with a 2in scale but, with a front projection less than 1in, has a scale length of 7in.

The dial which can be supplied in various colours to customers specification is made of a special translucent material and is unbacked. This, together with the positioning of the movement in the corner of the dial facilitates illumination from the back of the instrument and results in a uniform spread of light over the whole scale.

The quadrant meter is available as a milliammeter or ammeter with any range from one milliamp upwards or as a voltmeter and can be supplied with a moving-coil movement for D.C. circuits

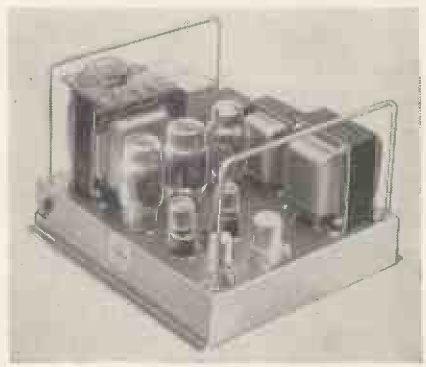
or a rectifier can be incorporated for use on A.C. This meter complies in all respects with B.S.89:1954 for industrial grade instruments.

Measuring Instruments (Pullin) Ltd,
Electrin Works,
Winchester Street,
Acton,
London, W.3.

Wideband Audio Amplifier

(*Illustrated below*)

AN engineered version of the Williamson amplifier is now being produced by Radford Electronics. In addition to its normal audio uses this amplifier is suitable for use on low frequency vibration and ultrasonic research. The frequency response is



claimed to be flat within ±2db from 5c/s to 300kc/s and the distortion less than 0·1 per cent at 15W output. The standard output transformer provides output impedances of 1·7, 6·8, 15·3, 27, 42·5, 61, 83 and 109Ω, but other impedances can be supplied to order.

Radford Electronics Ltd,
149 Newfoundland Road,
Bristol, 2.

Miniature and Sub-miniature Precision Potentiometers

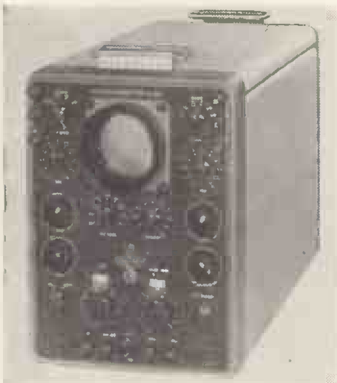
TWO new precision potentiometers have been added to the range recently introduced by Salford Electrical Instruments Ltd. They comprise linear miniature and sub-miniature types, and have been developed for use in applications where space is particularly limited.

The miniature potentiometer, the D.2, though it weighs 15g and is only $1\frac{1}{8}$in in diameter, covers a resistance range from 1kΩ to 90kΩ and has a power rating of 1W. The maximum resolution obtained is 6 turns/degree and the standard electrical angle is 340°. There are two taps on the standard model, each being made to a single turn of wire. The housing is fitted with a keyed locating spigot and this allows up to three units to be ganged on a single shaft. The torque with plain bearings is approx. 5g.cm, but ball bearings can be fitted if necessary and the torque is then of the order of 2 to 3g.cm.

The sub-miniature potentiometer, known as the S.D., features a special resistance card and clamp ring assembly. The complete potentiometer weighs only 4g and measures $\frac{5}{8}$in in diameter by $\frac{1}{2}$in in length; the latter measurement is taken from the front of the case to the rear of the housing and does not include terminal or spindle projection.

The resistance range covered by this instrument is from 200Ω to 8kΩ. The maximum resolution is 2 turns/degree and the power rating $\frac{1}{2}$W. There are two taps and the electrical angle is 280°. The sub-miniature potentiometer cannot be ganged, and is essentially a "trimmer" and unsuitable for servo application. It can, however, be supplied with a $\frac{1}{2}$in long shaft or with a screw adjustment as required.

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



Servo and General Purpose Oscilloscope

(Illustrated above)

THE model 0·100 oscilloscope has been developed from the type 1684D/2 and retains its main features, viz: symmetrical, direct-coupled, X and Y amplifiers (zero to 4Mc/s), instantaneous shift with high datum stability and functional independence of controls. In addition, the 0·100 embodies a triggered time-base, together with calibrated time and voltage measurement. It is also suitable for photographic recording as well as direct observation of both transient and recurrent phenomena. Cameras suitable for single shot or continuous film trace photography are available.

Furzehill Laboratories Ltd,
Shenley Road,
Boreham Wood,
Hertfordshire.

High Q Inductors

(Illustrated above right)

THE Components Division of Mullard Ltd. can now supply a wide range of ready-wound high Q inductors. Most of these are on Ferroxcube pot core assemblies and are suitable for use at frequencies from 300c/s to about 500kc/s. In due course, additional types will become available which will extend both the frequency and inductance ranges.

Owing to the low material hysteresis loss, it is possible, with very high values of inductance, to obtain extremely high values of Q, for example, a Q of 700 at 10kc/s.

