Electronic Engineering

Vol. XXIV. **DECEMBER 1952** No. 298.

Commentary

ON the whole, the relationships between publishers and authors of scientific works are among the happiest, and the law of copyright, designed to protect their mutual interests, would appear to be in little need of revision. But scientific works do not in general create great public interest and once we embark on matters which affect the wider public we are immediately aware that all is not well with law of copyright.

Very welcome therefore is the report of the Copyright Committee* which has just been published and which, if the recommendations contained therein are accepted, will go a long way to clearing up the present state of confusion, although with regard to some of the more controversial problems the Committee finds itself unable to make any suggestions. The Committee was set up in April, 1951, under the chairmanship of Lord Reading (who, however, resigned on his appointment as Under-Secretary of State for Foreign Affairs and was replaced by Sir Henry Gregory), and under its terms of reference was "To consider and report whether any, and, if so, what, changes are desirable in the law relating to copyright in literary, dramatic, musical and artistic works in particular regard to technical developments and to the revised International Convention for the Production of Literary and Artistic Works signed at Brussels in June, 1948, and to consider and report on related matters."

The present Copyright Act based on the original Berne Convention of 1908 was passed three years later to remove the state of confusion existing at the time and "to put the British Law on an intelligible and systematic footing." Since that date there have been revisions to the Berne Convention of which the latest, to which Great Britain is proposing to accede, was signed at Brussels in 1948. But the original Act was passed more than forty years ago and there were few technical developments at the time which called for special treatment in the Act. The gramophone was, of course, the first of what was referred to as a technical development and though special recommendations were put forward at the time, they were not embodied in the Act.

Since the gramophone, have come the cinematograph, radio and television in which public interest is great, and the Committee feels, quite understandably, that the present recommendations will be subject to considerable criticism and debate.

There has been in the post-war years a tremendous revival of interest in the gramophone due, no doubt, to the improved methods of recording and reproduction but controversy between the gramophone recording companies and performers is not so keen as it was. The main problem today is to break the deadlock existing between the BBC on the one hand and the various sporting and theatrical interests on the other.

To a limited extent the BBC and the sporting associations have been able to come to terms in the past and only shortened sound broadcasts have been allowed for some of the major sporting events. But the advent of television has brought to the fore the thorny problems of copyright and fees, and the Committee does not find any ready solution. The sporting associations fear that the televising of a sporting event may seriously prejudice attendances not only at the event itself but at other events which might be running concurrently and that if the televising be allowed, the associations should be given some measure of control. They, argue, for example, that if an event is taking place in London it should not be televised in the London area and certainly not nearer than, say, Birmingham. All this, of course, presupposes that both sides have agreed on the scale of fees to be paid by the BBC.

Matters are made more difficult where a public performance of a televised programme is given, say in a cinema. The Committee recommends that promoters of sporting or other events should not be allowed to prevent the copying or recording of their events and that the BBC—or any other broadcasting authority—should have the right to control public performances of its television programmes. This would, it is stated, also allow the payment of extra fees and at the same time the promoters' interests would be safeguarded.

If this were the only source of contention there would be hope that an amicable solution would be reached, but the BBC is in a difficult position for not only is the BBC a user of copyright material, but it assembles programmes of its own at considerable cost and skill, paying for copyright material where required. At the present moment there is nothing to prevent subsequent copying of these programmes and indifferent reproductions have already been made and sold with embarrassment to the BBC from both financial and prestige points of view.

As we see it, the existing copyright law does not prevent the musical enthusiast from building up a library of great music performed by some of the world's best artists merely by recording on tape selected items of the BBC programmes. Not only need no copyright fee be paid, but such a library could be obtained at a fraction of the cost of an equivalent in gramophone records.

* *Cmd 8662 H.M.S.O. Price 4s. 6d.*

The Transbooster

An Improved Method of Rectifier Voltage Control

By A. H. B. Walker*, B.Sc., A.C.G.I., D.I.C., A.M.I.E.E.

This article describes a new system of controlling the output voltage of a rectifier by saturated chokes (transductors), which overcomes many of the disadvantages, (e.g. large size of chokes, slow response speed, poor efficiency) usually associated with transductor control methods.

A 3 kW "Transbooster" Constant Potential Rectifier

THE D.C. saturated choke or "transductor" has been known for many years as a means of controlling A.C. fed equipment. It provides the advantages of completely static operation, unlimited life and simple construction, while since these advantages are also offered by the metal rectifier it is not surprising that the transductor has been widely used in the past for rectifier control. However, the development of modern high efficiency metal rectifiers has extended their field of use into higher power ranges, and this has emphasized certain defects of transductor control which, although long known, have not been so serious in smaller equipment.

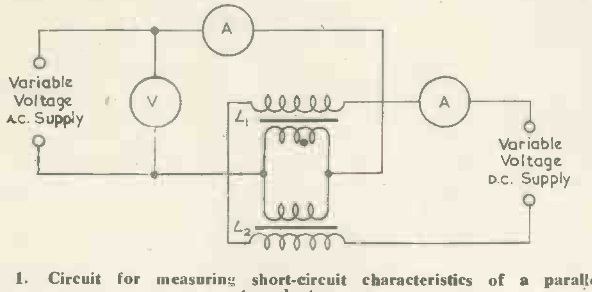


Fig. 1. Circuit for measuring short-circuit characteristics of a parallel transductor

Basic Principle of Transductor

Although the method of operation of transductors will be familiar to many engineers, a brief description of the basic principle will be given, since this will assist in a clearer explanation of the new "Transbooster" circuit to be described.

Transductor control is sometimes explained as a method of varying the inductance of chokes by D.C. saturation, but this is not strictly true and the operation may be explained by considering the magnetizing current of the two chokes L_1 and L_2 shown in Fig. 1. This arrangement is known as a parallel transductor.

Here the two chokes are connected in parallel across a variable voltage A.C. supply, and they are provided with secondary windings which are connected in series to a variable D.C. supply. Since the two main choke windings are connected in parallel the A.C. voltages across them must be equal, and if the windings are identical the A.C. voltages induced in the secondary windings must also be equal. The two secondary windings can therefore be connected in opposition, as shown, so that no A.C. voltage appears across the leads to the D.C. supply. If the combined magnetizing characteristic of the two chokes is now measured with various fixed values of direct current in the secondary windings, the family of curves shown in Fig. 2 is obtained.

Clearly the first curve, without D.C. saturation, is the normal magnetization curve of the chokes, and depends on the characteristics of the iron. The effect of various fixed values of D.C. is to shift the curve to the right, but note that it still retains a considerable portion which is nearly vertical, over which range the A.C. current remains almost unaffected by the applied A.C. voltage, and is almost proportional to the direct current. The effect of the D.C. is therefore seen to be much more than a simple variation of the inductance of the choke, and the proportional current effect is used, for example, to produce D.C. current transformers for measuring very heavy direct currents.

Simple Circuit for Controlling a Load

The most usual method of using the transductor characteristics to control the power in a load is to connect

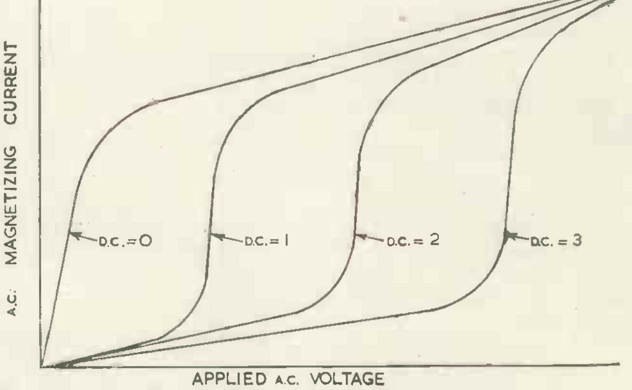


Fig. 2. Transductor magnetizing current or "short-circuit characteristic" as measured in Fig. 1

* *Westinghouse Brake and Signal Co., Ltd.*

the transductor in series with the load across the supply. The direct current fed to the transductor, hereafter termed the control current, can then be varied by a relatively low-power hand control or an automatic device, to regulate the voltage applied to the load as desired. Thus in Fig. 3 the transductor, which is now shown diagrammatically as a single choke winding with a saturating winding at right-angles to it, is used to control the power in the load over a considerable range by means of a relatively small variable resistor.

Conventional Transductor Control of a Rectifier

The main objective of transductor control of rectifiers is usually to achieve a constant voltage output from the rectifier despite mains voltage and frequency variations and large changes in the D.C. load. Although various detail modifications to the circuit are possible, the basic control method is to use the transductor as a variable series impedance between the A.C. terminals of the rectifier and the supply. The simplest possible form of this arrangement is a single-phase rectifier having a transductor connected in one A.C. line, and for the purpose of comparison, this circuit is shown in Fig. 4 with all non-essentials removed.

This can be analysed as follows, to show that the transductor size is determined absolutely by the range of

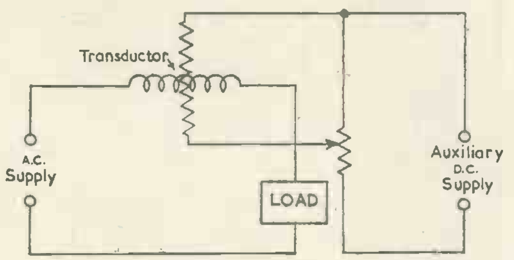


Fig. 3. Series control of an A.C. load by a transductor

control required, i.e. by the sum total of all the variables which have to be compensated in order to preserve a constant D.C. voltage.

Ignoring harmonics, the conditions on the A.C. side of the rectifier are illustrated by the vector diagram in Fig. 5. Taking first one limiting condition, that is, with light load on the rectifier (I') and the supply voltage at its top limit of variation (V_s'') the transductor voltage must be regulated to V_{T_1}.

If the supply voltage now falls to the lower limit (V_s'), the transductor voltage must be regulated to V_{T_2} to preserve the rectifier input voltage (V_{R_1}) unchanged. If the load is now increased to the maximum value (I'') while the supply is still low, the other limiting condition is reached since the rectifier input voltage must now be increased to V_{R_2} to maintain the D.C. voltage constant. The transductor voltage must therefore be reduced to V_{T_3} in spite of the heavy increase of current through it, by increasing the control current considerably. In order to achieve the necessary regulating effect with the greatest economy, it is necessary to proportion the quantities roughly as illustrated in the vector diagram of Fig. 5, the transductor voltage being of the same order as the rectifier input voltage, both being approximately equal to the supply voltage/$\sqrt{2}$ at full load. In this conventional circuit, therefore, the transductor must be wound to withstand a maximum A.C. voltage of about $V_s/\sqrt{2}$, and it must also have windings sufficiently heavy to carry the maximum output current. Thus, although maximum current and maximum voltage never occur simultaneously in the transductor, its size is determined by their product as in a normal transformer. Moreover, there is no possibility of working the core at a high flux density as the natural magnetizing current (with D.C. control current nil) must be kept low. The transductor in this conventional control system is therefore inevitably large, heavy and expensive, and is usually about 20 per cent larger than the main transformer.

It can be shown that the VA rating of the transductor is given by the following expression: —

Transductor VA rating =

$$I''\sqrt{V_s''^2 - [\sqrt{2}.V_s'' - V_s' - R(I''-I')]^2} \quad \ldots\ldots (1)$$

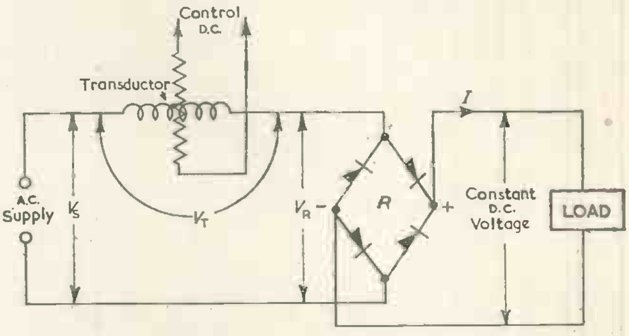


Fig. 4. Basic circuit for conventional transductor control of a rectifier

where V_s'' = maximum supply voltage
V_s' = minimum supply voltage
R = Equivalent circuit resistance (including rectifier and other resistances)
I'' = maximum load current
I' = minimum load current

Use of a Transductor Regulated Booster or "Transbooster"

If we examine the output regulation curve of a straight transformer and rectifier combination for high and low limits of mains voltage (Fig. 6), a possible way of greatly reducing the size of the regulating transductor suggests itself.

If, for example, a straight transformer and rectifier are so designed that the desired output voltage is only reached at point A, that is, at very light load with the mains voltage

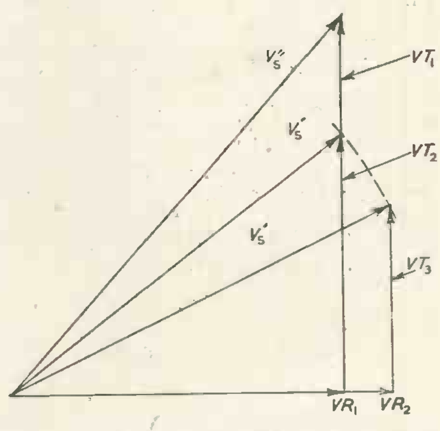


Fig. 5. Vector relationships in Fig. 4 for simple constant voltage rectifier

high, then it is apparent that the output voltage can never exceed the desired value, but that it will, in fact, be slightly low for all other conditions of load and mains voltages. All that is required therefore is a low voltage boost which can be added to the "natural" rectifier output voltage. If this boost voltage can be varied from nil at point A, through a range of intermediate values (XX) to a maximum value (YY), the total output voltage of the rectifier can be held constant at the desired value and the regula-

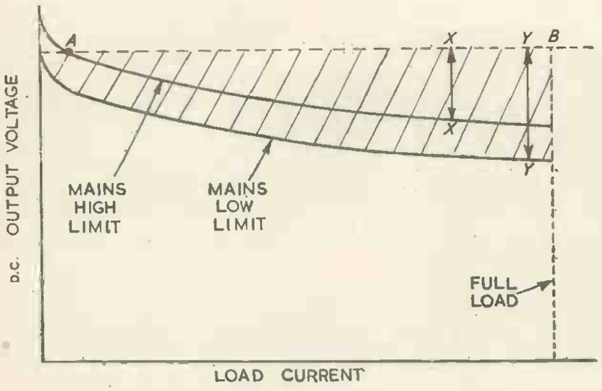


Fig. 6. Regulation curves of straight transformer and rectifier for two supply voltages, showing boost required to provide constant output voltage

tion curve will become the straight line AB. In this case the booster only has to supply the power inside the shaded area, and although it must still be rated for the full load current it now has only to provide the boost voltage, which is but a fraction of the full output voltage.

It is apparent that if any attempt is made to use a transductor to control a series voltage boost on the A.C. side of the rectifier, no advantage will be gained, since the circuit will always reduce effectively to Fig. 4, and it will be found that the transductor will still have to withstand $V_s/\sqrt{2}$. To realize the saving indicated as possible in Fig. 6, the boosting transductor must be transferred to the D.C. side of the main rectifier when it becomes, in effect, isolated from the supply voltage and can be designed simply to deal with the total boost voltage required to cover rectifier regulation and mains variation. Such a transductor controlled D.C. booster, or "Transbooster", therefore becomes as, in its simplest form, shown in Fig. 7.

Allowing for the quadrature vector relationship at full load between the transductor voltage and the voltage at the input to the boost rectifier, the boost transductor need only be wound to withstand about $\sqrt{2}$ times the boost voltage, instead of mains voltage/$\sqrt{2}$ as in the conventional circuit.

Another way of looking at the saving is that the conventional transductor has to be designed in any case, in a way which would make it possible to operate it down to short-circuit although this condition is not actually required; whereas in the Transbooster circuit we can in fact take full advantage of this facility by operating the transductor right down to short-circuit (zero-boost) conditions. This means that in the Transbooster circuit a greater vector swing occurs than in the conventional circuit. This is shown in Fig. 8 which is drawn for two A.C. voltages on the input side of the booster rectifiers for comparisons with Fig. 6.

Fig. 7. Simple form of Transbooster circuit

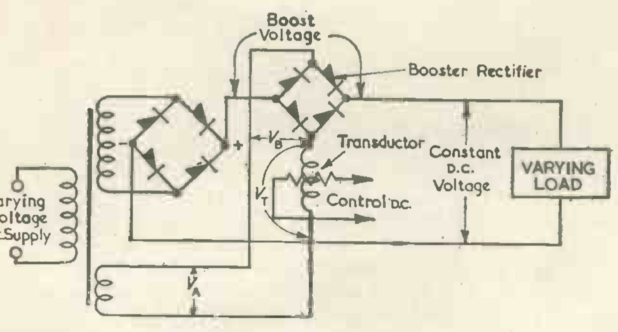


Fig. 8 shows that a swing of over 45° occurs, giving a very large variation of the boost voltage (V_B) from substantially nil to the maximum required, while the transductor voltage (V_T) varies from almost the full applied auxiliary voltage ($V_{A\max}$) down to a value slightly lower than the maximum input voltage to the boost rectifier.

An expression for the VA rating of the transductor when used in the Transbooster circuit can be derived in a similar manner as was done for the conventional circuit:

Transductor VA rating =

$$I'' \times \sqrt{2}[V_s'' - V_s' + R(I'' - I')] \quad \ldots\ldots\ldots (2)$$

where the symbols are the same as for Equation (1).

Example of Saving of Transductor kVA Rating in Transbooster Circuit

An approximate comparison of the transductor ratings in the two circuits for a typical case may be made by assuming, for example, that a 3kW rectifier having a smoothed output of 50V at 60A is required. The mains variations to be compensated are (say) −10 per cent and +6 per cent and the natural regulation of the rectifier, transformer and smoothing filter is assumed to be 15 per cent.

From (1) the transductor rating in the conventional

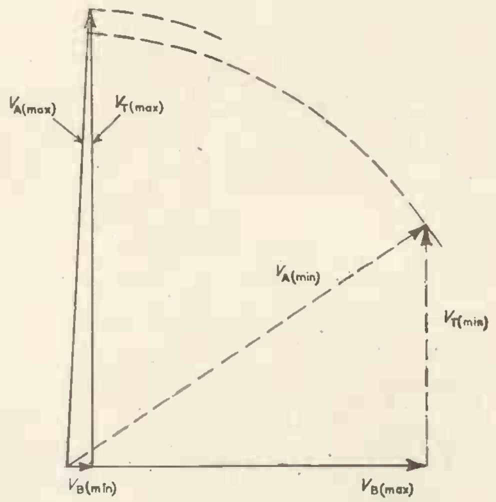


Fig. 8. Vector conditions in single phase Transbooster circuit of Fig. 7. *Full lines: high mains, minimum load, negligible boost. Dotted lines: low mains, maximum load, maximum boost*

circuit is given by:

$$\text{Transductor VA} = 60 \times \sqrt{95^2 - [\sqrt{2} \times 81 - 95 - 7{\cdot}5]^2}$$
$$= 5{\cdot}6\text{kVA}$$

From (2) the transductor rating in the Transbooster circuit is given by:

$$\text{Transductor VA} = 60 \times \sqrt{2}[95 - 81 + 7{\cdot}5]$$
$$= 1{\cdot}8\text{kVA}$$

In this typical example, therefore, the Transbooster circuit enables the transductor to be reduced to a third of the size which would otherwise be required. The exact saving obtained naturally varies with the application and the conditions to be dealt with (e.g. extent of mains voltage variation) but it is always very substantial.

Other Advantages of the Transbooster

In addition to this great saving in size, weight, and cost, the Transbooster offers further advantages. Since the transductor is smaller, the power required in the D.C. control winding is also smaller, so that the controller, which has to provide this power is also smaller and cheaper. For the same reason the overall power factor of the equipment is high, and since the booster becomes less active as the load is reduced, its losses tend towards nil and the efficiency is particularly well maintained down to low loads (see Fig. 11).

By comparing Fig. 4 and Fig. 7 it will be seen that in the Transbooster circuit there is no reactive impedance between the mains and the load. This means that the Transbooster can deliver heavy surge currents to variable loads without the output voltage ever dropping below the regulation curve of the basic rectifier, (see Fig. 6). Since it is impossible to change the D.C. saturation current in the transductor instantaneously in either circuit, this is a great advantage over Fig. 4, in which the output voltage might instantaneously drop nearly to zero as a result of a sudden load increase. In the Transbooster circuit the surge current is solidly fed from the mains and the booster rectifier "free-wheels", rapidly trimming up the voltage to the required value immediately after the surge. This lack of series impedance also enables the Transbooster to supply heavy transient overloads, up to say five times full load, the voltage merely falling gently down the main rectifier regulation curve and recovering immediately to the controlled value when the overload is removed. This feature is required, for example, when a constant voltage

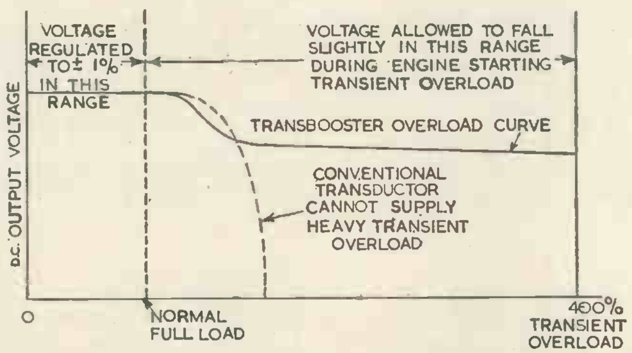


Fig. 9. Output characteristic of Transbooster designed for heavy transient overloads (e.g., Engine starting). *Normal transductor control would have dotted characteristic and could not supply the overload*

rectifier is required to supply a heavy engine starting current, during which time the voltage can be allowed to fall (see Fig. 9). It would be impossible to provide such an overload characteristic by conventional transductor control.