In addition to high efficiency, coils on Ferroxcube pot cores have the advantage of compactness. A 20H inductor may be wound on a core with overall dimensions of 1 1/16in by 1 1/16in by $\frac{3}{4}$in, including all necessary clamping screws.

The inherent self-screening effect of Ferroxcube, due to its high material permeability, makes it possible to stack coils or place them near other components without danger of stray inductive coupling. This property makes them particularly useful for compact, efficient filters.

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

Differential A.C. Ammeter

THIS instrument is sensitive enough to indicate a difference of 1A between two A.C. circuits each carrying 1 000A or 5μA between two circuits each carrying 5mA.

It consists of a centre zero moving-coil instrument and two bridge rectifiers, the output of these being in opposition and connected to the moving-coil instrument. Usually each rectifier is connected with its A.C. side to the output of a transformer, the transformers being well balanced and identical in every characteristic.

The Electrical Instrument Co. (Hillingdon) Ltd,
Boswell Square Industrial Estate,
Hillingdon, Glasgow, S.W.2.

Klystron Power Supply Unit

(Illustrated below)

THIS power supply unit has been developed to furnish all the voltages necessary for the operation of equipment containing klystrons under development conditions. Each supply is fully and independently metered, and all are

insulated for use at a minimum potential of 2·5kV to earth, positive or negative. The following voltages are available, with reference to the klystron cathode :—

E.H.T. Adjustable from 1kV to 2·5kV, at from zero to 15mA. Internal resistance not exceeding 500Ω, and regulation for changes of 10 per cent in mains supply voltage not exceeding 0·01 per cent.

Grid Bias Supply. Adjustable from zero to 250V at 100μA maximum.

Reflector Supply. Adjustable from zero to 500V at 1mA maximum. Stability and ripple percentage as for H.T. supply.

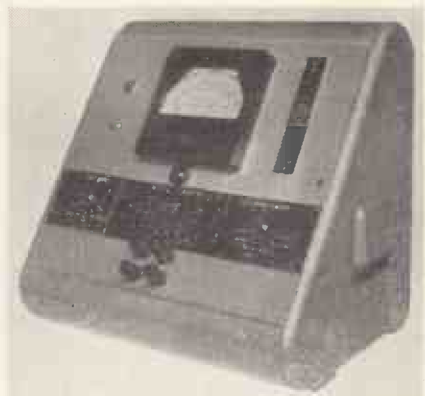
Heater Supply. Adjustable from zero to 8V A.C., at 2A maximum.

Radio Aid Ltd,
29 Market Street,
Watford,
Hertfordshire.

Electronic Multi-range Meter

(Illustrated below)

THE Clare Instrument Company's wide range linear scale ohm meters are now available with built-in ranges



for voltage and current employing a stable cathode-follower indicator system. Current ranges down to 0·3μA and voltage ranges up to 30kV at 3·3MΩ/V full-scale can be provided. A typical instrument would cover resistance ranges from 3Ω to 1 000MΩ, current from 0·3μA to 300mA D.C. and voltage from 3V to 300V D.C., F.S.D. Additional A.C. and D.C. ranges can be supplied to order.

Clare Instrument Co,
Rickmansworth,
Herts.

100db Step Attenuator

THE TF1073 is a 75Ω R.F. attenuator variable in 1db steps from 0 to 100db. Using the full 100db at 100Mc/s the maximum error does not exceed 0·6db. It consists of two sections in cascade, the first has an attenuation range of 80db in 20db steps and the second has a range of 20db in 1db steps. Input power up to a maximum of 0·25W may be applied to the attenuator (approx 4V R.M.S. at 75Ω).

The complete unit is housed in a metal case measuring 11in high by $7\frac{1}{2}$in wide by 7in deep.

Marconi Instruments Ltd,
Longacres,
St. Albans,
Hertfordshire.

BOOK REVIEWS

The Present State of Physics

Edited by Frederick S. Brackett. 265 pp. 60 figs. Demy 8vo. American Association for the Advancement of Science. 1954. Price $5.

THIS volume embodies a symposium presented on 30 December 1949 at the New York meeting of the American Association for the Advancement of Science. Despite its title—though understandably—it deals with only a selection of recent developments in four main subjects; those of elementary particles, physics of the solid state, chemical physics and biophysics. Any writer under such a broad title has to choose the extent to which he will sacrifice thoroughness to comprehensiveness. Most of the present authors have evidently decided in favour of the thorough discussion of relatively narrow but typical problems, though J. C. Street's paper on Developments in Cosmic Radiation, 1945-1950 and Karl Lark-Horowitz's on The New Electronics in particular combine breadth of purview with considerable detail of treatment. The book should therefore be read more as an authoritative and representative sample than as a balanced general report of current progress in physics.

Undoubtedly of greatest interest to most readers of this journal will be the three papers on physics of the solid state, which deal almost exclusively with semi-conductors and ferroelectrics. Lark-Horowitz's painstaking survey already mentioned occupies 71 pages and provides no fewer than 352 references to the literature on semiconductors and related topics, some of course pre-dating by many years the current spate of papers on germanium.

From a relatively brief but excellent historical survey covering the initial discoveries of non-metallic conduction, Hall effect, thermoelectric effects, contact rectification, and photoelectric effects, the author proceeds to a highly detailed discussion of the electrical properties of germanium in bulk form. A clear account is given of the effect of chemical impurities and lattice defects, and of current methods of producing these to a controlled extent. A good deal of space is devoted to theoretical attempts to correlate experimental findings, but the tentative and exploratory nature of some of the theory is made clear even to an outsider.

J. Bardeen follows with a lucid discussion of the flow of electrons and "holes" in semiconductors. Designers of transistor circuits will find much of value here, particularly towards an appreciation of the limitations of transistors. Holes in n-type germanium, for example, are reported to have lifetimes sometimes as long as $1\,000\mu$sec. Hole mobility in germanium, 1 700cm/sec per volt/cm, is roughly half that of electrons, and mobilities of both holes and electrons in most other semiconductors are lower. The fundamental limitations on the frequency response of transistors are thus very different from those on thermionic devices.

The paper by von Hippel on barium titanate ferroelectrics is refreshingly written and exciting in its implications. Only three groups of substances are so far known to show electrical remanence effects analogous to those of ferromagnetism, and of these only barium titanate offers interesting properties in an easily processed material over a suitable temperature range. Relative dielectric constants (ϵ'/ϵ_0) as high as 10 000 have been observed near the critical temperature (about 120°C). The value of ϵ'/ϵ_0 depends strongly on temperature and applied D.C. field strength, dropping by a factor of 5 over the ranges 120-20°C and 0-2·5MV/m. Technical applications of these properties are mentioned but not discussed.

In the section on elementary particles, P. Kusch reports the first evidence for the theoretical prediction that an electron interacting with a quantized radiation field should show a magnetic moment higher (by $1/(2\pi \times 137)$) than that predicted from Dirac's classical theory neglecting the interaction. An accuracy better than 1 part in 10^6 is claimed for the atomic-hydrogen-beam resonance measurements from which the calculations were made. Edward P. Ney and J. C. Street in two papers on cosmic radiation give a clear picture of progress in this fertile breeding-ground of new elementary particles.

The sections on chemical physics and biophysics are interesting if only to show how wide a range of topics now comes under the name of physics. The contributions range from P. J. W. Debye's on the Structure of Polymers and Lumry and Eyring's on Chemical Kinetics of Biological Systems, to those of F. Brink, Jr., on Conduction of Nerve Impulses and F. H. Johnson on Bioluminescence.

The reader may be intrigued by an arresting contrast in tone between the second and third of these papers. Lumry and Eyring begin: "A conceptual basis for the understanding of the living organism is probably complete at the present time". Brink concludes his review of the neurophysiologist's theoretical goal by remarking more modestly, "We are far from this desirable state of affairs". There is, of course, no necessary contradiction between the two statements . . . Both papers contain most illuminating information on the properties of the nerve fibre as a communication channel.

The volume, as a whole, is well produced and remarkably free from misprints. It contains a useful if rather slender index. In the circumstances of its origin one can scarcely complain of omissions, but the mention of bioluminescence draws attention to the absence of electroluminescence from the topics covered in the solid state symposium. These and other omissions would be a blemish in any volume having this title if written by a single author. But as a stimulating selection from current work, the present symposium is well worth reading.

D. M. MacKay.

Electronics

By A. T. Starr. 395 pp., 64 figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1954. Price 32s. 6d.

THIS book is one of the engineering degrees series produced by the publishers and was specially written to cover the syllabus for the subject of electronics in the University of London.

The field of electronics is now so large and varied that it is obviously impossible to cover it all thoroughly in a book of this size and price. That it is done so well indicates the familiarity of the author with his subject and his economy of words and skill in technical writing.

Chapter I deals with physical fundamentals in a most interesting way, emphasizing wave mechanics and the quantum theory as explaining more completely the behaviour of matter than the classical and simple picture of the electrons revolving round the nucleus like planets around the sun. The author is therefore able to bring into this chapter the conditions for thermionic emission, cathode ray tubes, photo electric emission, contact potential, and, of great interest now, the behaviour of semi-conductors. The operation of transistors and their main characteristics are neatly covered.

Chapters II and III are devoted to thermionic valves of many types and include cathode ray tubes and camera tubes. Rectifiers and rectification circuits are particularly well described. Sometimes valves are referred to as tubes and it is a pity that a uniform style of drawing is not adopted throughout the book. In many places "envelopes" are not included around the electrodes and some circuit diagrams are thus made more difficult to understand.

Chapter IV covers circuit theory starting with some of the theorems, laws, transformations and networks. Steady state conditions, unit step and impulse functions are then treated. Filters are particularly well covered as would be expected from the author.