For floating battery installations it is desirable to provide some form of current limiting to protect the rectifier from sustained overload after a supply failure, when the rectifier has to charge the battery and supply the load simultaneously. This can be achieved with the Transbooster by suitable controller design so that protection against overload down to the minimum battery voltage of 1·8 volts per cell is achieved. The output characteristic of a Transbooster provided with this feature is then as shown in Fig. 11.

Typical Design of Transbooster Equipment

A circuit diagram for a three-phase Transbooster regulated rectifier is shown in Fig. 10. A three-phase Transbooster is used but it is also possible to use a single-phase Transbooster in association with a three-phase main rectifier if desired, as this may prove economic in small equipments. The Transbooster is fed from an auxiliary low voltage delta winding on the main transformer, and the phase shift thus provided, together with the phase shift in the transductors assists in reducing the output ripple.

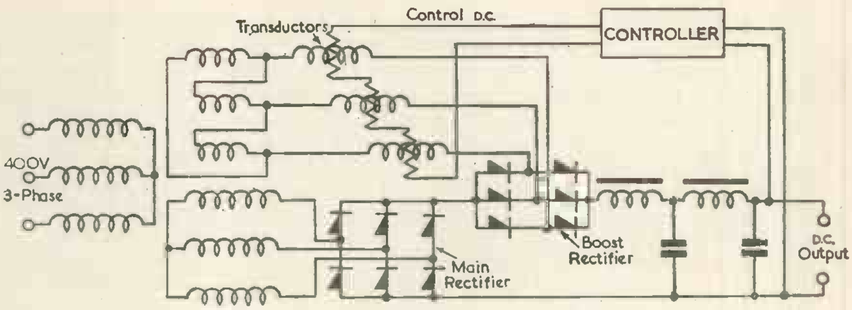


Fig. 10. Three-phase "Transbooster" Constant Voltage Rectifier

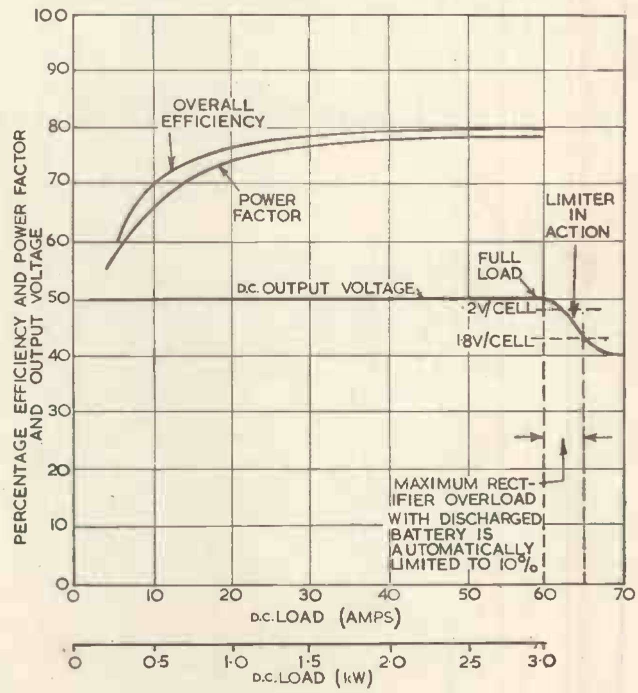
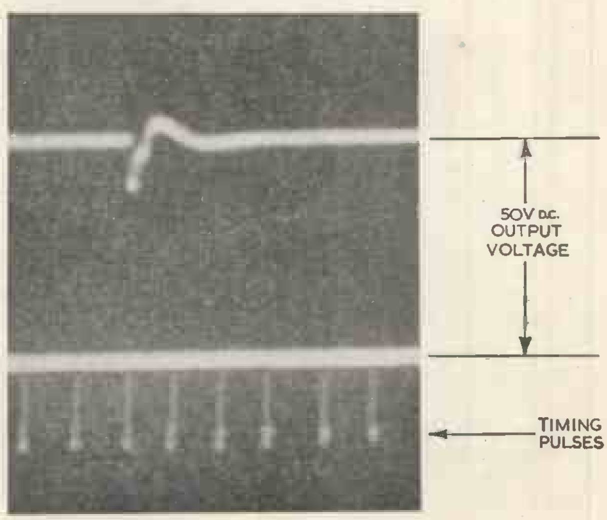


Fig. 11. Overall performance of 3kW Transbooster Rectifier, for floating battery applications. *Note the high efficiency even at low loads. (Controller losses and smoothing filter losses are included)*

Fig. 12. Oscillogram showing the effect of sudden large load increase (50 per cent of F.L.) on Transbooster fully smoothed output voltage. Initial correction is made in 0·05 seconds and is complete in 0·4 seconds. (0·25 sec Marker Pulses)



The performance of a typical 3kW Transbooster rectifier, rated at 50 volts 60 amperes for telephone power supply is given in the curves in Fig. 11. These curves show the high overall efficiency (82 per cent) and also illustrate the fact that the efficiency is exceptionally well maintained down to light loads, i.e. 70 per cent at 1/6th of full load and 65 per cent at 1/8th of full load.

The output voltage is held within limits of ± 1 per cent from 1/20th up to full load and is independent of mains voltage variations from −10 per cent to +6 per cent and frequency variations from 48 to 52c/s.

The rapid rate of correction of the output voltage which is achieved after a sudden large load increase (50 per cent of full load) is shown by the oscillogram in Fig. 12.

This shows that in spite of the delay inevitably caused by the long time-constant of the smoothing filter, initial correction of the D.C. voltage is made in 0·05 seconds, and that correction is complete within approximately 0·4 seconds.

In conclusion, it is clear that many of the fundamental disadvantages of conventional transductor control have been overcome by the Transbooster circuit which appears likely to widen the scope of entirely static constant voltage rectifier equipment.

Transmitter Combining Circuits

By A. R. A. Rendall,* Ph.D., B.Sc., M.I.E.E.; G. A. Hunt,* B.Sc.

UNDER the present conditions of overcrowding of European broadcasting stations in the medium waveband the only practical, though limited, means of improving broadcast coverage on medium waves in the United Kingdom is by supplementing the existing high-power transmitters with a number of low-power stations operating on a shared wavelength basis. The BBC has recently brought eight such stations into operation and others are planned or under construction. In the interests of the economic use of technical manpower, these small stations are designed to operate either automatically or by remote control without the need to have local staff in attendance. Under these conditions, in order to improve reliability, the full output power of the station is provided by two or three transmitter units arranged to feed into a common aerial, so that if one transmitter should fail the service is maintained, although at a reduced power. If these transmitters are paralleled directly on to the common aerial circuit, the arrangement will suffer from the following disadvantages.

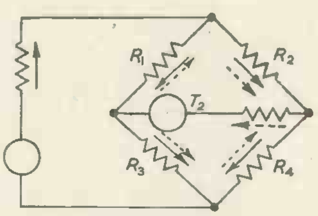


Fig. 1. A Wheatstone Bridge arrangement of Transmitter loads

In the first place, certain types of faults from one transmitter (such as a short-circuit) will affect the common circuit.

Secondly, a phase difference between the transmitters would lead to the current of one transmitter circulating in the output of the other and would be likely to cause non-linear distortion.

Thirdly, there is no direct way of monitoring the output of one transmitter independently of the other, so that a faulty transmitter can be detected and switched off.

Fourthly, if one transmitter is switched off, the effective load impedance on the other transmitter is reduced with possible deterioration in power efficiency and change of modulation depth. It may also cause excessive anode currents which would trip the remaining transmitters.

The circuit of an A.C. Wheatstone bridge has properties which could be applied to overcome some of these difficulties.

Fig. 1 shows such a bridge and, if the resistances R_1, R_2, R_3, R_4 are all equal, then it is clear that an E.M.F. from transmitter, T_1, will feed into the resistance arms of the bridge but not into T_2. Likewise, T_2 will not feed into

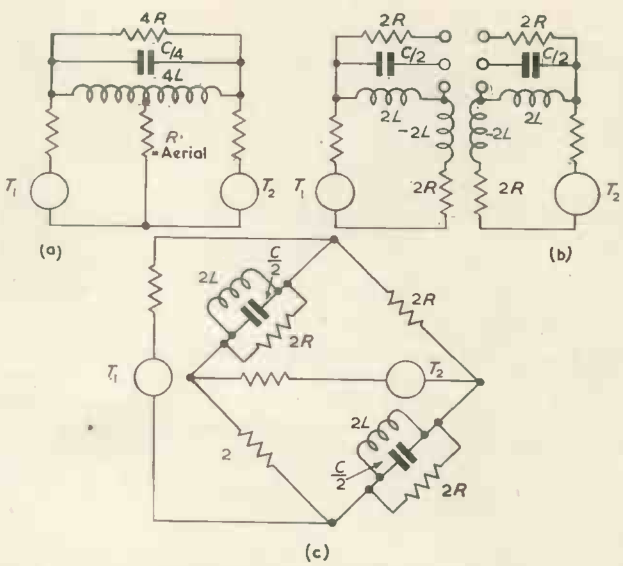


Fig. 2. (a) Hybrid Coil Coupling for 2 transmitters; (b) Bartlett bisection of Fig. 2(a); (c) Equivalent bridge circuit

T_1. Furthermore, if T_1 and T_2 are in phase, then, at any instance of time, the directions of flow of currents in the resistance arms are as shown. These currents oppose in R_1 and R_4 and add in R_2 and R_3. If R_2 and R_3 can be made the aerial, and R_1 and R_4 the balancing resistances, then it is clear that no power will be dissipated in the balancing resistances and the total power will be supplied to the aerial. If the E.M.F.s of transmitters T_1 and T_2 are not in

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* *Designs Dept., BBC Engineering Division.*

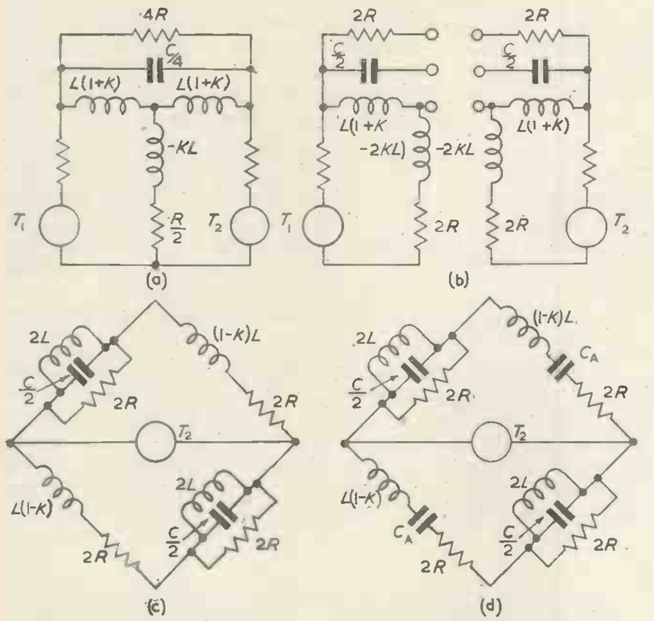
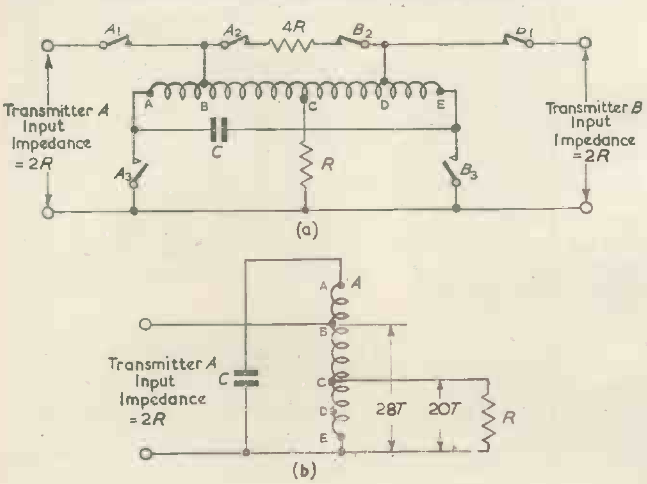


Fig. 3. (a) Equivalent circuit Fig. 2(a) but with coupling coefficient of K. (b) Bartlett bisection of Fig. 3(a). (c) Equivalent Bridge Circuit of 3(a). (d) Equivalent Bridge Circuit with reactance correction

phase, this cancellation does not occur and there is power dissipation in the balancing resistances, R_1 and R_4, but still no power from T_1 would be developed in T_2 or vice versa. Thus, although the output circuit efficiency would be impaired, there would be no possibility of interaction.

The bridge circuit, as shown, is not particularly convenient to use, but there are very many other electrical networks which perform in a similar way. The hybrid coil, which is well known to communication engineers, is a convenient example and is shown in Fig. 2(a). This consists of two close-coupled windings and transmitters feeding as shown. By Bartlett's bisection theorem, the hybrid coil of Fig. 2(a) can be replaced by networks shown successively in Figs. 2(b) and 2(c). The coupling between the two halves of the coil is assumed to be unity. A bridge circuit results which is perfectly in balance at the resonant frequency of L and C. If the coupling between the two halves of the coil is not unity then the equivalent network of Fig. 2(a) is as shown in Fig. 3(a) which, by the bisection theorem,

Fig. 4. (a) Circuit for restoration of transfer efficiency when one transmitter fails. (b) Equivalent with one transmitter off



could be represented as Figs. 3(b) and 3(c). One method by which Fig. 3(c) can be brought into balance at the resonant frequency, is by the addition of a capacitor C_A as shown in Fig. 3(d). There is a wide variety of bridged-T networks which can be made to satisfy the required conditions and the most convenient of these may be selected for each particular case. From Fig. 2(a) it will be seen that, if the aerial impedance is R, then the impedance facing either transmitter is $2R$. If, now, one transmitter were switched off, the impedance facing the other transmitter would still be $2R$, but half its power would be dissipated in the balancing resistance and half in the aerial. Thus, compared with the case where both transmitters are giving power, a 6db loss in radiated power would result if one transmitter closed down. Fig. 4 shows how the network can be re-switched, so that the impedance relations are maintained correct and no power dissipated in any balancing resistance[1]. In this particular case, a tapped coil

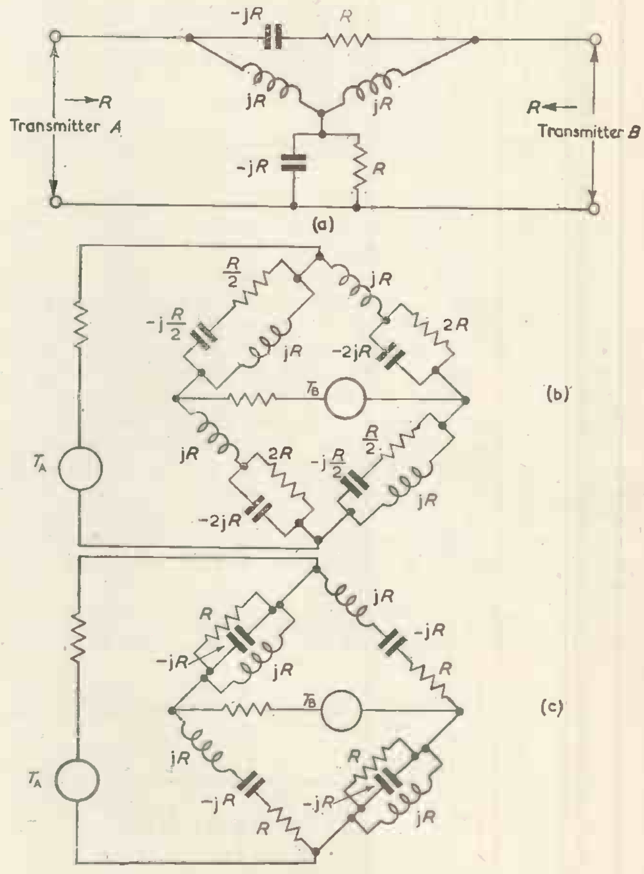
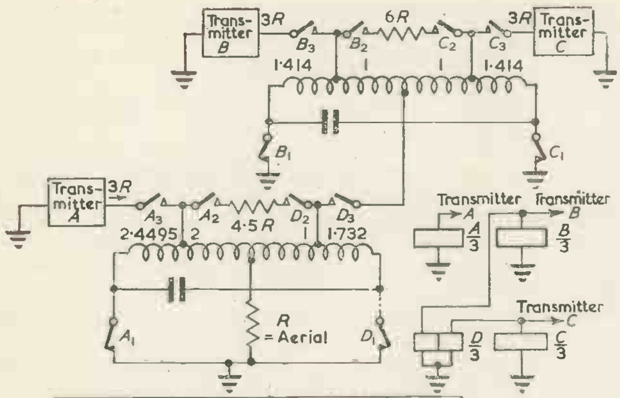


Fig. 5. (a) Bridged-T circuit for coupling 2 transmitters to one aerial. (b) Equivalent Bridge Circuit of Fig. 5(a). (c) Equivalent Bridge Circuit of 5(b) at carrier frequency

is used, but there are other circuits avoiding the use of tappings whereby the switching can be carried out and the proper power and impedance relations maintained. Such a circuit developed by Marconi's Wireless Telegraph Co. Ltd. is shown in Fig. 5(a)[2]. There is no mutual coupling between the coils and the equivalent bridge circuit is shown in Fig. 5(c). It should be readily seen from this that, under normal balance conditions, the match is perfect, and the impedance looking into the bridge terminals is equal to the aerial load resistance. The impedance relationships shown in Fig. 5(a) illustrate that when one transmitter is disconnected, a direct connexion can be made between the



Transmitters	Relay	Efficiency percentage	Power
A B C	A B C D	100	3
B C	B C D	100	2
A B	A B D	97·15	1·943
A C	A C D	97·15	1·943
A	A	100	1
B	B D	100	1
C	C D	100	1

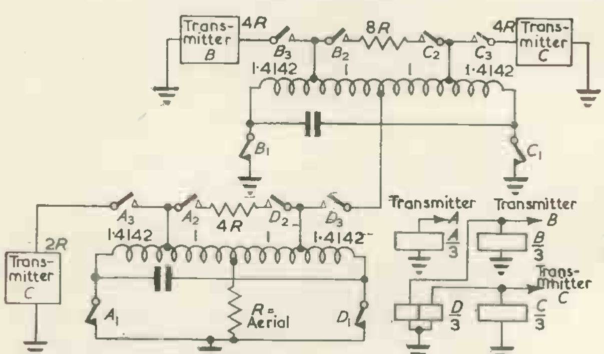
Fig. 6. System for coupling 3 transmitters to an aerial using one asymmetrical hybrid coil and one symmetrical hybrid coil

transmitter and the aerial circuit, giving perfect matching and optimum power conditions.

In many cases three transmitters are used to give increased reliability. In these cases it is possible to combine two of the transmitters to form one of double power by using one of the previous circuits, and these two transmitters may then be combined with the third by using an asymmetrical hybrid or bridged-T circuit. Some loss of efficiency is inevitable with the simple switching circuits shown in Figs. 6 and 7, but since this does not exceed about 3 per cent it is unimportant, and at all times the loading of each transmitter is maintained at its correct value. With the circuit shown in Fig. 6, the inefficiency occurs when either transmitter *A* or *B* has been closed down. In Fig. 7 the inefficiency occurs when all three transmitters are working.

In all the circuits considered here the voltage at any transmitter terminal is due to that transmitter alone, and

Fig. 7. System for coupling 3 transmitters to an aerial using two symmetrical hybrid coils



Transmitters	Relay	Efficiency percentage	Power
A B C	A B C D	97·14	2·9142
A B	A B D	100	2
B C	B C D	100	2
A C	A C D	100	2
A	A	100	1
B	B D	100	1
C	C D	100	1

as a result a simple crystal detector connected at that point will provide an audio output derived from that particular transmitter alone. Hence the output of each transmitter can be compared with its input by means of an automatic monitor and a faulty transmitter thereby identified and closed down if the distortion becomes too great[3].

Both the hybrid circuit and the bridged-T circuit are employed by the BBC. At Ramsgate, where each unit transmitter has a power of less than 1 kilowatt, the iron dust cored coil shown in Fig. 8 is used together with standard 3 000 type relays in the circuit of Fig. 4(a). The closing down of either transmitter causes the re-switching of the coupling circuit so as to maintain correct load of the transmitter still operating, without interruption of the service.