Amplifiers, oscillators and detectors are described in Chapter V. Positive and negative feedback is discussed and the effects of each. Multivibrators and time-base circuits are also covered. Waveforms at the various parts of the circuits are shown in most cases and this is always a help to the understanding of circuit behaviour.

Electronic applications are shown in Chapter VI. Servo mechanisms and motor controlled equipment are described in some detail and a number of devices used commercially such as welder control circuit, dielectric heating, and fluorescent lighting. A circuit and description of a photographic timer is included but the circuit has a resistor missing and another incorrectly connected and could not work in the form shown. The circuit was described by the reviewer in the March, 1950, issue of ELECTRONIC ENGINEERING.

The book finishes with several appendices, mainly of a mathematical nature such as theorems and functions, steady state A.C. theory including complex numbers and vectors and passes on to Fourier analysis, and impulse and unit step functions.

The appendix concludes with operational calculus forms due to Heaviside and Laplace and the relation between them, the M.K.S. system, and a useful summary of noise in circuits.

The book has a great number of questions and answers throughout and considering the enormous field it covers does so in a most workmanlike way.

C. H. BANTHORPE.

Power System Transients

Edited by E. Openshaw Taylor. 176 pp., 70 figs. Demy 8vo. George Newnes Ltd. 1954. Price 21s.

THIS symposium, based on a series of lectures delivered at the Electrical Engineering Department of the Heriot-Watt College, Edinburgh, brings together in a convenient form the latest available information on the various aspects of the subject.

Power system engineers, transmission engineers and designers of transformers, switchgear and protective gear will find this work useful for purposes of reference. Students intending to specialize in electrical machinery design, power generation or transmission and distribution will find the book provides a valuable complement to their standard text-books.

A Text Book of Radar

Edited by E. G. Bowen. 617 pp., 300 figs. Royal 8vo. 2nd Edition. Cambridge University Press. 1954. Price 45s.

THE authors of this book, which was first published in 1947, are members of the staff of the Radiophysics Laboratory in Australia and the main emphasis of the book is on the work done there. At the same time a very balanced account is given of the British and American work and the result is a comprehensive survey of radar techniques.

The main portion of the book, consisting of an introductory chapter, one on the fundamental principles of radar operation and a number, each dealing with a particular component of a radar system, have been revised from the first edition. These chapters are each written by the radiophysics laboratory worker, specifically responsible for the component being discussed, so that the book is unquestionably an authoritative one. The level of the treatment is, however, somewhat variable, ranging from the section on cavity resonators, which is suitable for a student, to that on transmission line technique, which is more appropriate to someone already working in this field. On the whole, the second approach predominates and the principal value of the book is as a general reference for those working on radar design.

The concluding chapters, which have been completely rewritten, integrate the remainder by discussing complete radar systems. Military radars are discussed historically and the way in which the development progressed up to the end of the late war is concisely covered. The peace time uses of radar as an aid to air and marine navigation have led to the development of other types of system and these are also discussed. The final chapter is a most interesting one, in which an account is given of the advances resulting from the application of radar techniques to fundamental research. The topics discussed in this chapter include microwave spectroscopy, radio astronomy and linear accelerators.

J. BROWN.

Television Engineers' Pocket Book

Edited by E. Molloy and J. P. Hawker. 228 pp., 80 figs. George Newnes Ltd. 1954. Price 10s. 6d.

THIS book has been specially prepared to meet the growing needs of engineers, dealers, servicemen and television enthusiasts for a handy compendium of practical information and technical data. In addition to useful reference information on cathode-ray tubes, valves, and the leading particulars of more than 300 popular models, the book contains valuable guidance on installing, servicing and aligning receivers.

Practical Television Circuits

By F. J. Camm. 288 pp., 156 figs. Demy 8vo. George Newnes Ltd. 1954. Price 15s.

THIS book gives constructional details, with wiring diagrams, of a number of television receivers which have been described in *Practical Television*.

The contents also include details and diagrams for a spot wobbler, a black spotter, E.H.T. generators, a pattern generator, preamplifiers, etc.

Electron Optics

By O. Klemperer. 471 pp., 120 figs. Demy 8vo. 2nd Edition. Cambridge University Press, 1953. Price 50s.

AMONG the many good textbooks on electron optics this work, perhaps more than any other, bears the individual stamp of its author, a distinguished electron physicist of 28 years' standing, with almost equally long industrial as academic experience. His own work has a deservedly prominent place in almost every chapter, and much of the mass of useful data collected in this book is the result of his own indefatigable researches.

The science and technique of electron optics has now grown to such an unmanageable size, that a book like the present, of 437 text pages cannot give even comprehensive abstracts of the majority of the important papers, but has to refer to them often in a few lines only. The research worker who wants to consult this book with a view to learning something in his own special field, must be content with finding in it an unusually complete enumeration of the points at which special fields of research branch off the main body of electron optics. In almost all cases he will have to complete his study by referring to the original publications. This cannot be helped, else the book would have become an encyclopedia.

Dr. Klemperer's book can be compared to a road map, in which all the highways are traced, and all cross roads are neatly and completely labelled with signposts, without tracing all the way the second and third class roads. This is what gives the work its special and useful character, even at this late hour when it has to compete with so many other textbooks. Compared for instance with Walter Glaser's Elektronenoptik, which is almost twice its size and four times its price, this book gives only a cursory introduction to theoretical electron optics, but it opens up many times more vistas into practical applications.

Some of the chapters deserve special mention, such as those on space charge, on emission systems, and on line focus systems, as in these fields the author's own work, published here for the first time in many instances, is of first class importance.

The first two chapters deal briefly with the fundamentals of electron optics and with the cardinal points of electron lenses. The third chapter is a comprehensive and very useful survey of almost all practical ray tracing methods. Chapters IV and V contain in a condensed form much useful reference material on the cardinal data of electric and magnetic lenses. Chapter VI, on the geometrical aberrations, represents a brave attempt to condense into 48 pages a subject of extraordinary intricacy and complexity. The next chapter which bears the somewhat unusual title "Electronic Aberrations" is chiefly devoted to the chromatic aberrations, their avoidance in lenses and their utilisation in spectrometers. The last is a subject in which the author is a pioneer; it was he who devised, in 1933, the first lens spectrometer. The diffraction errors receive only cursory treatment. Chapters IX and X, on emission systems and on line focus systems will be of special value to the designer. This too is a field in which the author's contributions are among the most important. The chapter on deflecting fields, which includes spectrometers as well as television tubes is of the "first-class highways and signposts" type, while the last chapter on the applications of electron optics in industry and research consists mainly of signposts. This last is a field in which there is no highway; it is still more or less free for the inventive imagination which does not run on rails.

It would be unkind to point out the few errors, and the few omissions of important references in a book which represents such an enormous amount of careful work, and which contains about 600 bibliographic entries, all digested.

The readers to whom this book will be useful and even indispensable are in the first line all electronic engineers and physicists engaged in research or development, and those students who do not only want to pass their examinations, but wish to prepare themselves seriously for a career in electronics.

D. GABOR.

Transistoren—Wirkungsweise, Eigenschaften und Anwendungen (Transistors—Mode of Action, Properties and Applications)

By M. J. O. Strutt. 166 pp., 121 figs. Demy 8vo. S. Hirzel Verlag, Zurich. 1954. Price S.Fr.21.

THE book deals systematically with the theory of the semiconductor triodes. In an introductory chapter the author explains what is understood by a transistor, gives the characteristics of a point transistor and compares them with those of a passive impedance quadrupole. Equivalent diagrams of a point transistor are developed by simple matrix algebra and practical values are given for the impedances. The power gain is calculated and the three different quadrupole connexions of the transistor are discussed depending on which two of the three electrodes, emitter, collector or base, are used as common input and output electrodes. Stability questions are briefly dealt with.

The next chapter is devoted to the electronics of materials. After a brief reference to the movement of slow electrons the electrons within atoms are discussed on the basis of the quantum theory and the periodical system of elements. The wave nature of the electron is mentioned and various cases of diffraction and reflexion are illustrated and calculated. Dealing with electrons in solids and with their energy levels the author shows the difference between conductors, semiconductors and insulators and particularly explains the behaviour of n-semiconductors containing "donors" and p-semiconductors containing "acceptors." Formulae are given for the quantitative investigation of these relations.

The third chapter deals with what is called contact electronics. The contact between different metals, between metals and n- and p-semiconductors and between different semiconductors is discussed as well as the rectification at contacts. Here the actions of barrier layers is explained.

After these more general explanations the fourth chapter shows their application to point transistors. Their effective resistances and current amplification at the collector electrode are calculated. The improvement gained by activation is discussed and an estimate is made of transition times, capacitances and inductances. A short section deals with noise questions and finally the point transistor using a p-semiconductor is treated.

The fifth chapter is devoted to junction transistors of the n-p-n type and other kinds and to their current amplification. Fieldistors and photo-transistors are explained.

The sixth chapter shows how the duality principle and the analogy principle may be used to explain the behaviour of transistors in comparison with vacuum tubes.

In the seventh, eighth and ninth chapters the use of transistors in input stages, output stages and as oscillators are discussed. For the input stages special consideration is given to noise values and it is stated that these have now been reduced to 2db. Limiting frequency and distortion are discussed at some length. Connexions for the output stages are first dealt with using idealized characteristics and comparisons are made with the behaviour of vacuum valves. Then the efficiency and distortion with non-idealized characteristics are discussed. Various connexions are shown for oscillator circuits, the production of negative resistances and their application for the generation of relaxation oscillations are discussed about which quite an extensive literature exists.