At Daventry, where the third programme high-power transmitter is also remotely controlled and operated on an unattended basis, a power of 150kW is provided by two transmitter units working in parallel[2] and, due to the high voltages and currents involved, it is not practicable to make a hybrid coil with the requisite coupling factor, so the circuit of Fig. 5 is used. If one unit transmitter fails,

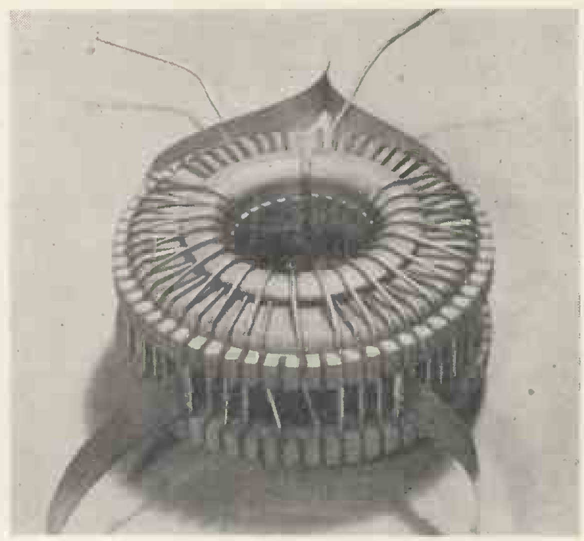


Fig. 8. Iron Dust Cored Hybrid Coil for coupling M.W. Transmitters. Used for powers up to 1kW

it is necessary to remove power from both unit transmitters during the re-switching operation and a short break in transmission occurs before the output circuit efficiency is restored.

Both these systems have been in use for a considerable time and have given satisfactory service. As a result of this experience further stations using one of these systems are likely to be built in the near future. The type of bridge to be employed will be determined by particular conditions at these stations.

Acknowledgment

The authors are grateful to the Chief Engineer of the British Broadcasting Corporation for his permission to publish this article.

REFERENCES

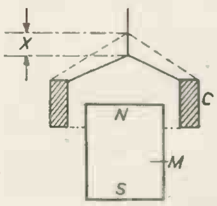
1. Pro. Pat. 15481/52.
2. LESLIE, R. W., GUNN-RUSSELL, C. Automatic Broadcasting. *Wireless World*, 58, 449 (1952).
3. PEACHEY, F. A., TOOMBS, R. BBC New Automatic Unattended Transmitter Technique. *Electronic Engg. 24*, 446 (1952).

Vibration Measurements

By Rose Winslade*

The study of mechanical vibration phenomena has become an important factor in the efforts made to improve and to control the quality of technical products. This article is intended as an introduction to some of the instruments at present being used for this study. Of necessity some aspects of the subject, such as phase distortion, losses in the integrating and differentiating circuits, etc., have been treated superficially, but it is hoped that these might be dealt with more fully at a later date.

A VIBRATION detector can be sensitive to the displacement, the velocity or the acceleration of a vibration. Since it is usually necessary to know the form of all three of these components of vibration, it is necessary to integrate or differentiate the signal from the detector. An electronic network will perform the mathematical operations of integration and differentiation making it simple, for instance, to take the output from an accelerometer and by integration to convert it to a signal representing the velocity of the vibration. Conversely, differentiating the output from a position sensitive instrument, i.e., a displacement pick-up, gives the same signal as that which would be obtained from a velocity sensitive instrument.



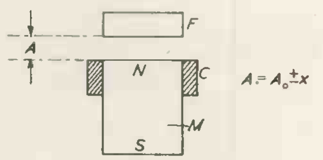


Fig. 1. (Left). Electro-dynamic system

The voltage induced in the windings of the coil c is at any instant directly proportional to the velocity dx/dt of the relative displacement x.

Fig. 2. (Right). Electro-magnetic system

The magnet M has a coil c wound directly upon it. When a ferromagnetic body F placed opposite one of the poles of the magnet is moved with respect to the magnet, a change is effected in the magnitude of the magnetic flux linking the coil. This causes a voltage proportional to dφ/dt to be generated in the coil.

Displacement can be measured by means of a detector employing strain gauges, making use of the differential transformer, or by means of a capacitance type instrument. Of these three types the capacitance pick-up is usually preferred because of its linear frequency response and its ease of construction.

Two types of *acceleration* conscious detectors employ (a) the strain gauge principle and (b) the differential transformer. The basic element of each consists of a mass mounted on a spring suspension that permits motion in one plane only. In the strain gauge type a seismic system is formed by a cylindrical mass suspended between two leaf springs. The flexure of the leaf springs is measured with the aid of four strain gauges fixed to the springs. The springs are arranged to permit axial movement only. The force required to accelerate the suspended mass changes the length of the strain gauges and hence their resistance, this in turn causes an unbalance of a strain gauge bridge; temperature change is usually automatically compensated for. The second type of instrument also contains a spring suspended mass, in this case the movable core of a differential transformer. Here, too, acceleration in the axial plane of the core of the transformer (or the component of acceleration in that plane) causes a relative movement between the core and the rigid frame and thus effects the electrical output.

Unfortunately, any spring mounted mass is a mechanically resonant system and vibrations with a frequency approaching the natural resonant frequency of the detector cause deflexions out of proportion to the applied motion. This can be corrected by filling the case of the detector with a viscous liquid to damp the motion.

With the *velocity* conscious pick-up there are two main methods of electro-mechanical conversion, the first is to allow the movement of a coil perpendicular to a field of constant strength as in Fig. 1. In this case the voltage induced in the coil c is at all times directly proportional to the velocity (dx/dt) of the relative displacement of the vibration, as variation in the rate of displacement causes a change in the rate of cutting lines of flux and therefore a change in the induced volts. In the second method, Fig. 2, a coil is wound upon a magnet M. When a ferromagnetic body F is moved relative to one of the poles, a change in flux linkage is effected. This again causes a voltage to be induced in the coil and this voltage is proportional to $d\phi/dt$, and may be expressed thus:

$$d\phi/dt \,.\, d\phi/dA \,.\, dA/dt = \frac{d\phi/dA(A_o \pm x)}{dt} = d\phi/dA \,.\, dx/dt$$

where A_o = the (constant) average distance.

In both types of instrument the voltage developed is proportional to the velocity of the mechanical movement being followed, and is dependent on the average distance. ($d\phi/dA$ may be regarded as a sensitivity factor).

There are several advantages to be gained from the use of these methods of conversion, viz:

1. Simple . . . no external current source necessary.
2. Construction can be as sturdy as desired.
3. The effects of moisture and temperature do not qualify the electrical effect; furthermore, with modern magnetic steel there is no ageing effect.
4. Because of the low impedance of these systems long cables may be used without shunting the output voltage from the pick-up.
5. It is necessary to integrate or differentiate only once to convert the voltage proportional to velocity to one proportional to the displacement or to the acceleration of the mechanical vibration.

Electrical Integration and Differentiation

Fig. 3 shows a simple network which will adequately perform the mathematical operation of integration and differentiation. These circuits involve the approximation that the majority of the applied voltage appears across one of the two circuit elements. In the differentiating circuit, for example, the voltage drop across R must be so small that the applied voltage closely equals that across C. This means that there is a loss in the circuit, but this can be compensated for by amplification, it also means that the useful range of the circuit is limited, because a frequency can be found at which the capacitative reactance approaches the value of the resistor. Thus the differentiating circuit will cover from zero frequency up to a limit value determined by the circuit design. The integrating circuit on the other hand, can cover all frequencies above

* *Philips Electrical Ltd.*

that for which R is several times greater than X. A similar network is incorporated in an electronic calibrator, the particular function of which is to convert the "velocity voltage" produced by the pick-up to a voltage proportional to displacement, velocity or acceleration and to measure the voltage so obtained by comparing it with one of known value. The voltage produced by the pick-up can be changed from one proportional to velocity

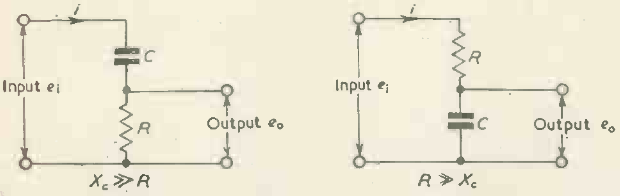


Fig. 3(a). Differentiating network ; (b) integrating network

(dx/dt) to a voltage proportional to acceleration (d^2x/dt^2) by means of a differentiator. The calibrator described also contains two integrating elements "x(freq. $>$ 10c/s)" and "x(freq. $>$ 1c/s)" the second circuit satisfactorily integrating from a frequency ten times lower than that which can be handled by the first element. The differentiating circuit is such as to give a very low phase and amplitude error for all frequencies below 1 000c/s per second. The unit provides for a comparison output between 1mV and 25V peak value and this is obtainable from an accurately calibrated potentiometer. The voltage from the potentiometer is compared with the output voltage from the pick-up and the peak amplitudes of the specific vibration component, i.e. x, dx/dt, d^2x/dt^2 can be determined.

Frequency of the vibration can be measured by feeding a suitable known but variable frequency into the calibrator unit and comparing the waveform of the known frequency with the oscillogram produced by the vibration; though this is practicable only when the vibration has a simple waveform.

Fig. 4 illustrates the seismic element of a simple type of vibrograph. The base A is attached to the structure whose vibration it is required to measure, and means are included for recording the amplitude of the relative displacement between A and the mass m.

The electronic form of this instrument Fig. 5 is a seismic system, comprising two coils suspended on spring membranes. The coils move with respect to a permanent magnet fastened to the housing of the pick-up. Spring sag due to gravity is kept within reasonable limits. The relation $d = 25/f_o^2$ exists between the sag d of the spring due to gravity and the resonant frequency f_o of the spring mass system. So that at, say, $f_o = 12$c/s, sag due to gravity $= 25/144 = 0{\cdot}174$cm. This in practice means that such a pick-up can be used in any desired position.

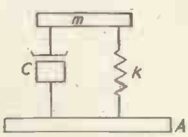


Fig. 4. Seismic element of simple vibrograph

The movable mass which moves in the housing of the pick-up is unrestricted by friction, hence the lowest limit of the intensity of measurable vibrations is determined by the sensitivity of the indicating instrument. Because the internal relative movement of a low tuned pick-up is practically equal to the absolute movement applied to the housing, such a pick-up is suitable for the measurement of very weak vibrations. Instruments of this type being velocity conscious, the conversion sensitivity is expressed in terms of voltage/velocity or volts/centimetre/second.

When using a seismic system for the measurement of absolute vibrations it is considered that half critical damping is to be preferred. The advantage being that the value of the relative movement immediately above f_o is approximately equal to that of the absolute movement to be measured, and, that in the case of impulse phenomena, the vibration at the f_o of the mechanical oscillatory system rapidly dies off. The disadvantage of the system lies in the fact that (a) there is a deterioration in the phase relationship between the relative movement of the pick-up elements and the absolute movement of the object as a function of frequency (where this phase relationship is important it is usual to apply a correction factor); and (b) when firmly affixed to the object for measurement the precise measuring range of the instrument extends over some hundreds of cycles per second, but as the frequency increases beyond about 1 000c/s, the inertia of the seismic systems begins to qualify the output from the pick-up and measuring accuracy decreases.

In order that the maximum degree of accurate representation of mechanical vibration can be maintained, it is essential that the source of vibration is not materially affected by the loading of the pick-up. The pick-up here described has a mass of about 600 grams and in the majority of cases of mechanical engineering would not prove an objectionable load; the pick-ups being individually calibrated from about 50c/s upwards, specific differences in f_o and damping—due to various tolerances—have no influence on the measuring accuracy obtainable. In the range below f_o where the phase and amplitude relations between absolute and relative internal


Fig. 5. Modern seismic type pick-up

movement depend to a great extent upon the frequency, it is preferable that the instrument be used for purely sinusoidal investigations. For vibrations measured in the longitudinal axis of the pick-up, the maximum permissible amplitude of displacement, velocity and acceleration as a function of frequency is represented by the uppermost limit lines on the three measuring range diagrams shown in Fig. 6. The broken line curves show the safe values for short duration measurement (horizontal 4mm, vertical 1·5mm), while the full lines represent the maximum values permissible on continuous loads. The lower limit lines indicate the maximum measurable x, dx/dt, and d^2x/dt^2 amplitudes when using the pick-up with an electronic integrator and differentiator, and correspond to 1mm amplitude after the electronic circuits, i.e. the electronic calibrator. The direction of vibration may easily be determined as the sensitivity of the pick-up decreases proportionally as $\cos\beta$ where β is the angle which the longitudinal axis of the pick-up makes with the direction of vibration. Consequently, in the direction perpendicular to the axis of the pick-up the sensitivity is virtually nil.

The electro-mechanical pick-up shown in Fig. 7 is intended primarily for carrying out very accurate measurements of relative oscillatory movements between objects but can also be used for the measurement of absolute vibrations, then having some advantages over the seismic pick-up. A third application of this same pick-up is the

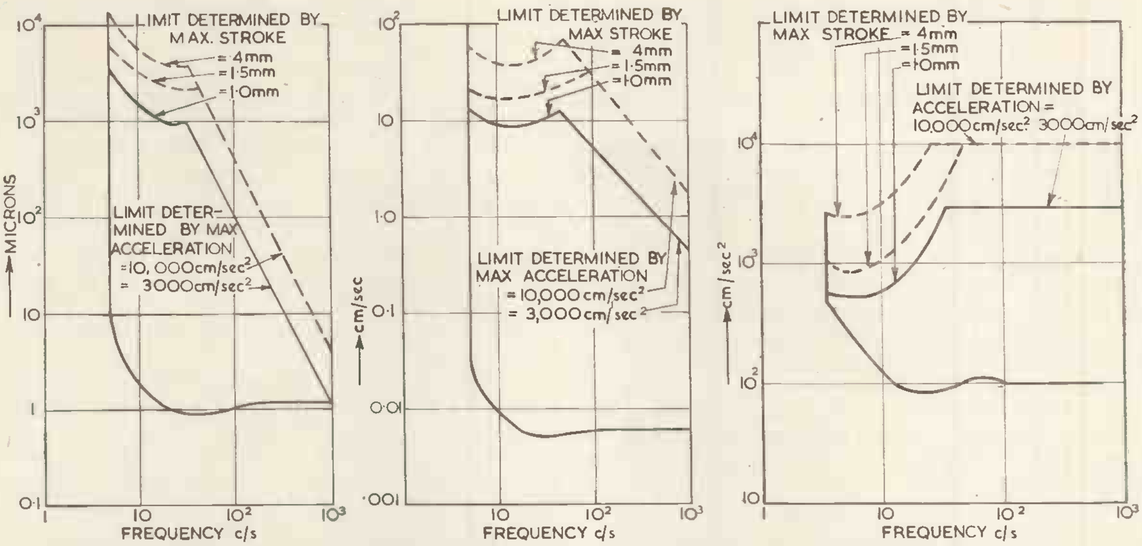


Fig. 6(a). The maximum permissible and minimum measurable amplitudes of displacement as a function of frequency when using the combination of seismic pick-up with the electronic calibrator. (b) The maximum permissible and minimum measurable amplitudes of velocity as a function of frequency using the same combination. (c) The maximum and minimum measurable amplitudes of acceleration as a function of frequency using the same combination

generation of alternating forces for measuring purposes.

The mechanical poles of the pick-up consist of a steel housing from which projects a contact pin. Mounted in the housing is a permanent magnet, while the contact pin, which is suspended by a spring, has a coil attached which can be made to move in the field of the magnet. The relative motion of the two mechanical poles in the longitudinal axis of the instrument generates in the windings of the coil a voltage proportional to the velocity of the vibratory motion. As the contact pin is connected to the coil assembly by means of a ball and socket joint, the object under test is permitted to vibrate freely in any direction, the electro-dynamic system being sensitive only to the vibration component that occurs in the axial plane of the pick-up. The spring suspension of the coil system allows no bearing friction and the damping condition of the vibrating object is hardly affected. Although the two mechanical poles are not completely independent of each other their coupling is negligible. When using this pick-up for the indication of relative movement the housing is attached to one object while the contact pin, pressed in about 2mm, is held against the other vibrating object with a pressure of about 850 grams. The total mass of the moving system is only 10 grams and can follow relative movement faithfully, provided the acceleration of that movement does not exceed 85 times the acceleration due to gravity, i.e. 850 grams force divided by 10 grams mass; though the pick-up will not suffer damage if this figure of $85g$ is exceeded its measuring accuracy will be less. The conversion sensitivity is sufficiently high to allow for the measurement of very weak vibrations and at higher frequencies much stronger vibrations can be measured than with the seismic form of pick-up. The contact frequency f_c determined by the combined resilience of the contact pin, test object and mass of the moving system is high for metallic objects—about 3 000c/s to 4 000c/s—and this is mainly due to the small mass of the moving system. The housing of the pick-up is rigidly attached to the stationary object in such a way that the f_c of the spring mass and of the housing is of the same order. Vibrations with frequencies up to 1 000c/s can be measured with accuracy. Due to the pick-up characteristics within this measuring range there is no phase shift and the pick-up can thus easily be used for the study of phase relations in vibrations.

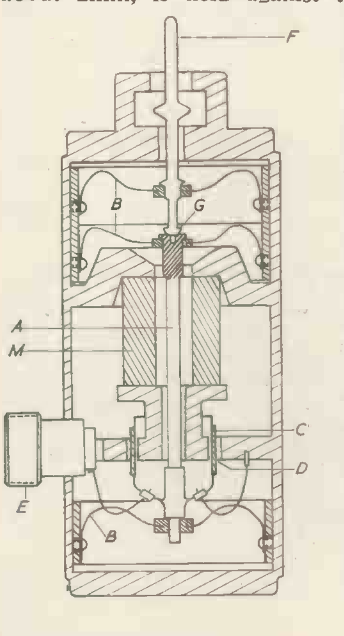


Fig. 7. Cross-section of electro-mechanical pick-up (relative)

M, permanent magnet. A, spindle of the moving system suspended in the casing by means of three springs B; to this spindle is affixed the coil C which moves in the air-gap D. E, connecting cable socket. The spindle A is connected to the contact pin F by means of a ball joint G, which allows small transverse movements of the probe tip.

Since for the measurement of absolute vibrations the housing of the pick-up must be considered as a mass stationary in space, e.g. held in the hand with the contact pin against the object being measured and pressed to a depth just greater than half the length of its full free stroke, involuntary arm movements will, at low frequencies, naturally affect measurement of small displacements but will have no effect when considering velocity nor when considering acceleration. The velocity sensitivity is adjusted accurately and, over the full free stroke of the moving system (± 2mm from the centre position), deviates only 1·2 per cent from the mean value.

The maximum acceleration that can be followed is, to within ± 1 gram, equal to the spring tension divided by the mass of the moving system, i.e. $85g \pm 10$ per cent.

The maximum stroke in the longitudinal direction together with the maximum acceleration that can be followed by the pick-up determines the upper limit of the measuring range (Fig. 8). The broken lines represent the permissible values for loads of short duration, the full lines indicate permissible values of loads for long periods. The lower limit lines or base lines indicate respectively the minimum measurable x, dx/dt and d^2x/dt^2 amplitudes corresponding to 1 millivolt out of the calibrator.

Under conditions where there is an absence of resonance the pick-up introduces no phase change. As to the phase-angle at the working frequency f, the contact frequency being f_c, the phase relationship between the absolute movement y of the coil and the absolute movement x of the object under test is given by:—

$$\phi yx = \text{arc tan}\ \frac{-\,n^3\delta}{1 - n^2\,(1 - \delta^2)}$$

where $n = f/f_0$ and δ the damping coefficient of the contact resonance system.

For $n < 1/3$ the phase-angle ϕ appears to be negligible so that on a cathode-ray oscilloscope non-sinusoidal phenomena are reproduced within the frequency range meeting the condition $n < 1/3$.

Mechanical Impedance of the Pick-ups

The two mechanical poles of the instrument are masses of 1 000 and 10 grams respectively. Between these two masses lies the electro-dynamic conversion element and in addition, a spring with an inertia constant smaller than 150 grams/centimetre.

The electrical load on the coil and the internal friction in the springs produce a slight mechanical damping between the poles. The value of the electro-dynamically introduced damping

$$= r_d = 9{\cdot}12 \times 10^5\ R_b\ \text{dynes/cm/sec.}$$

(1 dyne = 1/981 grams of force).

$r_d = 930/R_c$ grams per cm/second, where R_c is the resistance of the electrical circuit containing the coil.

With the contact pin resting directly against a vibrating object the result of a sinusoidal transverse movement of the object is a vibratory movement in the longitudinal axis of the pick-up in accordance with

$$y\ \delta/50\ \hat{x}\ \sin\omega t - \hat{x}/200\ \cos 2\omega t.$$

where y = resultant movement in the longitudinal axis (instantaneous value in mm)

δ = deviation in mm of the top of the contact pin in position of rest from the centre line of the instrument.

$\hat{x}$ = displacement amplitude in mm of the transverse movement.

$\omega = 2\pi f$ (where f is the transverse frequency). From the equation it can be seen that the sensitivity of the pick-up to transverse vibrations is very slight. It follows that if $\delta = 0$, i.e. if the rest position of the pin lies in the centre line of the pick-up, the resultant transversal vibration is of the second harmonic. If the displacement is 0·5mm and the deviation 0·5mm then the amplitude of the resultant in the fundamental is 1/200mm and in the second harmonic 1/800mm.