The final chapter deals with measuring methods and results of measurements referring to semiconductors in general and to transistors in particular. It should be noted that the "life" of the electron as here defined, i.e. the time taken for the recombination of an electron and a hole, must not be confounded with the life of the transistor, the former amounting to some 100μsec while the latter has been stated to be about 8 000h or even up to 100 000h. Some block diagrams show how the characteristics of transistors may be taken on cathode ray oscillographs and how noise values may be measured.

A bibliography of 148 items to which reference is made at the end of each section is given and covers the literature to about the middle of 1953. A combined subject and name index concludes the book which is well produced. It contains much useful information in a concise manner and should be on the shelves of every laboratory dealing with transistors. As its price is moderate, students and research workers in this and neighbouring fields will be well advised to possess the book.

R. NEUMANN.

The Practical Electrician's Pocket Book 1955

Edited by R. C. Norris. 550 pp. Demy 16mo. 57th Edition. Odhams Press Ltd. 1954. Price 5s.

THIS edition contains several new and extended chapters on electrical plan for homes, apprenticeship in the electrical contracting industry, illuminated signs and floodlighting, electrical equipment for catering, electricity in agriculture and horticulture, etc. This reference book should prove of interest.

Short News Items

Sir Anthony Eden, the Foreign Secretary, is to be the guest of honour at the annual dinner of the Radio Industry Council at the Dorchester Hotel, London, on 1 December. Lord Burghley, president of the Radio Industry Council, will preside.

The Fulmer Research Institute held an open day on 2 November when His Royal Highness the Duke of Edinburgh opened the new engineering laboratory. The Institute was founded in 1946 by the late Colonel W. C. Devereux. One of its principal aims is to bring research facilities within the means of firms which themselves are unable to maintain a full-time research laboratory. The opening of the new engineering laboratory marks another stage in the development and expansion of the Institute.

The Physical Society's autumn meeting will be held in the Poynting Physics Building, The University, Edgbaston, Birmingham, on 13 and 14 December. The subject of the meeting, which is being organized by Professor R. E. Peierls will be "A Survey of Field Theory". The meeting will consist of a series of lectures which will attempt to describe the present knowledge of the interactions between mesons, nucleons, photons and electrons, without assuming specialist knowledge. Non-members are welcome to attend this meeting and application forms may be obtained from the Physical Society, 1 Lowther Gardens, Prince Consort Road, London, S.W.7.

The Leicester firm of Taylor, Taylor & Hobson Ltd have inaugurated a new factory-to-buyer delivery service by van direct to their customers in Western Germany, Italy, France, Holland, Belgium and Switzerland. The first delivery was an electronic measuring instrument for one of Germany's largest manufacturers of ball bearings.

The 1955 Physical Society Exhibition of Scientific Instruments and Apparatus will be held in the New Hall of the Royal Horticultural Society, Westminster, from Monday, 25 April, to Thursday, 28 April. It is anticipated that about 140 exhibitors will take part. The change of venue from Imperial College to the Royal Horticultural Society's New Hall will in no way change the unique character and well-known purpose of the exhibition.

The London Electric Wire Company and Smiths Ltd, and its Associates, have removed their Birmingham branch to new and larger premises. Their address is Lewcos House, 119/120 Moseley Street, Birmingham 12. Telephone: Victoria 3731/4.

Her Majesty the Queen has been graciously pleased to approve a recommendation made by the Council of the Royal Society for the award of the Royal Medal for the current year to Sir John Cockcroft, K.C.B., C.B.E., F.R.S., for his distinguished work on nuclear and atomic physics.

The President and Council of the Royal Society have awarded the Hughes Medal to Mr. M. Ryle, F.R.S., for his distinguished and original experimental researches in radio astronomy.

The Board of Associated Electrical Industries Ltd have reorganized the AEI Companies into four groups: BTH Group, Ediswan-Hotpoint Group, M-VE Group and Overseas Group. The chairman of AEI has become chairman of each of the groups. The following are the group managing directors: BTH Group, Mr. E. H. Ball; Ediswan-Hotpoint Group, Mr. A. N. E. McHaffie; M-VE Group, Dr. C. Dannatt; Overseas Group, Dr. I. R. Cox. The group managing directors are members of the AEI Board and form, with the chairman, the Executive Committee of the Board.

Redifon Ltd announce that two 650 ton trawlers now being built in a German shipyard for the Standard Steam Fishing Co Ltd of Grimsby, and an associated company, are to be fitted with Redifon marine radio equipment. In addition, Redifon will supply V.H.F. radio and inter-communication equipment.

The Eighth Canadian International Trade Fair, sponsored by the Canadian Government, will be held in Toronto from 30 May to 10 June 1955.

The New Marconi College at Chelmsford, Essex, has recently been completed and provides facilities for technological training for nearly 100 students and residential accommodation for 50. The majority of the students are already university graduates and, in the case of British students, they are mainly employees of the company; the courses being designed to enable them to play a useful part in the company's research and development departments. Most of the overseas students are sent by their governments or by customers of the company, frequently for training on specific equipments.

A Television Receiver Intermediate Frequency of 34·65Mc/s with a high oscillator has been recommended in a recent report issued by the British Radio Manufacturers' Association. In this report, comprehensive reasons and arguments leading to the recommendation of this frequency are given; these are based mainly upon interference received from various sources and interference to other services caused by radiation from the local oscillator of the receiver. This report has received general endorsement by members of B.R.E.M.A., and has the approval of the G.P.O.

Pye/Technograph Agreement. It has been announced that Pye Ltd of Cambridge have acquired an interest in Technograph (Printed Circuits) Ltd, who hold patents for printed circuits all over the world. Mr. H. Vezey Strong, chairman of Technograph, who made this announcement recently, stated that it was the intention of both companies to work closely in touch with each other in furthering the development and application of printed circuits, not only in this country but also through their organizations all over the world.

The General Electric Co Ltd announce that they have recently equipped the Coryton Oil Refinery, owned by the Vacuum Oil Co Ltd, with a comprehensive V.H.F. radio installation. This provides two-way communication between the vehicles and a control station at transport headquarters.

Ferranti and Powers-Samas announce that they have agreed to collaborate in the development, production and marketing of electronic data processing equipment.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 29 December. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London.
Discussion: Education and Training of Radio Engineers.

Merseyside Section
Date: 2 December. Time: 7 p.m.
Held at: The College of Technology, Byron Street, Liverpool 3.
Lecture: Electronics in Materials Handling.
By: L. Landon Goodman.

Scottish Section
Date: 2 December. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders. Elmbank Crescent, Glasgow.
Lecture: Some Interesting Applications of Electronics to Photography.
By: D. M. Neale.

North Eastern Section
Date: 8 December. Time: 6 p.m.
Held at: Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: Logic, Algebra and Relays.
By: Emrys Williams.

West Midlands Section
Date: 8 December. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Wulfruna Street, Wolverhampton.
Lecture: Industrial Applications of Electronic Control.
By: J. A. Sargrove.

THE BRITISH KINEMATOGRAPH SOCIETY

Date: 15 December. Time: 7.15 p.m.
Held at: The Gaumont British Theatre, Film House, Wardour Street, London, W.1.
Lecture: Optical Developments Concerning Anamorphic Lens Systems.
By: G. H. Cook.

THE INSTITUTE OF NAVIGATION

Date: 17 December. Time: 5 p.m.
Held at: The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.
Lecture: The Accuracy of Dead Reckoning in Air Navigation.
By: C. S. Durst.

THE INSTITUTE OF PHYSICS

Date: 15 December. Time: 6.30 p.m.
Held at: The Institute's House, 47 Belgrave Square, London, S.W.1.
Lecture: Photoelectric Measurement of Polarized Light.
By: A. M. Taylor.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Radio Section
Date: 1 December.
Lecture: The Vertical Radiation Patterns of Medium-Wave Broadcasting Aerials.
By: H. Page and G. D. Monteath.
Date: 13 December.
Discussion: Practical and Economic Problems in the Maintenance of Domestic Television Receivers.
Opened by: W. L. Greenwood.

Informal Meeting
Date: 6 December.
Discussion: The Applications and Limitations of Electronic and other Computors.
Opened by: L. G. Brazier.

Measurements Section
Date: 14 December.
Lectures: An Attracted-Disc Absolute Voltmeter.
By: G. W. Bowdler.
The use of an Electron Velocity Analyzer to stabilize a 50kV D.C. Voltage source to a few parts in a million.
By: M. E. Haine and M. W. Jervis.
A Precision Direct-Current Stabilizer.
By: M. W. Jervis.

Supply Section
Date: 15 December.
Lecture: Short-Circuit Forces on Turbo-Alternator End Windings.
By: J. B. Young and D. H. Tompsett.

East Midland Centre
Date: 16 December. Time: 6.30 p.m.
Held at: The Albert Hall, Nottingham.
Faraday Lecture: Courier to Carrier in Communications.
By: T. B. D. Terroni.

Cambridge Radio Group
Date: 7 December. Time: 6 p.m.
Held at: Cambridgeshire Technical College.
Lecture: Transistor Circuits.
By: E. H. Cooke-Yarborough.

Mersey and North Wales Centre
Date: 6 December. Time: 6.45 p.m.
Held at: The Philharmonic Hall. Liverpool.
Faraday Lecture: Courier to Carrier in Communications.
By: T. B. D. Terroni.

Tees-Side Sub-Centre
Date: 1 December. Time: 6.30 p.m.
Held at: The Cleveland Scientific and Technical Institution, Middlesbrough.
Lecture: Fluorescent Discharge-Tube Circuits and Operating Problems.
By: J. Cates.

Northern Ireland Centre
Date: 14 December. Time: 6.30 p.m.
Held at: The Engineering Department, Queen's University, Belfast.
Lecture: The Possibilities of a Cross-Channel Power Link between the British and French Supply Systems.
By: D. P. Sayers, M. E. Laborde and F. J. Lane.