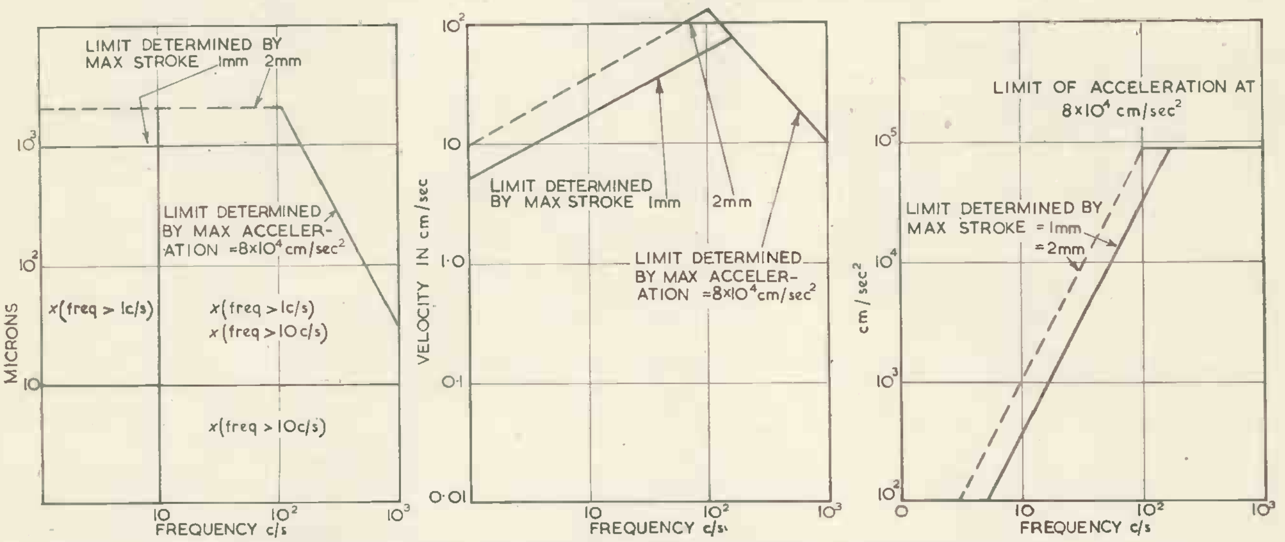


Fig. 8(a). The maximum permissible and minimum measurable amplitudes of displacement as a function of frequency using the "relative" type pick-up and the electronic calibrator. (b) The maximum permissible amplitudes of velocity as a function of frequency when using the combination as in (a). The base line of this diagram corresponds to 0·005cm/sec and gives the minimum measurable velocity. (c) The maximum permissible amplitudes of acceleration as a function of frequency when using the same combination as in (a) and (b). The base line of the diagram corresponding to 100cm/sec/sec gives the minimum measurable acceleration

total resistance in the coil circuit in ohms. The inevitable purely mechanical losses in the individual pick-ups are of the order of r_d 125 dynes/cm/sec.

The mechanical impedance of a vibrating system may be defined as the force required to give the system a certain velocity thus $Z_m \approx \dfrac{\text{force}}{\text{velocity}}$. The reciprocal value of $Z_m \approx \dfrac{\text{velocity}}{\text{force}}$ gives a value for mobility. Greater mobility therefore corresponds to small mechanical impedance. Provided the frequency of the vibration is less than $\frac{1}{3}$ the contact frequency, the mechanical impedance is determined by the formula:—

$$Z_m = r + j\,(\omega m - c/\omega)\ \text{dynes/cm/sec. in which:—}$$

ω = 2π times the frequency of the vibration.

m = mass in grams of the moving system.

c = dynamic spring constant, the value lies between $1{\cdot}5 \times 10^5$ dynes/cm.

$r = 9{\cdot}12 \times 10^5/R$ + approx. 125, where R is the

For the prolonged use of the pick-up the maximum tolerable amplitude of the transverse vibrations is 0·5mm and for short periods 1mm. If the amplitude of transverse vibrations exceeds these values the use of an extension pin would reduce the influence of transverse vibrations by a factor of about 3.

Properties of the Pick-up used as a Vibration Exciter

The mechanical-electrical effect of an electro-dynamic system is reversible, when the current is passed through a coil a mechanical force in phase with that current and proportional thereto is set up between the moving system and the magnetic structure. It can be easily deduced that the force in dynes per ampere of current through the coil is 10 times the velocity sensitivity ($\alpha x = 0{\cdot}302$V/cm/sec) all quantities expressed in identical values.

x = displacement $= x$

$\dot{x}$ = velocity $= dx/dt$

$\ddot{x}$ = acceleration $= d^2x/dt^2$

The coil of the pick-up can safely take a current of

20mA R.M.S. without danger of overheating. The peak value of the force then amounts to:—

$$20 \times 10^{-3} \times 0{\cdot}302 \times 10^{7} \times \sqrt{2} = 8{\cdot}54 \times 10^{4} \text{ dynes} = 87{\cdot}1 \text{ grams.}$$

When the casing of the instrument is again mounted stationary in space and the contact pin pressed against the object under test, for frequencies lower than 500c/s the same values for the mass and stiffness loads have to be taken into account as when the pick-up is used as a detector. The damping resistance added to the object by the exciter at the point of contact may, however, be somewhat greater, depending upon the internal resistance (R_g) of the alternating current generator by which the exciter is energised. The value of the damping introduced electrodynamically is:—

$$r_{dyn} = \frac{\alpha^2 x \cdot 10^7}{R_{elec}} = \frac{9{\cdot}12 \cdot 10^5}{2250 + R_g} \text{ dynes per cm/sec.}$$

Fig. 9. Some of the equipment in use

The coil resistance being about 2 250 ohms, the maximum r_{dyn} amounts to 405 dynes/cm/sec at $R_g = 0$. A high value of R_g as compared with the coil resistance is recommendable not only in order to minimize this electrodynamic damping but particularly to keep the excitation current constant when passing through an object-resonance while varying the frequency. The electromotive force proportional to the velocity, excited in the current supply circuit by the vibration of the object, is also effective as a change of impedance in the current circuit. This equivalent impedance variation (ΔZ) has the magnitude

$$\Delta Z = \frac{\alpha^2 x \cdot 10^7}{Z_m} = \frac{9{\cdot}12 \cdot 10^5}{Z_m} \text{ ohms}$$

Where Z_m represents the mechanical impedance, expressed in dynes/cm/sec of the vibration system. When a mechanical resonance is reached Z_m undergoes a certain change and the current in the coil can only be kept constant when that circuit has a high impedance with respect to ΔZ.

Since Z_m can never be smaller than the damping resistance of the instrument itself (approx. 125 dynes/cm/sec) the theoretically highest value that can be reached for ΔZ is 7 300 ohms. Of course, this is assuming that energy is not supplied to the vibrating system by any other means than the exciter. The alternating force (maximum 85 grams) is superimposed upon the static pressure exerted by the springs in the instrument. The resultant pressure is, therefore, during the half-cycles, that the system is drawn inwards, less than the static pressure, being determined by the difference between alternating force and static pressure.

Consequently, used as an exciter, the instrument can follow only about 90 per cent of the acceleration that it is sensitive to as a detector.

The maximum continuous current rating of the coil, i.e. 20mA R.M.S., cannot be applied for all frequencies, since no method of limiting increasing electrical impedance with the increase of frequency has been incorporated. In order to minimize the risk of insulation break-down in the coil at frequencies higher than 2kc/s, the current should be limited to about 13mA R.M.S. for 5kc/s and to about 7mA R.M.S. up to 16kc/s. The impedance of the pick-up at 16kc/s is approximately 7 500 ohms. This limitation may be a drawback when examining small objects, which often have very high natural frequencies, but this may in practice be offset by the fact that in such cases smaller excitation forces are required.

Definition of Terms

In practical mechanics the maximum displacement from the zero position is often referred to as "single amplitude" or "amplitude." Since these terms might lead to confusion when dealing not only with displacement but other phenomena coincident with vibration; e.g. velocity, acceleration, frequency, strength, etc. a terminology has been used which applies to physics generally and to electronics in particular.

When the alternating voltage converted from the mechanical vibration varies sinusoidally with time, the maximum deviation from zero is referred to as the "amplitude." When the change is not sinusoidal then the term "peak value" applies. Thus for distinction the following terms are used:—

"displacement amplitude"
"acceleration peak value"
"voltage amplitude."

With non-sinusoidal changes the peak value below zero may not equal that above zero, in these cases the terms "negative peak value" or "positive peak value" (of velocity, for instance) would be used. When the vibrations are not of constant magnitude, it will be possible to state for a given period of time the maximum peak value of amplitude observed during that time*. *Absolute movement* is defined as that movement with respect to a stationary point in space, in most practical cases with respect to the earth. *Relative movement* is defined as that movement existing between two specific points, for example, as between one part of a machine and another.

Movement of a point is characterized by its displacement as a function of time, and if this movement is sinusoidal then the motion may be described in terms of displacement, amplitude and frequency.

The velocity (dx/dt) of the movement is defined as the differential of displacement (x) with respect to time, and the acceleration (d^2x/dt^2) as the differential of velocity.

Acknowledgment

The author wishes to thank Messrs. Philips Electrical Ltd. for permission to publish this article, also to acknowledge the provision of facilities by that company for investigation of the subject. Thanks are also due to my colleagues, S. F. Smith and W. J. Donnelly, for their constructive criticism of the draft for this article.

BIBLIOGRAPHY

1. Hill, W. R. Electronics in Engineering. (McGraw Hill).
2. Manley, R. C. Fundamentals of Vibration Study. (Chapman & Hall, Ltd.)
3. Electronic Measuring. Vol. 2, No. 12, Vol. 3, No. 4. (Philips Electrical Ltd.)

* *One seldom finds the term "effective" or "R.M.S." values in vibration measuring techniques, although these two terms are frequently met with in electronics.*

Geiger Counter Tubes

A Production Range

By N. B. Balaam*, B.Sc., A.M.I.E.E.

THE usefulness of the Geiger counter tube as an instrument for detecting and measuring radioactivity is based upon the amplification of ionizing events within the tube into discrete pulses which are large enough to be counted and of which the frequency, in terms of counts per minute, is a measure of the strength of radiation incident upon the tube. The simplest method of measurement is obviously by direct comparison with a known field or field-producing source, provided the radiations are of the same nature. A full discussion of this aspect is given by Dr. D. Taylor[1].

The Geiger counter tubes hereafter described are gas-filled diodes of cylindrical geometry having glass envelopes which are operated commonly in the corona discharge region[2] with the wire positive and with a gas amplification of the order of 10^8. In this condition the ionization produced by a β-particle or by a secondary electron knocked out of the cathode by γ-ray bombardment creates an avalanche which spreads along the whole length of the thin central anode wire. This avalanche is terminated by the space-charge due to the formation of a positive-ion sheath around the wire and production of further avalanches is prevented by the action of a quenching agent added to the noble gas filling. During the avalanche a current flows through the resistor R in Fig. 1 and produces a voltage pulse whose amplitude depends upon the geometry and gas-filling of the tube and upon the values of R and C.

The minimum voltage at which the avalanche will spread along the anode so that all pulses have the same amplitude irrespective of the amount of primary ionization is known as the Geiger Threshold Voltage and the voltage range over which all the pulses—at a fixed voltage—are the same size is known as the Geiger Plateau. Fig. 2 indicates that the name is derived from the shape of the graph of counts-per-minute against applied voltage in this range. The slope of the line *ab* in Fig. 2 expressed as a percentage of the mean count rate is known as the slope of the Geiger plateau and may vary in good counter tubes between 0 and 0·1 per cent per volt according to their nature and quality. The deviation of the line *ab* in Fig. 2 from the horizontal is mainly due to the change in the effective volume of the tube brought about by increasing the applied voltage. The less significant effect of spurious pulses following close upon the main pulse is discussed by Wilkinson[3], who also refers to the role of the quenching agent.

The quenching agent is generally a vapour which is added to the noble gas and its main function is to ensure that secondary emission shall not take place at the cathode surface. This effect may be achieved by the production of excited states of comparatively low energy in the quenching agent itself (e.g. halogens[4] and organic quenching agents[5]) or by dissociation of its molecules (e.g. organic vapours such as ethyl formate). The reduction of the field near the wire implies that the tube is inoperative for a period known as the Dead Time[6] and partially operative during the Recovery Time[3]. Corrections necessary in assay observations are best considered in relation to the Resolving Time[1] of the complete equipment employed.

The electrical performance and quality of a Geiger counter tube are generally expressed in terms of its threshold voltage and the length and slope of its plateau. Large scale manufacture over many years has made possible very close control of these three properties and has established standards for counter tubes which are quenched with organic vapours such as alcohol or ethyl formate. Tubes of this type are illustrated at Fig. 3 and a typical tube has the following properties:

Geiger Threshold	1 000 volts
Plateau Length	200 volts
Plateau Slope	0·02 per cent per volt
Life	4 × 10^8 counts

The comparatively modern requirements for counter tubes which will operate at low voltages readily available from batteries and in portable equipments gave rise to the development of a series of tubes operating at about 400 volts. These tubes are filled with a trace mixture of halogen and argon in neon[7] at a total pressure of about 20cm mercury. Experiments with various proportions of these and other constituents are reported by Leibson and Freidman[8] and by Le Croissette and Yarwood[9].

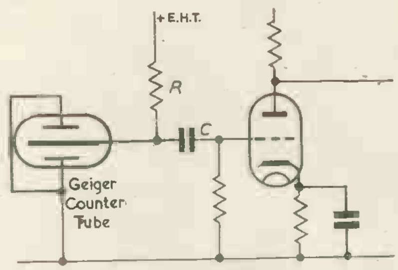


Fig. 1. Geiger counter tube circuit

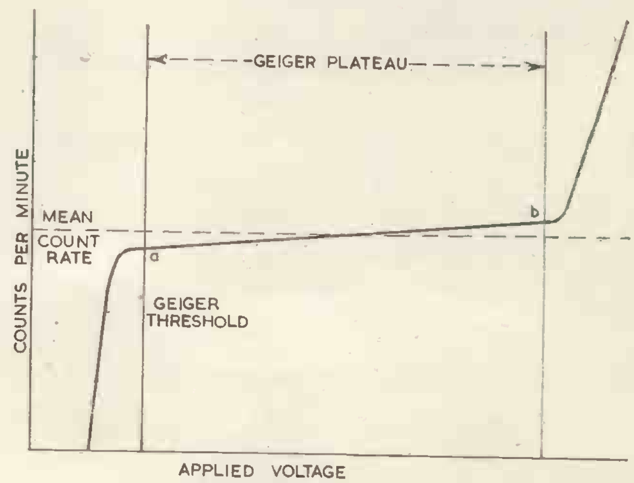


Fig. 2. Counts per minute/applied voltage

Early work upon the design of practical tubes of this type was undertaken for the Ministry of Supply and in close operation with Mr. W. R. Loosemore[4] of A.E.R.E., Harwell.

A group of halogen-quenched counter tubes is shown

* *20th Century Electronics Ltd.*

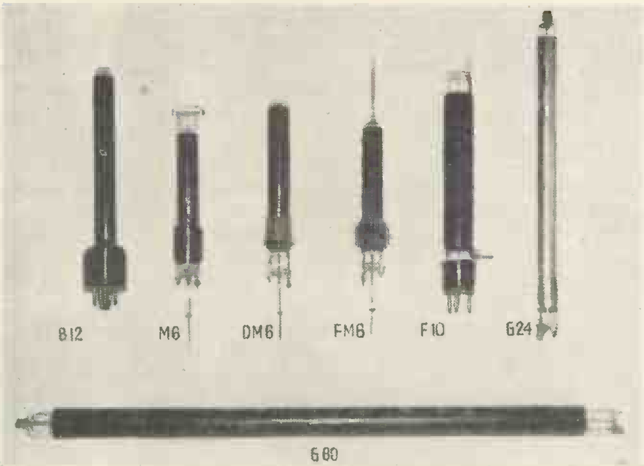


Fig. 3. Typical Geiger counter tubes

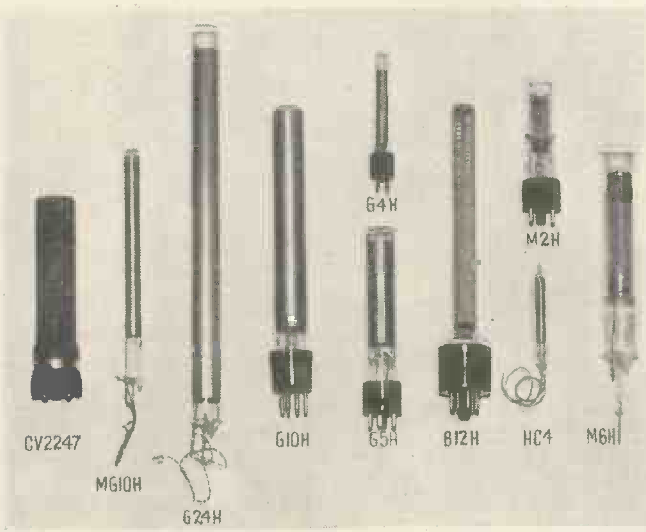
at Fig. 4 and typical properties are as follows:

Geiger Threshold	360 volts
Plateau Length	100 volts
Plateau Slope	0·02 per cent per volt
Life	Unlimited by use.

Apart from the low operating voltage, which makes a plateau length of 100 volts quite acceptable, halogen-quenched counter tubes offer the advantages of unlimited counting life, better stability and improved electrical robustness over their organic quenched counterparts. From the manufacturing point of view, however, it is necessary to employ materials which are inert to the halogens and techniques which eliminate absorption of the halogen during shelf life, and it is only the precise control of all processes that has made possible the production of consistently good halogen-quenched low voltage counter tubes in large numbers.

The influence of the circuits associated with the counter tube upon the results obtained may be considerable and has been analysed by the manufacturers[10]. In an organic-quenched counter tube the charge per avalanche is constant at a fixed voltage and the influence of R and C in Fig. 1 is quite conventional, whereas with halogen quenching the charge per avalanche cannot necessarily be regarded as independent of the impedance of the input circuit and the relationship between the pulse amplitude, circuit constants and applied voltage may be complex or even discontinuous[11].

Fig. 4. Typical halogen-quenched counter tubes



A Range of Counter Tubes

The necessity for a range of tubes is imposed by differences in the nature of the radiations characteristic of various substances. The following loose classification of these radiations illustrates some of these differences and the explanatory notes in each case refer specifically to the practical aspects of Geiger counter tube production.

(a) Hard γ quanta are of sufficient energy to penetrate glass walls of normal thickness (1 to 2mm) and produce secondary electrons with an efficiency of about 1 per cent from cathodes of metal foil or carbon.

(b) Soft γ quanta will penetrate glass walls of normal thickness but require cathode materials of high atomic weight such as tungsten or lead to produce efficiencies of the same order.

(c) Hard β-particles can penetrate walls up to 30mgm/sq.cm with no more than 50 per cent loss and are counted with 100 per cent efficiency when once in the counting volume.

(d) Soft β-particles cannot penetrate windows thicker than 3mgm/sq.cm without at least 50 per cent loss, but can be counted with 100 per cent efficiency when once in the counting volume.

TABLE I

Approximate classification of some isotopes indicating their influence on Geiger counter tube design

RADIATION	ISOTOPE	ENERGY MeV
Hard γ	Ra	0·8 average
	Co^{60}	1·33 1·17
Soft γ	I^{131}	0·4
Hard β	P^{32}	1·69
	Na^{24}	1·4
	I^{131}	0·6
	Na^{22}	0·58
Soft β	C^{14}	0·14
	S^{35}	0·11
	H^{3}	0·01

Table I indicates this classification, together with the approximate energies, of some of the isotopes in common use.

Hard γ Counter Tubes

These tubes are identified by a type number of which the first or second letter is G and the figures indicate the approximate active length in cm. A final letter H shows that a tube is halogen-quenched. In applications where large areas are to be monitored or coincidence methods are to be used tubes of the "double-ended" type such as the G24 or G60 in Fig. 3 are desirable. For prospecting and bore-hole logging and for checking contamination either in the open or indoors, tubes from the Halogen range illustrated in Fig. 4 are ideal. G24H, G10H or MG10H may be selected for lower field intensities and a halogen-quenched version of the G60, to be known as the G60H, is being developed for the very low intensities. G4H and G5H are most useful in strong fields and the latter has been protected against shock and tropical conditions by the provision of a thick rubber cover (CV.2247). The approximate measurement of fields of very high intensity may be achieved by the use of the HC4 in suitable circuits. This halogen-quenched high-current tube, illustrated at Fig. 4, may be used to give a current varying logarithmically with change of field.

Soft γ Counter Tubes

The widespread use of I^{131} in medicine, where the total amount of activity permissible is limited by health considerations, requires the best sensitivity to soft γ-radiation in a cheap and manœuvrable instrument. A thin lead foil pressed into a copper mesh forms a cathode which has the advantages of rigidity, high atomic weight and large surface area so that tubes containing such cathodes meet most of the clinical research worker's requirements. Such tubes are particularly suitable for hand and foot monitoring and for circulation investigations while the G4Pb, used in a lead "pistol" shield, has directional properties

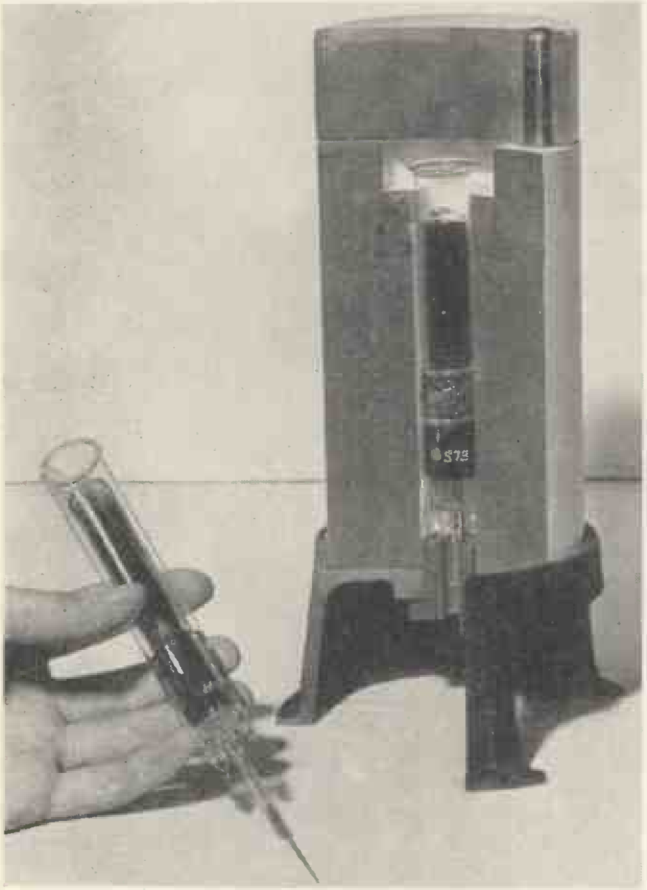


Fig. 5. Lead castle for use with M6 and M6H tubes

desirable in the detection of local concentrations[12]. These tubes are not available with halogen-quenching because of the chemical action of halogen on lead and copper.

Hard β Counter Tubes

Although these tubes are identified by the initial letter B in the type number, they are sensitive to both β- and γ-radiation and tubes such as the B12 (Fig. 3) and the B12H (Fig. 4) are commonly used for general monitoring purposes. Their sensitivity to β radiation is some 50 times that to γ, so that even in the presence of γ they may be used quite accurately for β assay.

The glass walls are reduced, over the active length, to a thickness of about 30mgm/sq.cm and the cathodes are thin films of either carbon or preferably, evaporated metal. This form gives, for example, an overall efficiency of about 50 per cent to Na^{22}. The fragility of the envelope is a limitation quickly overcome with the care taught by experience, but for applications where contamination by contact with isotope necessitates frequent cleaning and handling special arrangements have been made.

In the M6 (Fig. 3) and the M6H (Fig. 4) a 6cm thin-walled counter tube has been surrounded by a test-tube which is robust and of fixed volume. This enables a constant source geometry to be obtained with radioactive solutions and simplifies comparison measurements as well as offering a convenient form of permanent calibration.

A special lead castle for use with the M6 and M6H is shown in Fig. 5. The M6 was primarily designed in collaboration with N. Veall[13] of the Medical Research Council for use in the medical field[14].

Modified versions of the M6 are the DM6 (Fig. 3) which, by reason of its detachable outer container, is more easily cleaned and the FM6 (Fig. 3) which provides for tubular connexions to the outer container in cyclic or liquid-flow measurements. For liquid-flow experiments requiring greater sensitivity or where the available samples are necessarily small the F10 is particularly useful. This tube is illustrated at Fig. 3 and is a conventional cylindrical counter having a very thin glass tube spiral of 1cc total capacity within the counting volume. Liquid access to each end of the spiral is through capillary tubes protruding from the counter body. A further tube for use with liquids where the isotope dilution cannot be closely con-

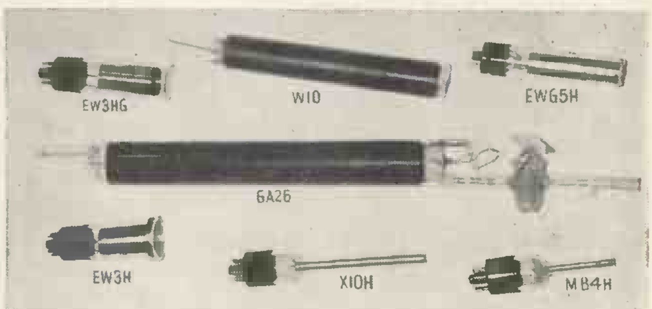


Fig. 6. Typical special purpose counter tubes

trolled, as may be the case in the field, is the M2H illustrated in Fig. 4. The outer jacket has a liquid capacity of about 5cc and the tube has a wall thickness of 15mgm/sq.cm with a 2cm active length. It is halogen-quenched and operates at 400 volts.

Recent research at A.E.R.E. Harwell has shown that the M2H may also be employed usefully for the counting of S^{35} in liquid solution with an efficiency of the order of 10^{-5}.

Where hard β-contamination is to be detected or solid samples assayed, tubes of the B12 and B12H form (B6 is a shorter version; B24 and B15H are longer) are commonly used; but when the material is in the form of dust or swarf it may be convenient to collect it on a fine wire which can be passed through a counter tube. For such purposes the W10 illustrated in Fig. 6 is ideal since it has a very thin glass tube (20mgm/sq.cm) passing straight through the counting volume of a conventional counter, parallel to the anode. Both the W10 and F10 have been developed in collaboration with A.E.R.E Harwell.

Soft β Counter Tubes

The most popular method of carrying out measurement of soft β-radiation is with an end window counter such as the EW3H illustrated in Fig. 6, which has a mica disk of about 2mgm/sq.cm sealed over the open end of a bell-shaped glass tube 30mm in diameter. The sample to be measured is spread in a specimen dish which may be presented to the window, the whole being enclosed in a lead castle to minimize the background[1]. The tube is halogen-quenched and is filled to a pressure of approximately one atmosphere, giving an operating voltage of about 650 volts.

Apart from its sensitivity to soft β-rays this type of tube has good directional properties with hard β-radiation and finds many applications in industry and medicine. It is often desirable to have a tube in a form in which it may be washed or, in cases where it is used in the operating theatre, sterilized. For these purposes[15] a tube having a glass bubble sealed directly to the open end of the "bell-jar" is available. The EW3HG, MB4H and EWG5H illustrated in Fig. 6 are all of this type and have glass windows of 5 to 10 mgm/sq.cm, according to diameter. All the end window tubes illustrated in Fig. 6 are halogen-quenched and have the advantages of uniform sensitivity across the window, absence of hysterisis, long life and a rigid anode construction enabling their use in any position.

In order to obtain the best possible sensitivity to the very soft β-radiation of C^{14} and H^{3} it is often necessary to introduce the sample into the active volume of the Geiger counter tube itself in the gas phase. Such tubes as the GA26 illustrated in Fig. 6 are manufactured in many sizes to meet varying conditions of experiment and they can be connected to the gas-producing system by cone joints, rubber hose or glass seals. Common vehicles for the isotopes referred to above are water, carbon dioxide, methane, butane; any of which may conveniently be condensed in the refrigerant tube during generation and allowed to vaporize into the active volume for counting[16,17].

X-ray Counter Tubes

The application of Geiger counting techniques to X-ray spectrometry was discussed[18] in 1948 and the advantage of a more rapid assessment of the spectra of samples than can be obtained photographically is generally accepted. A suitable tube is the Type X10H illustrated in Fig. 6, which is halogen-quenched with a vehicle gas filling of argon at about 60cm mercury pressure. Its basic design is due to Dr. Arndt of the Royal Institution. An effective length of 10cm gives 80 per cent efficiency in the counting of CuKα and the glass bubble end window is of 7mgm/sq. cm. The use of halogen-quenching gives the necessary long counting life and permits the use of a rigid anode to minimize the dead space at the window. The tube is not perfected but versions filled with krypton and xenon are envisaged to cover other energy ranges when the development is complete.

Economics

The four major groups of counter tubes, as well as those for special applications have been developed around the common constructional features of glass envelopes and cylindrical geometry. In practice, this basic design has proved to be adaptable to most Geiger counting problems and the techniques of manufacture are sufficiently flexible to provide tubes of all types at a comparatively low cost.

The "G" series of γ counter tubes with their thick glass walls are the least expensive. The "B" series of β-γ tubes, though slightly more costly on account of the fine limit thin glass wall required, find many applications in the laboratory while, in the medical field, the "M" series of liquid-jacketed tubes of this type is particularly useful. End window tubes are obviously more expensive to manufacture by reason of the special manipulative skill required for very thin windows, but they are essential to a complete range of tubes serving isotope users in medical and industrial physics.

Halogen-quenched counter tubes require a very high degree of process control to maintain good quality in production and to realise the inherently high performance of their type. Careful control and application of the new processes within the same basic glass-bodied design enables the price of these tubes to be kept low and thereby increases the popularity already established by their high stability, low operating voltage and long counting life.

Acknowledgments

The author wishes to thank R. D. Phillips and other helpers; in particular, J. Sharpe, of A.E.R.E., Harwell, for his technical advice and the correction of the early manuscript.

REFERENCES

1. Taylor, D. The Measurement of Radio Isotopes. (Methuen, 1951).
2. Freidman, H. Geiger Counter Tubes. *Proc. Inst. Radio Engrs. 34*, 791 (1949).
3. Wilkinson, D. H. Ionisation Chambers and Counters. (Cambridge University Press, 1950).
4. Franklin, E. and Loosemore, W. R. A Survey Equipment Using Low Voltage Halogen Quenched Geiger Muller Counters. *Proc. Instn. Elect. Engrs. 98*, 237 (1951).
5. Korff, S. A. Electron and Nuclear Counters. (Von Nostrand, 1946).
6. Stever, H. G. The Discharge Mechanism of Fast G-M Counters from the Dead-Time Experiment. *Phys. Rev. 61*, 38 (1942).
7. Penning, F. M. Uber den Einfluss sehr geringer Beimischungen auf die Zundspannung der Edelgase. *Zeitschrift fur Physic. 46*, 335 (1927-28).
8. Leibson, S. H. and Freidman, H. Self-Quenching Halogen-Filled Counters. *Rev. Sci. Instrum. 19*, 303 (1948).
9. Le Croisette D. and Yarwood J. The Gas-Filling and some Characteristics of Bromine-Quenched Geiger Muller Counters. *J. Sci. Instrum. 28*, 225 (1951).
10. 20th Century Electronics Ltd. Recommendations for the Operation of Counter Tubes. 1952.
11. Phillips, R. D. Notes on the Use of G.M. Tubes. *Atomics. 3*, 169 (1952).
12. Corbett, B. D. and Honour, A. J. Design of Directional Counters for Clinical Use. *Nucleonics. 9*, 43 (1951).
13. Veall, N. G.-M. Counter for Measuring the Beta-Ray Activity in Liquids and its Application to Medical Tracer Experiments. *British Journal of Radiology. 21*, 347 (1948).
14. Rose, G. and Emery, E. W. Effects of Solution Composition in a G.M. Counter for Liquid Samples. *Nucleonics. 9*, 5 (1951).
15. Barron, J. N., Veall, N. and Arnott, D. G. The Measurement of the Local Clearance of Radioactive Sodium in Tubed Skin Pedicles. *Brit. Journal of Plastic Surgery. 4*, 16 (1951).
16. Glascock, R. F. Estimation of Tritium and Some Preliminary Experiments on its Use as a Label for Water. *Nucleonics. 9*, 28 (1951).
17. Hawkings, R. C., Hunter, R. F. and Mann, W. B. On the Efficiency of Gas Counters Filled with Carbon Dioxide and Carbon Disulphide. *Canada. J. Res. 27*, 555 (1949).
18. Freidman, H. and Birks, L. S. Geiger Counter Spectrometer for X-Ray Fluorescence Analysis. *Rev. Sci. Inst. 19*, 323 (1948).

HARBOUR RADAR AT ESBJERG

At the request of the United Shipping Company of Denmark, a test was recently carried out to demonstrate the possibility of giving assistance from Harbour Radar to vessels entering Esbjerg.

A Decca Harbour Radar Type 30, arranged as part of the Decca Mobile Demonstration Unit at present touring North West Germany and Scandinavia, was placed on a site 50ft above sea level on the island of Fanö near Esbjerg. From this position a good view of the narrow approach channel could be obtained and also the rest of the channel to Esbjerg itself, a total distance of approximately eight nautical miles.

A Pye v.h.f. radio telephone equipment Type PTC 704 was installed in the chartroom of the United Shipping Company's passenger vessel "Kronprinsesse Ingrid" to provide a two-way radio link with the Harbour Radar Mobile Unit, which was equipped with a Pye Mobile Unit Type PTC 115.

v.h.f. communication was established with the "Kronprinsesse Ingrid" at a range of just over 20 nautical miles as she approached Esbjerg on her routine run from Harwich. She was detected and identified on the Type 30 Harbour Radar equipment when she was 5 miles to seaward of the first channel marking buoy. During the whole of the vessel's inward run continuous information was passed to her giving her position relative to the nearest mark and keeping her acquainted with the movement of other shipping in the channel. This information was checked on the bridge of the "Kronprinsesse Ingrid" and the Captain subsequently stated that he found the information to be reliable and that it would have been of the utmost value to him if he had been bringing in his ship in bad visibility.

Discussions are in progress to determine whether a Type 30 Harbour Radar shall be permanently installed at Esbjerg to assist the regular operation of the United Shipping Company's vessels.

The CHORD ORGAN

A New Domestic Musical Instrument

By Alan Douglas

THE Hammond organization is noted for its novel approach to electronic musical problems. Few people realize that this company has no fewer than 69 patents on electrical keyboard instruments alone, apart from many circuit patents. From this fund of ingenuity new methods of tone production are embodied in commercial instruments from time to time, and this article describes a fundamentally new concept for the large numbers of musically inclined people who have never progressed beyond the stage of picking out a melody with one finger.

The "Chord" organ automatically provides a correct harmonic accompaniment to any note so selected, and furthermore generates two bass notes which change with the accompaniment. In addition, a wide range of tone-colours embracing strings, flutes, woodwind, brass, etc., is available and octaves can be played on the melodic section of the keyboard. Vibrato, percussion, different rates of attack and decay, mutes and accenting devices are also provided.

Fig. 1. The complete instrument

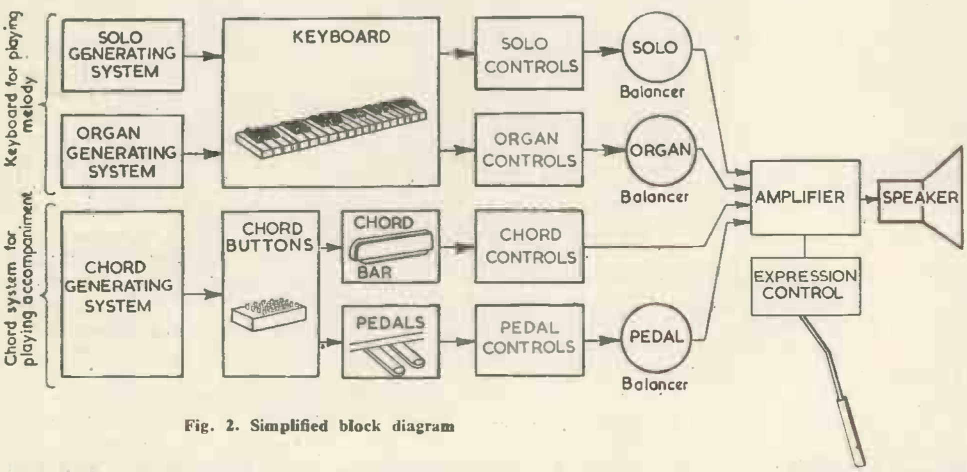


Fig. 2. Simplified block diagram

So simple and ingenious are the methods of control for all this that a special musical notation has been made available by the Hammond company, so that those who cannot read a note can play pieces so scored with ease.

Fig. 1 shows the appearance of the instrument. The case contains a 37-note keyboard from F below middle C upwards; 96 chord buttons; two bass pedals; 20 control tablets; an expression lever and an accent bar. All the generators, amplifiers and the 15-inch loudspeaker are contained in this console.

Fig. 2 is a simplified schematic diagram of the system. Two different generators feed the keyboard, another feeds the chord buttons and bass pedals. The outputs from these different sections are mixed, controlled in timbre, balanced for relative loudness and passed to the amplifier and loudspeaker.

Fig. 3 shows the chord buttons. Practically any usable chord for any note in any key is available by pressing the appropriate button. In all keys, some chords are more used, or of greater importance, than others; small detachable caps are provided to slip on these buttons, the more readily to locate them.

To play the instrument it is only necessary to press the desired note on the keyboard; pressure on the appropriate chord button will supply a harmonically correct chord and automatically pass a signal to the pedals, consisting of the fundamental pitch of the chord and, to the other pedal, the fifth above. These signals are two octaves lower in pitch than the chords. For those who can play, however, some chords can be played on the keyboard in the ordinary way. Another generator provides the signals for this. So this section can be used alone, and if the solo generator is also brought into circuit, the uppermost of any notes played will sound as a melody with a contrasting tone-colour and in several pitches. Of course, the chord buttons and pedals could be used as well.

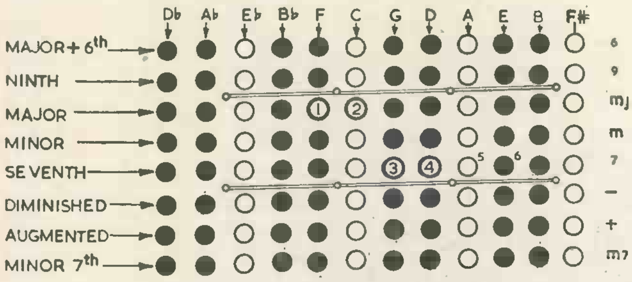
Although the resources of this instrument suggest it is complicated, this is not so, as will be clear from the following sequential description.

Keyboard

There are 37 keys, three full octaves F scale. Each key has four contacts which are depressed to touch four busbars running the length of the keyboard. The upper pair of contacts operate the melodic or solo generator whereas the lower pair excite the generator supplying notes for the rest of the keyboard. The upper of each pair of contacts is a tuning contact, which always closes before the speaking contact. All contact faces are tipped with oxide-resistant alloys but the busbars can be moved a short distance endwise to present a fresh surface or to dislodge a particle of dust. The whole of the keyboard-contact assembly is sealed in a dustproof enclosure before assembly.

Vibrato

The value of any simple tonal generator is enhanced by the judicious use of a tremulant or vibrato device. In this instrument there is a fixed frequency for this circuit, approximately 6c/s. This is found generally agreeable. A high gain triode valve is used in a phase-shift oscillator. The buffer valve or switch valve has a square wave output and its anode circuit acts as a switch to connect and disconnect small capacitors across the solo oscillator tuned circuit. There are compensating capacitors connected to the vibrato "off" tablet to maintain the mean frequency constant. The two other signal generators obtain the vibrato voltage from the cathode resistor of the switch valve because these operate by varying the grid bias of the oscillators in these sections.



BUTTON	NOTES IN CHORD				LEFT PEDAL	RIGHT PEDAL
C_6	G	A	C	E	C	G
C_9	G	A#	D	E	C	G
C_{mj}	G	C	E		C	G
C_m	C	C	D#		C	G
C_7	G	A#	C	E	C	G
C–	F#	A	C	D#	C	F#
C+	G#	C	E		C	G#
C_{m7}	G	A#	C	D#	C	G

Fig. 3. Chord buttons, showing button caps in position for locating chords in key of C and (below) the chords available in key of C

Solo or Melodic Oscillator

The oscillator drives frequency dividers providing an octave below and two octaves below the true generated pitch; therefore it is the highest pitch range which is tuned. 37 inductors are series connected so that, when the lowest note is played, the maximum inductance is such as to give the correct frequency. Depression of any other playing key then shorts out a part of the inductance chain, automatically tuning the oscillator to that note. Each inductor has an adjustable core for initial tuning. If two keys are pressed at the same time, only the upper note will speak; it is by this means that a melody of contrasting tone quality can be played when full chords are also played. Two panels of tuning capacitors in big and little steps are provided to set the pitch. A buffer valve follows the oscillator and provides a square wave output, but this waveform is used for tone forming and will not trip the dividers; to do this, a rectifier valve follows the buffer which generates waves having a steep wave-front; these feed the first divider, but are also used to reinforce the tone-forming circuits in an ingenious manner.

Frequency Dividers

Each employs three triodes, of the symmetrical feedback type. Either (but only one) of the two latter triodes can conduct at a time, for by drawing anode current it holds the other in a cut-off condition[1]. By cascading two dividers, tones one and two octaves below those of the oscillator are produced, and this is the only melodic instrument in existence (the Solovox is the same) on which this can be done.

Solo Register Controls

Three register or pitch division controls are provided; the soprano tablet derives its signals from the oscillator and its rectifier; the Tenor tablet from the first divider and the following driver valve; the Bass tablet from the second divider and a rectifier following. Thus the two incoming signals to each register control are of the same frequency but different waveshapes. After passing through suitable tone filters these signals provide woodwind tone-colours from one control and string tones from another.

Solo Tone Controls

After passing through a pre-amplifier, the signal is applied to five control circuits between the line and earth. A deep tone switch introduces a large capacitor which absorbs the upper frequencies; the full tone switch leaves the response substantially flat. First and Second voices introduce resonant inductive filters which give horn-like tones of considerable beauty; and brilliant tone produces a biting sound of the trumpet type. Thus, between the tone and register controls, a wide choice of tone-colours is available.