South-East Scotland Sub-Centre
Date: 7 December. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Technical Arrangements for the Sound and Television Broadcasts of the Coronation Ceremonies on 2 June, 1953.
By: W. S. Proctor, M. J. L. Pulling and F. Williams.

South-West Scotland Sub-Centre
Date: 15 December. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders. 39 Elmbank Crescent, Glasgow.
Lecture: Servo Machines.
By: W. S. Wood.

South Midland Centre
Date: 6 December. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Technical Arrangements for the Sound and Television Broadcasts of the Coronation Ceremonies on 2 June, 1953.
(Joint meeting with the South Midland Radio Group, and Supply and Utilization Group).

Rugby Sub-Centre
Date: 10 December. Time: 7 p.m.
Held at: The Temple Speech Room. Rugby.
Faraday Lecture: Courier to Carrier in Communications.
By: T. B. D. Terroni.

Southern Centre
Date: 1 December. Time: 6.30 p.m.
Held at: The College of Technology. Portsmouth.
Lecture: The Possibilities of a Cross-Channel Power Link between the British and French Supply Systems.
By: D. P. Sayers, M. E. Laborde and F. J. Lane.

Western Centre
Date: 13 December. Time: 6 p.m.
Held at: The South Western Electricity Board Offices. Colston Avenue, Bristol.
Repeat of Lecture given at Southern Centre.

RADIO SOCIETY OF GREAT BRITAIN

Date: 17 December. Time: 6.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place. London, W.C.2.
Annual General Meeting.

THE TELEVISION SOCIETY

Date: 10 December. Time: 7 p.m.
Held at: 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Television Circuit Refinements.
By: C. H. Banthorpe.

PUBLICATIONS RECEIVED

WIRELESS WORLD DIARY 1955. This diary is now in its 37th year of publication. It includes data for the forthcoming commercial television transmissions and the proposed V.H.F. sound broadcasts. Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 5s. 10d. (leather), 4s. 1d. (rexine).

CIRCUITS ELECTRONIQUES by J. P. Oehmichen is a well presented book of considerable interest to those concerned with the subject of electronic circuits. Societe des editions Radio, 9 Rue Jacob, Paris, 6. Price Fr. 1320.

MASS SPECTROMETRY by A. J. B. Robertson is a monograph in the Methuen series which attempt to give brief but authoritative accounts of the present state of knowledge in various departments of Chemistry. Methuen & Co. Ltd., London. Price 8s. 6d.

THE ZERO READER FLIGHT DIRECTOR, SPERRY GYROPILOTS, AIR DRIVEN HORIZON TYPE HL8 FOR AEROBATIC AIRCRAFT, AERONAUTICAL EQUIPMENT. DUPLEX FLIGHT DATA SYSTEM, A GYRO-MAGNETIC COMPASS FOR AIRCRAFT, ELECTRIC GYRO HORIZONS, and THE ZERO READER FLIGHT DIRECTOR TYPE ZL1 are the subjects of recent brochures produced by the Sperry Gyroscope Co. Ltd., Great West Road, Brentford, Middlesex.

BSI ANNUAL REPORT 1953-54 runs to a total of 200 pages and provides almost at a glance an impression of the very comprehensive range of industries which use the BSI facilities to prepare agreed standards for their products and services. British Standards Institution, British Standards House, 2 Park Street, London, W.1. Price 5s.

TELEVISION CIRCUIT REFINEMENTS is a book by C. H. Banthorpe explaining the purpose and operation of old and new circuit developments used to improve the performance, reliability or safety of commercial television receivers. Norman Price (Publishers) Ltd., 283 City Road, London, E.C.1. Price 5s.

VADE MECUM 1954 is the 11th edition of this comprehensive valve guide. Valves are tabulated in alphabetical and numerical order. P. H. Brans Ltd., Antwerp. Price 25s.

INDUSTRIAL ENGINEERING is the subject of the last of the reports of the series of British Productivity teams that have visited America. The main body of the report contains chapters on the techniques and functions which either come within the scope of industrial engineering or closely affect its practice, followed by others on human factors with which the industrial engineer is vitally concerned. The Anglo-American Council on Productivity, 21 Tothill Street, London, S.W.1. Price 5s.

RADIOLOCATOR IV is a brochure which provides, in pictorial form, an explanation of the controls and switch selection applicable to the Marconi Marine Radiolocator IV radar equipment. It should enable personnel to operate the apparatus with the minimum of reference to the more detailed technical instruction manual supplied with the installation. The Marconi International Marine Communication Co., Ltd., Chelmsford.

TRUVOX RADIO JACK leaflet describes a unique unit for use with a tape recorder which is designed to give direct reception from either of two local stations or to make recordings, for future playback, of any of the programmes radiated by the selected stations. Copies are available from Truvox Ltd., 15 Lyon Road, Harrow, Middlesex.