Solo Control Valves

It is necessary to give a suitable time-constant to the notes sounded to match the mood of the music and to prevent explosiveness in attack. The circuits for this are more comprehensive than have been made available on any instrument previously. First of all there are two keying valves in push-pull. These are normally over-biased so that the signal will not sound. The solo control contact on the playing keys earths the cathodes and so causes the note to sound. With the Solo fast attack and Solo accent tablets off, a large capacitor C_{58} makes the attack comparatively slow because a sudden decrease in the positive cathode voltage (when a key is pressed) causes a negative surge through this capacitor, charges C_{60} negatively, and so maintains the cut-off condition for an instant until the charge leaks off through R_{77}. When the Fast attack is in use, this effect is eliminated by disconnecting C_{58}. But when the Solo accent is on, it not only disconnects the capacitor C_{58} but connects a large capacitor across R_{80}. On a sudden decrease in cathode voltage, this capacitor reduces the bias for an instant, causing the note to be loud at first and so giving a percussive effect.

After these foregoing circuits, the signals pass through a balancing resistor to a pre-amplifier valve. Each section of this instrument has one of these balancers to set the relative levels.

"Organ" Oscillators

In order that chords can be played on the rest of the keyboard, a set of Hartley type oscillators is provided to give several notes each. By this means, the number of components is reduced although it means that some rather abstruse harmonies cannot be played. The upper oscillator provides three adjacent notes; the lower four, three notes each; all the others, two notes each. This permits some chords to be played and it is assumed that only two or three notes with the right-hand will be required as well as the melody, and this perhaps not often. The oscillators are tuned firstly by a main shunt capacitor across the whole coil, then for other notes by capacitors in series with the cathode tap. The coil taps are so adjusted that capacitors of the same values can be used for many purposes, thus greatly simplifying manufacture and servicing. Each coil can be tuned by adjusting the core.

There is a rather flute-like signal developed at the upper end of each coil, whereas that from the lower end of the coil is more brilliant and each may be selected by tablets. The grid resistor of each oscillator valve is taken to the top of the vibrato buffer cathode resistor; since there is a small voltage variation across this, the bias and hence the frequency of the "organ" oscillators is cyclically varied to produce a vibrato. This can be cancelled by earthing the cathode of the buffer valve. Another balancing resistor is inserted before the signals are passed to the pre-amplifier associated with this section.

similar to that of the solo section is used; pressing the chord bar removes a cut-off voltage and the signal can pass to the amplifier.

Bass Pedals

The left pedal always plays the fundamental or root note of the chord selected; the right one, the fifth above. When a pedal is pressed, the contact taking the signal to the pedal dividers closes first; then another contact operates yet one more control valve as above, with a control to allow of a fast decay to the sound. Since two stages of dividers are used in cascade (to lower the pitch two octaves), a mixed signal is available to feed the pedal tone forming filter; this provides a more satisfying tone-colour than a pure flute, which is not effective at very low frequencies. A balancer is also provided for the pedal control valve,

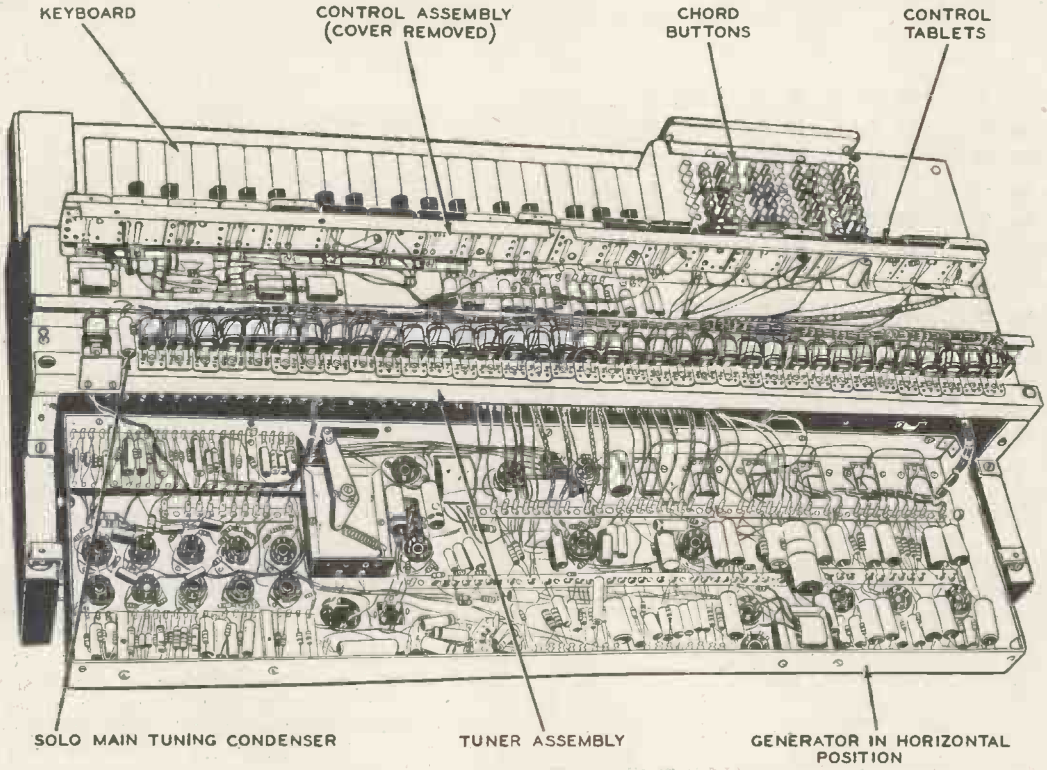


Fig. 4. Generator, tuner assembly and control assembly exposed

Chord Oscillators

The six chord oscillators are similar to those described above, and each can play either of two adjacent notes. The vibrato effect is also produced in the same way. It is possible to make the 12 chord notes available with only six oscillators, because no chord button uses more than five oscillators and no button requires two adjacent notes; however, these oscillators operate all the time, whereas the "organ" oscillators only speak when the anode circuit is keyed by one of the contacts.

Chord Buttons

Each of the 96 chord buttons selects the correct three or four notes for that chord, as well as the correct notes for the two pedals to use with that chord. The buttons tune the six oscillators as required to form the proper chords, all of which lie in the range 175 to 330c/s. For this purpose, busbars and sets of movable contacts like those of the playing keys are used; there are, in addition, extra contacts to route the signals to the pedal contact box. As the oscillators are operative all the time, a control valve circuit

but this does not attenuate the signal so much as the other sections as the ear requires a higher bass level at low volumes.

Amplifier

The main amplifier is conventional except that a differential variable capacitor is used as a volume control; this is noiseless and not subject to wear.

Power Supplies

All of these, including the several negative voltages, are derived from a common power pack. A single 5U4 supplies the H.T. circuits.

The compact nature of the electrical design will be apparent from Fig. 4, which shows the top of the instrument open; space will not permit the inclusion of the complete circuit diagram, but parts of it are given in Fig. 5.

Conclusions

The "Chord" organ is extremely small in size and characterized by the excellent mechanical construction and

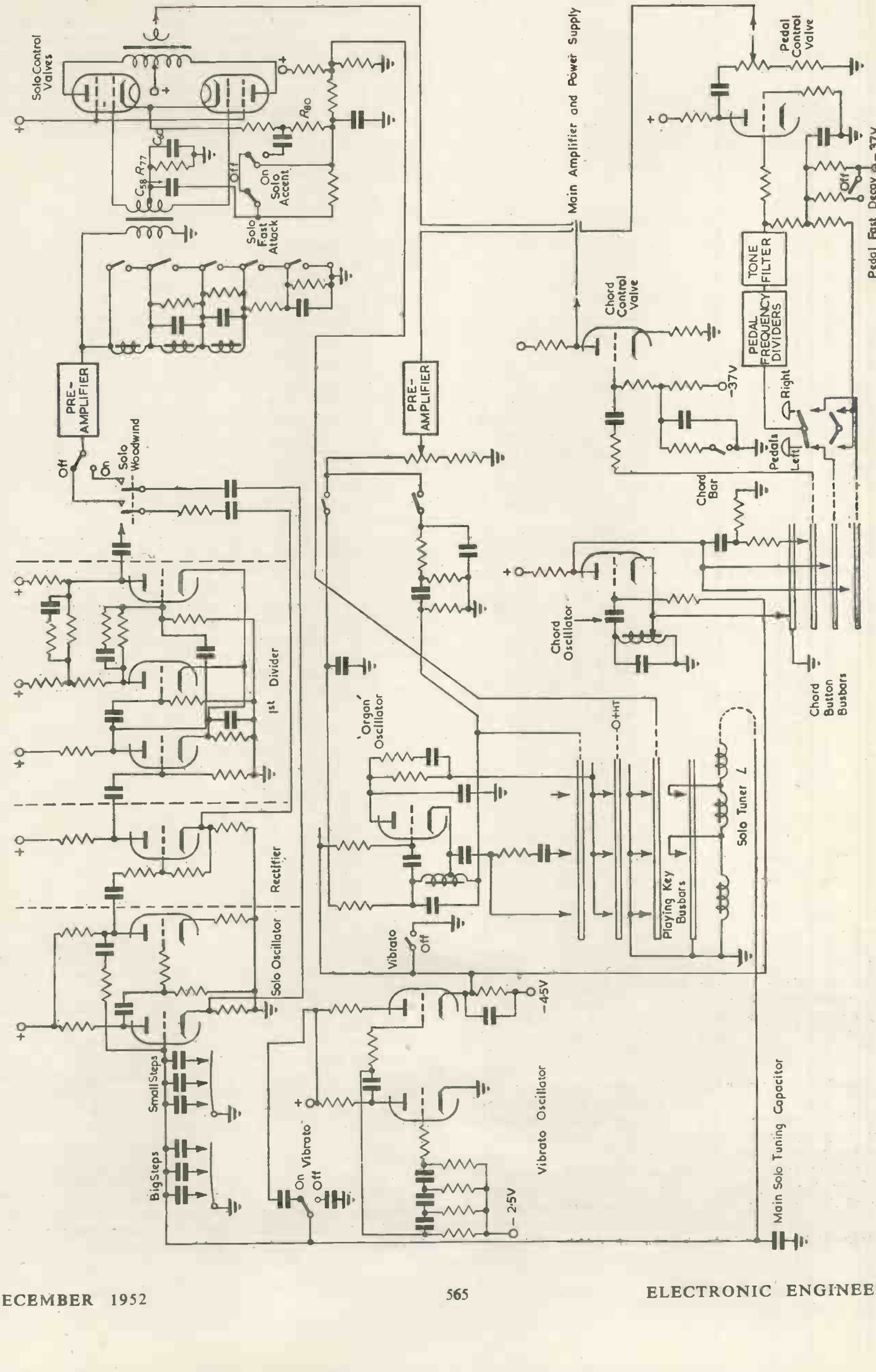


Fig. 5. Part of the circuit

finish of all Hammond products. It is only in limited production in the United States at present on account of the heavy defence programmes, and regrettably there is no prospect of its being available here. Technically it is interesting as confirming the trend towards the use of valve or static generators for producing complex musical tones, and also in showing the development of circuits for forming pleasant organ-like sounds in variety. As with all Hammond products, no attempt is made to synthesize exactly known organ tone-colours, and this undoubtedly leads to the best effect, since to meet the requirements of voiced organ pipes, more generators and other complications are required. Moreover, the kind of sounds available are more likely to be suitable for playing simple music of the light kind in the home. Independent research on extending the resources of the instrument is being carried out by the author, who has great pleasure in acknowledging the kindness of the Hammond Instrument Company in making the details available to him.

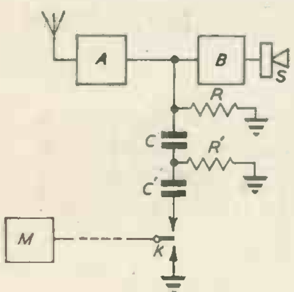
REFERENCE

1. DOUGLAS, A., The Solovox : *Electronic Eng.*, *22*, 275 (1950).

Traffic Aid for Heavy Transport Vehicles*

OF recent years road transport has been increasingly used for conveying very large objects, and consequently there are now on the roads many transport vehicles of such length that it is not easy for the drivers of these vehicles to hear motor signals from the rear of the vehicle. Besides, the engine noise in the cab of many transport vehicles gives rise to the same difficulty. Various proposals have therefore been made to enable the drivers of heavy transport vehicles to be better aware of motor horn signals made by a following car. One of the more practical proposals is that a microphone should be mounted at the rear of the transport vehicle to relay the normal acoustic motor horn signals to the cab of the transport vehicle, for instance by means of an amplifier and a loudspeaker mounted in the cab.

One difficulty that is met immediately with an arrangement of this kind is that the microphone tends to relay to the cab of the transport vehicle the general background of road noise. This of course is undesirable and it has been suggested that a 3 000c/s band-pass filter should be included in the microphone or amplifier circuit: at 3 000c/s most motor car horns have a peak in their output characteristic. If the filter is designed so as not to pass frequencies below about 2 500c/s then the majority of background noises are excluded, and in practice a simple method is to employ a high-pass filter which is transmitting for frequencies above 2 500c/s. Such a high-pass filter may be conveniently provided by means of series capacitance elements and shunt resistance elements. In some circumstances however the most convenient course may be to cause a coupling transformer to resonate at 3 000c/s so providing a band-pass circuit.



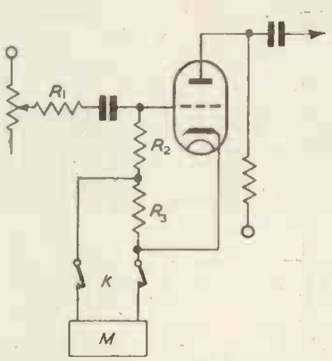


Fig. 1. Arrangement using existing radio receiver.

Fig. 2. Method of by-passing volume control.

If the transport vehicle is already provided with a radio receiver this receiver may easily be modified so that the signals from the microphone are repeated by the loudspeaker of the receiver. Fig. 1 shows an arrangement on these lines. The block A represents the carrier frequency portions of the receiver together with the detector circuit, the output from the detector is applied to an audio frequency amplifier indicated by the block B. This amplifier feeds the loudspeaker S and the resistor R, connected at one end to ground, represents the grid leak resistor of an amplifying stage of the audio frequency amplifier. This resistor is made the final shunt resistor of a resistance capacitance high-pass filter, the other elements of which are constituted by the series capacitance elements C, C' and the further shunt resistance element R'. To the input of this filter there are applied by means of the switch K the signals deriving from the microphone which is indicated by M in the diagram. In a simpler form of the filter the elements C' R' may be omitted, this omission depending upon the degree to which it is desired to discriminate against the general background.

With the arrangement just described, in the absence of any precaution to the contrary, the signals from the microphone will only be repeated in the cab of the vehicle so long as the receiver is in the operating condition; and if the receiver is one permitting only of station selection by push buttons it will clearly be possible on most occasions only to hear the motor horn signals in the presence of a broadcast programme, unless some precaution is taken so that the volume control for the receiver can be made to operate solely for the broadcast transmission leaving the audio amplifier working at full gain for microphone signals all the time.

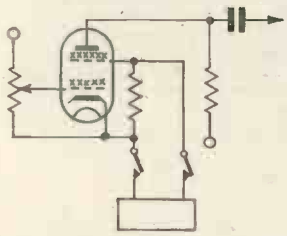


Fig. 3. An alternative arrangement.

Fig. 2 shows one arrangement in which the volume control can be worked in this way. In this figure the volume control slider is connected to the grid of the following amplifier valve not merely through the usual coupling capacitor C but also through the series resistor R_1. The leak resistance for the valve furthermore is not provided by a single resistor but by the pair of resistors R_2 and R_3 in series, to the junction point of these resistors the connexion from the microphone being effected via the switch K. By making the series resistor R' sufficiently large it is clear that the volume control slider may be set to a zero volume condition without however, short-circuiting the microphone.

Fig. 3 shows an alternative arrangement and in this a hexode valve is used and the broadcast signals from the volume control are applied directly to the inner control grid and the signals from the microphone are separately applied to the outer control grid.

* *Communicated by the Telefunken Company via E.M.I. Ltd.*

A Dekatron C.R.O. Time Marker

By J. H. L. McAuslan*, B.Sc.

THE usual way of obtaining a permanent record of a single incident phenomenon is to present it on a C.R.O. tube and photograph it. To obtain a time reference it is customary to present calibration marks in the form of a sinewave or a series of pulses on the second beam. The number of cycles or pulses per time sweep is dependent on the rate of rise of the fastest transient on the trace. When a rapid change occurs at a considerable time after the start of the time sweep, it is necessary to have a high repetition rate of marking to give significance to the time of rise of the transient, but this high rate makes it a laborious task to refer this incident to the start of the sweep by counting each individual time marker in the interval.

This measurement would be made much easier if each tenth and hundredth pulse could be separately discriminated. The instrument to be described achieves this by using scale-of-ten cold-cathode counter units. The discrimination is made possible by making the tenth pulse larger than the units and the hundredth pulse similar to the tenth but with a negative pulse superimposed. Fig. 1 illustrates the final form of the signal.

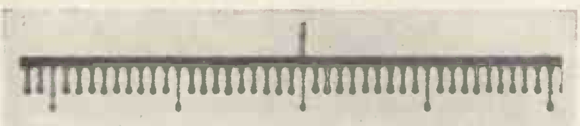


Fig. 1. Final form of the signal

Circuit Description

Fig. 2 illustrates how the Dekatron cold cathode counters type GC10B are used to obtain this. Starting with a 1kc/s crystal oscillator, the output is broken down to 100c/s by the first stage of counting. This output is then used to drive the second stage of counting through an amplifying and inverting triode.

The output from the anode of this counter is shown in Fig. 3, the large positive square pulses corresponding to the tenth counts and the thickening of the trailing edges of the other pulses to the movement of the glow from the cathode concerned to the following guide electrodes.

By cutting off the negative part of the pulses just above where this thickening occurs, the effect can be eliminated. This clipping is carried out by diode V_7, the necessary bias being obtained from potentiometer $R_{31}R_{29}$. The output-waveform is shown in Fig. 4.

* *Nobel Division, Imperial Chemical Industries Ltd.*

When this waveform is differentiated by time-constant $R_{33}C_{19}$, the output appears as in Fig. 5. The tenth pulse corresponds to the trailing edge of the large positive pulse of Fig. 3, and is consequently larger than the other nine, the relative sizes being dependent on the bias setting of R_{29}.

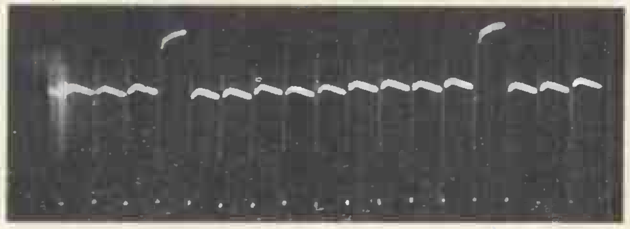


Fig. 3. Output from second stage counter

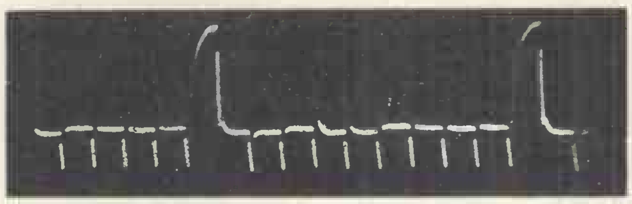


Fig. 4. Waveform after clipping

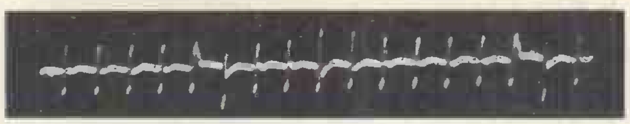
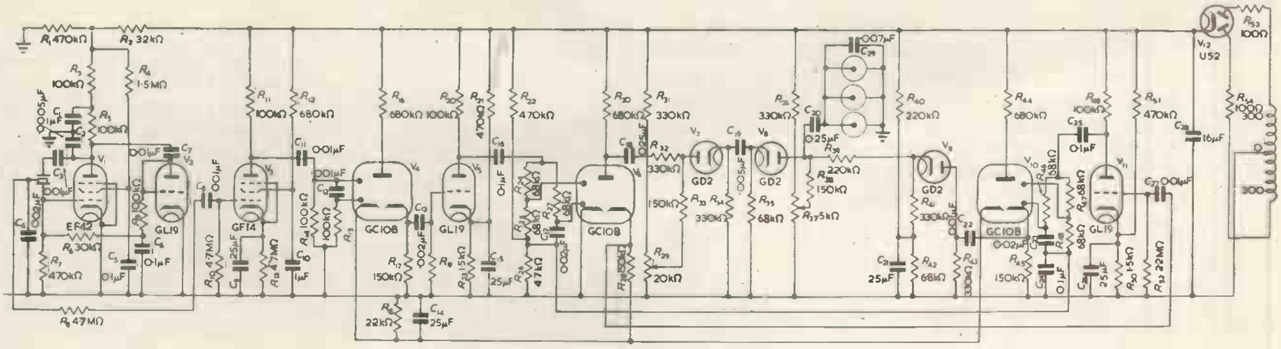


Fig. 5. Waveform after differentiation

To provide a straight base line from which measurements of displacement might be made, this output is passed through diode V_8 which has its anode biased positively. The diode therefore cuts off the erratic base line of Fig. 5 and refers the pulses to a steady potential determined by $R_{36}R_{37}$.

The third counter unit is driven through triode V_{11} in the conventional manner from the output cathode of the second counter. The output from this counter consists of a positive square wave of 100msec length with a repetition rate of one per second. As this pulse is too long for the purpose of time marking, it is differentiated by $C_{22}R_{43}$

Fig. 2. Circuit diagram



and the negative pulse from the trailing edge rejected by diode V_9. The setting of potentiometer $R_{40}R_{42}$ controls the size of this positive pulse and rejects the small variations due to the intervening nine counts of the second counter.

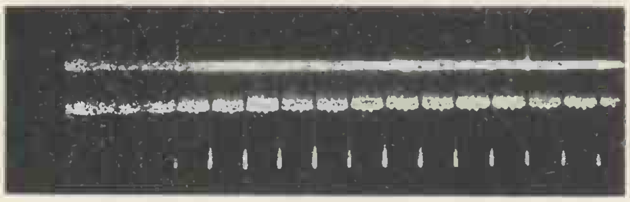


Fig. 6. Final output before integration

This pulse, recurring at intervals of one second, is then superimposed on the signal output from V_8 by feeding it on to the common load R_{38}. The resultant waveform is shown in Fig. 6. This photograph was obtained by brightening the beam with the differential of this signal, as without brightening the pulses are too sharp to register without over exposing the base line.

For practical purposes this is not convenient and the output has been integrated by C_{29}, the result being as shown in Fig. 1. This was obtained using a drum camera having a time sweep of approximately 6in. per sec, and using a steady beam brightness.

Conclusion

This particular instrument was designed for pressure-time curves having an overall period of several seconds and having a transient rise of approximately 100msec within that time. This principle can be applied to all frequencies within the range of the counter tubes available.

In addition the fundamental accuracy of a crystal oscillator has been coupled to a simple and inexpensive counter system. The use of a binary counting system would make the cost of such an instrument prohibitive.

Acknowledgment

The desirability of a long period time marker for cathode-ray tube traces was suggested by a colleague of the author, W. Nimmo.

Improved Stabilization from a Voltage-Regulator Tube

By M. D. Armitage*, B.Sc., A.Inst.P.

WITH the usual method of operating a gas-filled voltage-regulator tube in conjunction with a series resistor, a high degree of stabilization calls for a high value of resistance. This usually leads (especially if the load current is large) to an inconveniently high value of D.C. supply voltage. The circuit described in this note employs a barretter lamp in place of the series resistor, resulting in a considerable improvement in the degree of stabilization obtained for a given supply voltage. Experimental results show that the output voltage can be regulated within limits of the same order as the hour-to-hour variations to be expected in the operating voltage of the voltage-regulator tube itself. The circuit is best suited to relatively large values of load current.

Theory

The performance of a voltage stabilizer can be described by three parameters (see Hunt and Hickman[1], Gilvarry and Rutland[2], Benson[3]). These are defined as follows:—

(1) The "Stabilization Ratio", S, is the ratio in which variations in the supply voltage are reduced by the circuit.

(2) The "Range Regulation Factor", Q, is the rate at which the current in the voltage regulating element varies with the supply voltage, and enables one to estimate the range of values of supply voltage over which the stabilizer will operate.

(3) The effective "Output Resistance", R_o, may be obtained in the usual manner by considering the current due to a fictitious generator in series with the load.

In the normal circuit, where a voltage-regulating tube is connected via a series resistor to a D.C. supply, and feeds a load (assumed resistive) connected in parallel with the tube, the following expressions for the three parameters may be obtained (see Gilvarry and Rutland[2]):—

$$S = 1 + R(1/r_a + 1/R_L) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$Q = R.r_a(1/r_a + 1/R + 1/R_L) \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$R_o = R.r_a/(R + r_a) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

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where r_a is the average slope resistance of the voltage-regulator tube over the working range, R is the resistance in series with the supply, and R_L is the resistance of the load. The D.C. supply voltage required is given by:—

$$V_s = R(I_a + I_L) + V_o \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

where I_a and I_L are the currents taken by the voltage-regulator tube and the load, respectively, and V_o is the output voltage. It will be seen that a high degree of stabilization necessitates a high value of R and a correspondingly high value of V_s (unless V_o, I_a and I_L are small). In the usual case, where $R >> r_a$, the output resistance approximates to r_a.

The characteristics of a barretter lamp offer the possibility of obtaining a high dynamic series resistance with a comparatively low ohmic resistance, and it is clear from Equations (1) and (4) that this should lead to improved stabilization. A barretter lamp can be connected in place of the normal series resistor (and R in Equations (1) to (3) replaced by the average slope resistance R' of the barretter), provided the total current $(I_a + I_L)$ is within the limited range over which the lamp will operate satisfactorily. The combination of D.C. supply and lamp may be regarded as a generator of almost constant current; the small variations remaining will be absorbed by the voltage-regulator tube, and will cause only small variations in the output voltage.

Experimental Results and Calculations

Measurements were made on a barretter type 161 and on a voltage-regulator tube type STV280/80, three only of the four gaps being used to provide a convenient value of output voltage V_o. The characteristics of each tube were plotted and an average slope resistance estimated for each over its working range. In the case of the barretter, several minutes are required for thermal equilibrium to be established after a change of operating conditions, and so the slope resistance offered to sudden changes is less than that obtained from the static characteristic, which represents the resistance to slow changes. In the calculations below, the instantaneous slope resistance refers to this lower value, and is used because it relates to the most difficult type of operating condition.

The complete circuit is shown in Fig. 1.

Curves were obtained showing the performance of the voltage-regulator tube (a) with a suitable value of series resistance, (b) with the barretter, the load being approximately the same in either case. The power supply was a rectifier unit of normal design, giving about 400 volts D.C. on load; the voltage was reduced when required by means of a variac in the mains supply. The curves obtained are shown in Fig. 2, where the advantage of the barretter circuit is evident. With the barretter in circuit, the voltage changes being measured were only about 1 per cent of the output voltage, and in order to increase the accuracy of the measurement the change in voltage (consequent on a sudden change in the D.C. supply voltage) was observed directly by comparing the output voltage with a nominally equal voltage obtained from dry batteries. Changes of 0·1V could then be observed. Table 1 shows the results of such observations.

TABLE 1

LOAD RESISTANCE ..	2220Ω	1880Ω	1630Ω
LOAD CURRENT (I_L) ..	97mA	113mA	130mA
RANGE OF I_a	70-60mA	55-45mA	40-30mA
Change in V_o for 80V change in V_s			
Instantaneous ..	1·8V	1·8V	1·1V
over 30sec. ..	1·5V	1·4V	0·8V
over 2min.	1·2V	1·0V	0·2V
to new equilibrium ..	0·8V	0·4V	0·2V

The change in output voltage was measured for a given sudden change in supply voltage, the values being observed at various times from the instant of altering the supply voltage. Different values of load resistance were taken, as shown, in order to operate the voltage-regulator tube on different portions of its characteristic, and variations in

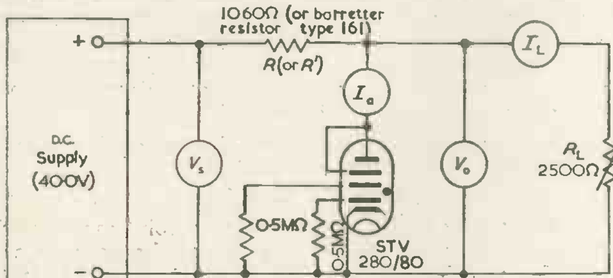

Fig. 1. The complete circuit

the slope of this characteristic account for the different results in the three columns of Table 1. It can be seen that, on the more favourable parts of the characteristic, and for slow variations of supply voltage, the constancy of the output is as good as could be expected from a voltage-regulator tube under conditions of constant input voltage. The performance of the circuit for smaller changes in supply voltage (20-30 volts) was checked; the changes in output were proportionately smaller than those in Table 1, and the stability (even for rapid variations in V_s) was approaching the ultimate stability of the voltage-regulator tube itself.

From the tube characteristics plotted (not shown here), the slope resistance r_a of the STV280/80 was estimated at 140 ohms (for the three gaps used), and the slope resistance R' (to instantaneous changes) of the 161 at 7 000 ohms (to slow variations it was as high as 20 000 ohms), compared to its maximum ohmic resistance of 1 060 ohms. This same value of series resistance R = 1 060 ohms was used with the STV280/80 to obtain curve (1) of Fig. 2. Using these values of r_a, R and R', values of the stabilization ratio S and range regulation factor Q were calculated for the two circuits, the results being shown in Table 2, together with values of S obtained from the experimental curves in Fig. 2. The agreement between calculated and experimental values is as close as can be expected in view of the degree of approximation in estimating r_a and R'. The improvement in stabilization (even to rapid changes) given by the use of the barretter is at least 5:1. The increase in the range of supply voltages over which regulation is obtained,

TABLE 2

CIRCUIT	SERIES RESISTOR	BARRETTER
S (calculated—eqn. (1))	9	55
S (Measured—Fig. 2)	8 (curve 1)	44 (curve 3)
Q (calculated—eqn. (2))	1260 ohms	7000 ohms

suggested by the values of Q in Table 2, is not realized in practice with these particular tube types. For the STV280/80 the permissible current range ($I_2 - I_1$) is 60mA, giving a range of 76 volts with the conventional circuit,

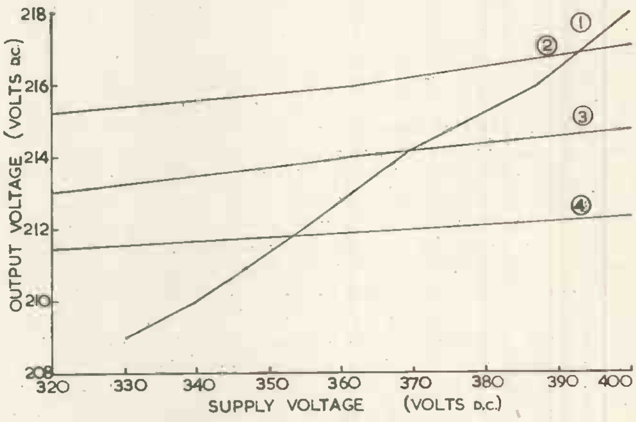

Fig. 2. Performance of STV280/80

(1) With R_S = 1 060Ω and R_L = 2 200Ω, (2) (3) and (4) with barretter and R_L = 2 220Ω, 1 880Ω and 1 630Ω respectively

using 1 060 ohms; with the 161 the range is only increased to about 90 volts, as the barretter itself cannot handle larger changes in voltage.

Conclusion

For trifling extra cost, the circuit described improves the performance of a voltage-regulator tube stabilizer by a significant amount, and with careful choice of operating points the irregularities in the burning voltage of the voltage-regulator tube can become the limiting factor. The circuit is most useful when the load current is relatively high; for lower values of load current than match a convenient barretter, a bleeder would be necessary. In any case, of course, permissible variations in the load current are restricted within a range equal to the range which can be accommodated by the voltage-regulator tube unless an adjustable bleeder is used, and the internal resistance of the complete stabilizer cannot be lower than the slope resistance of the voltage-regulator tube itself.

Acknowledgment

This circuit was initially designed to supply the current to the focusing coil of the Electron Diffraction Camera manufactured by W. Edwards & Co. (London), Ltd., while the author was on the staff of their Research Department. The author wishes to thank the Directors of Messrs. Edwards for permission to publish these results.

REFERENCES

1. HUNT and HICKMAN. Electronic Voltage Stabilizers. *Rev. Sci. Inst.* 10, 6 (1939).
2. GILVARRY and RUTLAND. The Operation of Voltage Stabilizing Elements with Current Stabilized Supplies. *Rev. Sci. Inst.* 20, 633 (1949).
3. BENSON, F. A. Voltage Stabilizers. (*Electronic Engineering*, 1950).

NETWORK ANALYSIS
By Repeated Voltage Superposition

By J. E. Parton*, B.Sc., Ph.D., M.I.E.E., A.M.I.Mech.E

IN the solution of electrical network problems, many methods have been evolved since the original statement of Ohm's law.[1] Kirchhoff's two rules[2] embodied the idea of current continuity and Ohm's law in forms quite convenient for solving networks by the branch-current method. This method seeks the current in each of the B branches of a network and requires the solution of B simultaneous equations. Maxwell[3], by the use of currents circulating round individual meshes, cut down the number of simultaneous equations to M, the least number of meshes required to specify the network. More recently its dual system, that of node potentials[4] has come into prominence and the field in which it has superiority over the mesh-current method is becoming known. It requires the solution of N' simultaneous equations, N' being the number of independent nodes in the network; (for the meaning of the term "independent node" see the paragraph including Equation 3). Lastly, the matrix methods of Kron[5,6] are coming into use, not because they have reduced the number of simultaneous equations to be solved, but because they have systematized the problem; they allow whole groups of relations to be handled with facility, and they can readily deal with networks having branches mutually coupled. There is the same computational work involved in the evaluation of inverse matrices as previously occurred in the older methods.

In addition to the development of these general methods of attack, which enable the complete voltage and current distributions throughout the network to be calculated, there have arisen many circuit theorems which enable particular problems to be solved with the minimum of computational labour. The commoner of these theorems[7] are Thévenin (Helmholtz), Superposition, Reciprocity, Delta-star and Star-delta transformations etc.

When Thévenin's theorem is first seen applied to such a problem as that of the unbalanced Wheatstone bridge, the impression is gained that the previously used mesh-currents or none-potentials, and simultaneous equations, have been replaced by the use of simple calculations requiring nothing more elaborate than a knowledge of Ohm's law and how to combine impedances in series and parallel. This is a false impression, however, and when Thévenin's theorem is applied to more complicated networks, it becomes evident that by its use an M-meshed problem is simply replaced by one with $(M - 1)$ meshes, the current in the last branch having been prevented from flowing by open-circuiting it, or alternatively by inserting a generator of just sufficient voltage in the branch. A single application of Thévenin's theorem is not of very great benefit therefore, and the question naturally arises as to whether it is possible to apply it repeatedly to a multi-meshed circuit so as to reduce the complexity to manageable bounds, thus avoiding the solution of large numbers of simultaneous equations, the evaluation of determinants or of inverse matrices.

The first difficulty is in the choice of the positions for the generators to prevent the flow of current in a sufficient number of branches, and it was in solving this difficulty that the method to be described here arose. Having found the branches in which to put the generators, and the magnitudes of their voltages, then the actual network conditions can be approached by letting one of these generators at a time, with its voltage reversed in direction, act on the remaining quiescent circuit. Finally by the Superposition theorem, all individual effects are combined to get the actual current and voltage distributions required.

After having used the method successfully for some considerable time, a search of the literature was made and this revealed that Tasny-Tschiassny[8] had in 1948 described a somewhat similar process. His method differs from the method to be described here in some details; for instance, instead of using "residual" voltage-generators in the final branches, he uses residual current-generators at the final nodes, and his arbitrary choices of current are not made in the manner used here. The "table of operations" he uses can be compared to the use of the diagrams of the superposed networks advocated here. Finally, Rogers[9] has in part used the same idea but restricts its use to simple "chains" or "ladders."

Topological Relations in Networks

The fundamental relation between branches B, meshes M and nodes N in any circuit is

$$M = B - N + 1 \qquad (1)$$

This relation has been referred to by Humphrey Davies[10] as Euler's theorem[11]; this latter theorem, however, relates to the faces, edges and vertices of a solid body and uses Equation (1) only as a step in the proof; Equation (1) possibly dates before Euler. A simple proof of Equation (1) has been given by Whitehead[12], and a more rigorous proof has recently been advanced by Synge[13].

For S separate networks considered as *one* related network Equation (1) becomes modified to

$$M = B - N + S \qquad (2)$$

In any electrically insulated network, one point can always be connected to earth, and if any one of the nodes is so chosen for earthing, there will be $(N - 1)$ "independent nodes" whose potentials are to be determined. If there are S sub-networks, one node in each can be earthed, leaving N' $(= N - S)$ independent nodes, so that generally

$$M = B - N' \qquad (3)$$

relating the meshes, branches and independent nodes.

Whitehead[12] also discusses the formation and properties of what are called "basic" networks; in these each of the nodes are formed by the junction of three and only three branches. All networks can be reduced to this basic condition by the process of opening nodes and inserting links of zero impedance. With such networks it is obvious that

$$2B = 3N \qquad (4)$$

Whitehead shows how to produce networks of M meshes from those having $(M - 1)$ meshes. The study of these basic networks reveals the field of usefulness of the mesh-current and the node-potential methods. For example, combining Equations (2) and (4), we have for basic networks

$$M = (3/2)N - N + S = N/2 + S \qquad (5)$$

The number of nodes must be even, and except for the cases where N is 2 or 4 and S is 1, the number of meshes is always less than the number of independent nodes. In

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general, therefore, the mesh-current method will be preferable for basic networks, and the node-potentials method can only be advantageous in non-basic networks in which there are several nodes formed by more than three branches.

We shall return to these topological relations after the proposed method has been illustrated by a few typical examples.

Illustrative Example

Consider Fig. 1(a). This is a planar net of 3 meshes in which all the nodes are basic, each being the junction point of three branches. In mesh 1 is a generator of voltage **E** and suppose the immediate problem is to calculate the current from the generator. Calling the area external to the network, mesh O, then the self impedances round the meshes are

$$Z_{11} = Z_{10} + Z_{12} + Z_{13}$$
$$Z_{22} = Z_{20} + Z_{21} + Z_{23}$$
$$Z_{33} = Z_{30} + Z_{31} + Z_{32}$$

The two subscripts indicate the meshes to which the impedance is common.

Suppose some generator of voltage $\mathbf{E}_1$ to be supplying the current $\mathbf{i}_1$ to the network. It is convenient to regard

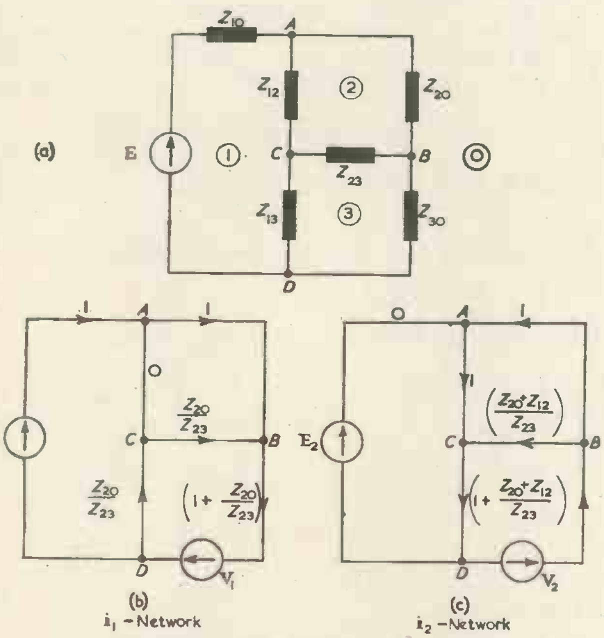

Fig. 1. Theoretical illustration of method

$\mathbf{i}_1$ as 1 unit (or 1 ampere) while the currents are being inserted in the "$\mathbf{i}_1$-network" of Fig. 1(b), remembering that all the values so found must be multiplied by $\mathbf{i}_1$ finally. At node *A* the current would divide along the branches *AB*, *AC* in some fashion not immediately ascertainable. Suppose the current in branch *AC* to be zero, so that the current *AB* is also 1 unit. There will be some conditions which can be applied to the network which would in fact force the current in *AC* to be zero; continuing through the network these conditions will emerge. If the current in *AC* is zero, then nodes *A* and C are at the same potential, and a current must flow from *C* to *B*

$$\mathbf{i}_{CB} = (\mathbf{V}_{AB}/Z_{23}) = (Z_{20}/Z_{23})$$

This current must also flow from *D* to *C*. From *B* to *D* the total current must therefore be

$$\mathbf{i}_{BD} = 1 + Z_{20}/Z_{23}$$

The current relations are now satisfied, but the voltage relations in mesh 3 are not. A source of voltage $\mathbf{V}_1$ must be included in this mesh, and applying Kirchoff's mesh rule to mesh 3 gives the relation

$$(\mathbf{V}_1/\mathbf{i}_1) = (Z_{33} \,.\, Z_{20}/Z_{23}) + Z_{30}$$

All of these relations and quantities can be indicated in the $\mathbf{i}_1$-network as the simple calculations proceed.

For mesh 1 a relation for the required voltage $\mathbf{E}_1$ can be found; it is

$$(\mathbf{E}_1/\mathbf{i}_1) = Z_{10} - (Z_{13} \,.\, Z_{20}/Z_{23})$$

Returning to branch *AC*, in which no current flows under the conditions of Fig. 1(b), in actual fact there will be some current, conveniently called $\mathbf{i}_2$, and it is the purpose of Fig. 1(c), the "$\mathbf{i}_2$-network", to determine the voltage conditions which will give rise to the current $\mathbf{i}_2$ in branch *AC*.

Suppose that the generator in mesh 1 is now replaced by one of voltage $\mathbf{E}_2$, the value of which is just sufficient to prevent the flow of current in the supply branch *AD* in the $\mathbf{i}_2$-network. In proceeding through the $\mathbf{i}_2$-network, it is again convenient to start with a current of 1 unit in branch *AC*, remembering that all the branch current values found must ultimately be multiplied by $\mathbf{i}_2$. These current values are found in the same way as before. Finally to satisfy Kirchhoff's mesh-rule, a voltage source $\mathbf{V}_2$ must be included in mesh 3 and this must be sited in the *same branch* as the first source $\mathbf{V}_1$. Hence for mesh 3

$$(\mathbf{V}_2/\mathbf{i}_2) = (Z_{20} + Z_{12})(Z_{33}/Z_{23}) + (Z_{33} - Z_{23})$$

Likewise for the mesh-1 generator, the voltage preventing the flow of current in branch *AD* is obtained from

$$(\mathbf{E}_2/\mathbf{i}_2) = (Z_{12} + Z_{13}) + (Z_{20} + Z_{12})(Z_{13}/Z_{23})$$

All that remains now is to superpose the $\mathbf{i}_1$- and $\mathbf{i}_2$-networks to produce the combined conditions in the original problem, Fig. 1(a). From this it is clear that the voltages in mesh 1 $\mathbf{E}_1$ and $\mathbf{E}_2$, must produce the supply voltage **E**; in mesh 3, the two voltages, $\mathbf{V}_1$ and $\mathbf{V}_2$, must cancel. As the voltages $\mathbf{V}_1$ and $\mathbf{V}_2$ have the opposite polarity, their magnitudes can be equated; this yields an equation for $\mathbf{i}_2$ in terms of $\mathbf{i}_1$, thus

$$(\mathbf{i}_2/\mathbf{i}_1) = \frac{Z_{20} \,.\, Z_{33} + Z_{30} \,.\, Z_{23}}{(Z_{22} - Z_{23})Z_{33} + (Z_{33} - Z_{23})Z_{23}} = \frac{Z_{20}Z_{33} + Z_{30}Z_{23}}{Z_{22}Z_{33} - Z_{23}^2}$$

For the supply voltage **E**, as the polarities of $\mathbf{E}_1$ and $\mathbf{E}_2$ are the same, $\mathbf{E} = \mathbf{E}_1 + \mathbf{E}_2$; thus, using the relation above

$$(\mathbf{E}/\mathbf{i}_1) = (\mathbf{E}_1/\mathbf{i}_1) + (\mathbf{E}_2/\mathbf{i}_2)(\mathbf{i}_2/\mathbf{i}_1)$$
$$= (Z_{10} - Z_{20}Z_{13}/Z_{23}) + (\mathbf{i}_2/\mathbf{i}_1)[(Z_{12} + Z_{13}) + (Z_{20} + Z_{12})(Z_{13}/Z_{23})]$$

After some manipulation, Equation (6) expressing $\mathbf{i}_1$ in terms of **E** and known impedances can be obtained.

$$(\mathbf{i}_1/\mathbf{E}) = \frac{(Z_{22}Z_{33} - Z_{23}^2)}{(Z_{11}Z_{22}Z_{33} - Z_{11}Z_{23}^2) - 2Z_{12}Z_{23}Z_{31} - Z_{12}^2Z_{33} - Z_{13}^2Z_{22}} \quad \ldots (6)$$

In order to check the accuracy of this result, the network can be solved by any of the other more usual methods. As this is a basic network, it is more suited to the mesh-current method. Using α_1, α_2 α_3 (to avoid confusion with the previous currents) as the mesh currents in meshes 1, 2 and 3 respectively, then in matrix notation

$$\begin{bmatrix} Z_{11} & -Z_{12} & -Z_{13} \\ -Z_{21} & Z_{22} & -Z_{23} \\ -Z_{31} & -Z_{23} & Z_{33} \end{bmatrix} \begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \alpha_3 \end{bmatrix} = \begin{bmatrix} \mathbf{E} \\ 0 \\ 0 \end{bmatrix}$$

from which

$$\alpha_1 = \mathbf{i}_1 = (\mathbf{E}/|Z|) \begin{vmatrix} Z_{22} & -Z_{23} \\ -Z_{23} & Z_{33} \end{vmatrix}$$

and this agrees with the previous analysis, as can be shown by expanding the determinants.

From the foregoing it might be wrongly concluded that the proposed method had no advantages over the more familiar ones. Its advantages are not well exhibited by such a theoretical case; even so, the evaluation of the third order determinant $|Z|$ has been avoided, the computation having been carried through by easy, obvious stages. When numerical values are given to the impedances, the advantages become more apparent, and a few such networks will now be examined.

Numerical Examples

These have been chosen to illustrate particular aspects of the method.

Network of 15 one-ohm Resistors

Fig. 2(a) shows a network in which all of the resistors shown are of value 1Ω. It is required to find the resistance measured between the input nodes AG. Connect a

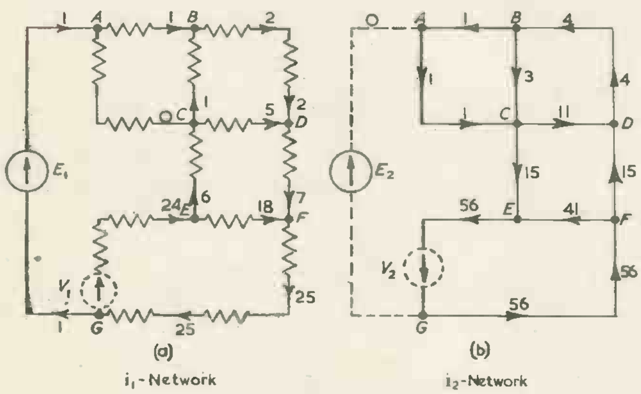

Fig. 2. Network of fifteen one-ohm resistors

the current through it also 3 units. Proceeding in this fashion round the network, the sustaining voltage required in the final mesh is

$$V_2 = (56 \times 5 + 41)\, i_2 = 321\, i_2$$

Similarly the voltage required in the supply branch to prevent the flow of current there is

$$E_2 = (2 + 15 + 2 \times 56)\, i_2 - V_2 = -192\, i_2$$

Now on superposing the two networks, the voltages V_1 and V_2 must cancel so that

$$i_2 = (141/321) = (47/107)\text{ A}$$

The driving voltage E is now obtained by summing E_1 and E_2, giving

$$E = E_1 + E_2 = 87 - 192 \times (47/107) = (285/107) = 2{\cdot}6636\text{V}$$

and since this voltage causes a current of 1A to flow into the input terminals, it is also numerically equal to the required input resistance.

The result can be checked by any of the usual methods; for instance using delta-star and star-delta transformations. In doing so the ease with which the result has been found by the above method will be readily appreciated.

A further observation can be made at this stage. There is in fact no need to determine the individual values E_1, E_2 etc. to find the supply voltage E. This can be done by applying Kirchhoff's mesh-rule, *after* the component networks have been superposed. For instance, after superposing the i_1- and i_2-networks of Fig. 2, from the outer mesh,

$$E = (1+4+7+75) - (1+8+15+168)i_2$$
$$= (285/107)\text{ volts}$$

as before. This method will be followed in later examples.

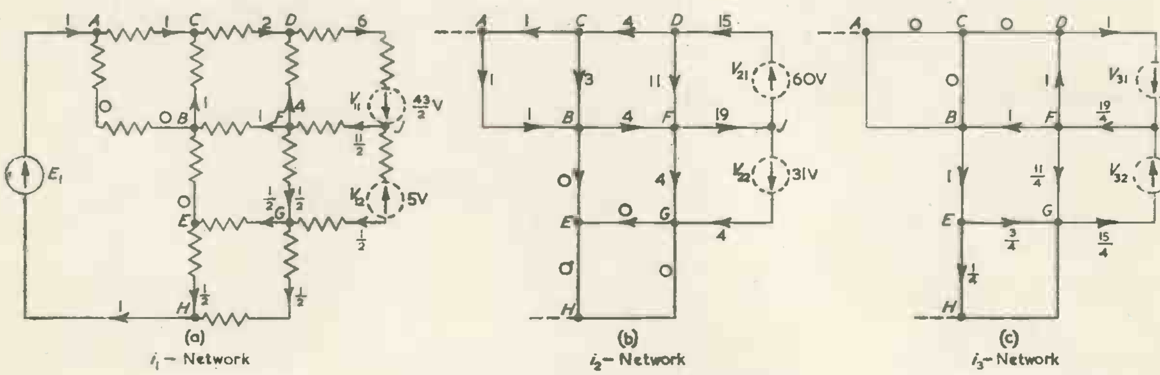

Fig. 3. Network of eighteen one-ohm resistors

generator supplying a direct current of 1A to the input terminals AG, and find the required voltage E to sustain it. The value of E will be numerically equal to the input resistance measured across AG.

In the i_1-network (Fig. 2(a)) supposing a 1A-current flowing into the node A, at A the first indeterminacy occurs. Let the current in branch AC be zero; then the voltage drop V_{AB} is 1V and so must be V_{CB}; hence I_{CB} is also 1A. Proceeding in the same manner I_{BD} is 2A and $V_{CD} = 1 + 2 + 2 = 5$V. Therefore the current I_{CD} is 5A, and so on.

In the final mesh a source of voltage V_1 is required to maintain the currents, and applying Kirchhoff's mesh-rule

$$V_1 = 25 \times 3 + 24 \times 2 + 18 = 141$$

Similarly, in mesh 1, the voltage required to inject 1A into the network, is obtained by applying Kirchhoff's mesh-rule

$$E_1 = V_1 - 24 \times 2 - 6 \times 1 = 87$$

Now turning to the i_2-network (Fig. 2(b)), taking the supply current as zero and the current in the branch AC as 1 unit, the P.D. across branch BC must be 3 units and

Network of 18 one-ohm Resistors

Fig. 3 is another network composed of square meshes having 1Ω resistors as the sides of the squares. Suppose the current distribution for a 10V D.C. supply across the terminals AH is required. Let a generator of voltage E_1 be first considered as supplying unit current (i_1) to the input of the i_1-network, Fig. 3(a). Choose the current in AB to be zero, giving $I_{BC} = I_{AC} = 1$ unit. Next choose $I_{BE} = 0$ giving the rest of the currents, and requiring *two* fictitious balancing generators, V_{11} and V_{12} of (43/2) and 5 units respectively in the branches shown.

In the i_2-network (Fig. 3(b)), starting with $I_{AB} = 1$ unit (i_2) continue as before, again choosing $I_{EB} = 0$. Again two generators V_{21}, V_{22} are required and these must be sited in the same branches as the previous pair.

Repeat for the i_3-network, starting this time with $I_{BE} = 1$ unit (i_3) and obtained the distribution shown in Fig. 3(c) in which a further pair of generators are required, V_{31}, V_{32}. The balancing generators in the supply branch, E_2 and E_3 are not shown.

Finally, superposing the three component networks, the balancing voltages in branches DJ and GJ must cancel; a third relation is obtained by applying Kirchhoff's mesh-rule to the supply mesh after the networks are superposed. These three equations are

$$86i_1 - 240i_2 + 31i_3 = 0$$
$$5i_1 - 31i_2 + 15i_3 = 0$$
$$\tfrac{1}{2}i_1 + 2i_2 + 5/4\, i_3 = E = 10\text{V}$$

Solving these for the three currents, either by elimination or by determinants, yields

$$i_1 = (13195/2711) = 4{\cdot}8672\text{A}$$
$$i_2 = (5675/2711) = 2{\cdot}0933\text{A}$$
$$i_3 = (7330/2711) = 2{\cdot}7038\text{A}$$

From these values, the current in any branch can be found. For instance suppose the current in branch DJ is required, from the individual networks we have

$$I_{DJ} = 6i_1 - 15i_2 + 1i_3 = (1375/2711) = 0{\cdot}5072\text{A}$$

Theoretical Examples

The method is not limited to numerical examples but can also be applied to theoretical studies. As illustrations, the Anderson bridge, and parallel-T networks will be considered.

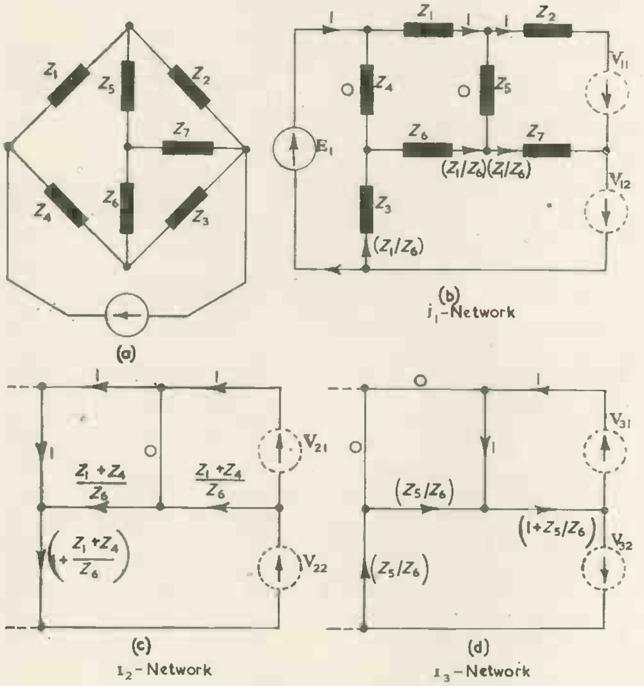

Fig. 4. Anderson bridge

ANDERSON BRIDGE[14]

Fig. 4(a) shows the usual arrangement of impedances, the suffixes corresponding to those used by Hague; for convenience, this has been re-drawn as in Fig. 4(b).

Suppose first that the bridge balance conditions only are required. Then, following the procedure already laid down, in the $\mathbf{i}_1$-network, no current is allowed to flow in Z_4 and Z_5; in the $\mathbf{i}_2$-network no current is allowed in the supply and in Z_5. For the simple bridge-balance conditions, the i_3-network is not required, and from the diagrams, the four balancing voltages are

$$\mathbf{V}_{11} = \mathbf{i}_1[Z_2 - Z_1Z_7/Z_6]$$
$$\mathbf{V}_{12} = \mathbf{i}_1(Z_1/Z_6)[Z_3 + Z_6 + Z_7]$$
$$\mathbf{V}_{21} = \mathbf{i}_2[Z_2 - Z_1Z_7/Z_6 - Z_4Z_7/Z_6]$$
$$\mathbf{V}_{22} = \mathbf{i}_2[(Z_1/Z_6)(Z_3 + Z_6 + Z_7) + (Z_4/Z_6)(Z_1 + Z_3 + Z_7) + Z_3]$$

On superposing, these voltages must cancel, i.e. $\mathbf{V}_{11} = \mathbf{V}_{21}$, and $\mathbf{V}_{12} = \mathbf{V}_{22}$. This yields a pair of equations in $\mathbf{i}_1$ and $\mathbf{i}_2$. Thus

$$(\mathbf{i}_1/\mathbf{i}_2) = [1 - Z_4Z_7/(Z_2Z_6 - Z_1Z_7)] \text{ from } \mathbf{V}_{11} = \mathbf{V}_{21}$$
$$(\mathbf{i}_1/\mathbf{i}_2) = [1 + Z_4/Z_1 + Z_3Z_6/Z_1(Z_3 + Z_6 + Z_7)] \text{ from } \mathbf{V}_{12} = \mathbf{V}_{22}$$

Hence equating

$$Z_4Z_7/(Z_1Z_7 - Z_2Z_6) = Z_4/Z_1 + Z_3Z_6/Z_1(Z_3 + Z_6 + Z_7)$$

is the balance condition and this can readily be rearranged to give Hague's Equation (9)[15]

$$Z_7(Z_1Z_3 - Z_2Z_4) = Z_2(Z_3Z_4 + Z_4Z_6 + Z_6Z_3)$$

Suppose next that the out-of-balance current $\mathbf{i}_3$ (in the detector Z_5) is required for an input current of 1 ampere. The $\mathbf{i}_3$-network, Fig. 4(d) is now also required. In this network there is no current allowed to flow from the supply, and no currents in the impedances Z_4 and Z_1. From Fig. 4(d), the further two balancing voltages are

$$\mathbf{V}_{31} = \mathbf{i}_3[(Z_2 + Z_5 + Z_7) + Z_5Z_7/Z_6]$$
$$\mathbf{V}_{32} = \mathbf{i}_3[(Z_5/Z_6)(Z_3 + Z_6 + Z_7) + Z_7]$$

On superposing the three networks, the conditions are

$$\mathbf{V}_{11} - \mathbf{V}_{21} - \mathbf{V}_{31} = 0$$
$$\mathbf{V}_{12} - \mathbf{V}_{22} + \mathbf{V}_{32} = 0$$

From these equations and inserting the value $\mathbf{i}_1 = 1$A, and rearranging, the pair of simultaneous equations become

$$\mathbf{i}_2[Z_2Z_6 - Z_7(Z_1 + Z_4)] + \mathbf{i}_3[Z_6(Z_2 + Z_5 + Z_7) + Z_5Z_7] = Z_2Z_6 - Z_1Z_7$$
$$\mathbf{i}_2[Z_3Z_6 + (Z_1 + Z_4)(Z_3 + Z_6 + Z_7)] - \mathbf{i}_3[Z_5(Z_3 + Z_6 + Z_7) + Z_6Z_7] = Z_1(Z_3 + Z_6 + Z_7)$$

From these the detector current $\mathbf{i}_3$ for a 1A supply current can be obtained by eliminating the unwanted current $\mathbf{i}_2$.

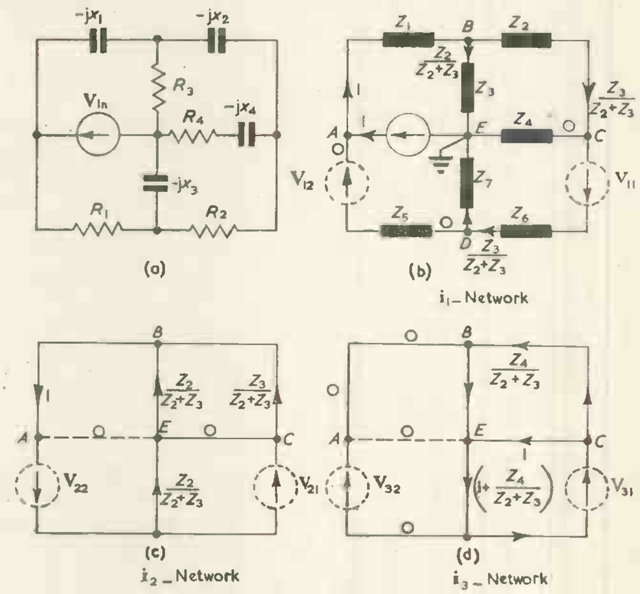

Fig. 5. Parallel-T network

PARALLEL-T NETWORK

As a final example, consider the parallel-T network which Tasny-Tschiassny[16] has treated by current residuals and a tabular process.

Fig. 5(a) shows the usual components, and these are generalized in Fig. 5(b). Three pairs of generators are found necessary, these being inserted in the branches having the impedances Z_5, Z_6. Fig. 5(b, c and d) give the current values for the $\mathbf{i}_1$-, $\mathbf{i}_2$-, $\mathbf{i}_3$-networks respectively.

Suppose it is required first to find the balance conditions, for which $\mathbf{i}_3 = 0$, and next to find the output/input voltage ratio or transfer function when the network is unbalanced.

From the networks of Fig. 5, to determine the unknown currents $\mathbf{i}_1$, $\mathbf{i}_2$, $\mathbf{i}_3$, the six injected voltages are

$$\mathbf{V}_{11} = \mathbf{i}_1[Z_3/(Z_2+Z_3)](Z_6+Z_7)$$
$$\mathbf{V}_{12} = \mathbf{i}_1[Z_1+Z_2Z_3/(Z_2+Z_3)-Z_3Z_7/(Z_2+Z_3)]$$
$$\mathbf{V}_{21} = \mathbf{i}_2[Z_3Z_6/(Z_2+Z_3)-Z_2Z_7/(Z_2+Z_3)]$$
$$\mathbf{V}_{22} = \mathbf{i}_2[Z_1+Z_5+(Z_3+Z_7)Z_2/(Z_2+Z_3)]$$
$$\mathbf{V}_{31} = \mathbf{i}_3[(Z_4+Z_6+Z_7)+(Z_6+Z_7)Z_4/(Z_2+Z_3)]$$
$$\mathbf{V}_{32} = \mathbf{i}_3[(Z_3+Z_7)Z_4/(Z_2+Z_3)+Z_7]$$

For balance conditions only the first four of these equations are required, and on superposing the networks, the conditions are $\mathbf{V}_{11} = \mathbf{V}_{21}$ and $\mathbf{V}_{12} = \mathbf{V}_{22}$ which give

$$\mathbf{i}_1 Z_3(Z_6+Z_7) = \mathbf{i}_2(Z_3Z_6-Z_2Z_7)$$
$$\mathbf{i}_1[Z_1(Z_2+Z_3)+Z_2Z_3-Z_3Z_7] = \mathbf{i}_2[(Z_2+Z_3)(Z_1+Z_5)+Z_2(Z_3+Z_7)]$$

From these, eliminating $\mathbf{i}_1$ and $\mathbf{i}_2$ and rearranging gives

$$Z_3Z_6(Z_2Z_5+Z_3Z_5+Z_2Z_7+Z_3Z_7)+ Z_3Z_7(Z_1Z_2+Z_1Z_3+Z_2Z_5+Z_3Z_5+Z_2Z_3) +Z_2Z_7(Z_1Z_2+Z_1Z_3+Z_2Z_3) = 0$$

Into this equation, the actual values of the impedances $Z_1 \ldots Z_7$ are next inserted; then equating reals and imaginaries separately to zero gives two equations

$$X_1X_3(X_1X_2-R_1R_3-R_2R_3)+R_3^2(R_1R_2-X_1X_3-X_2X_3)=0$$
$$R_3X_3(X_1X_2-R_1R_3-R_2R_3)+R_3X_2(R_1R_2-X_1X_3-X_2X_3)=0$$

Both of these equations will be satisfied when

$$X_1X_2 = R_3(R_1 + R_2)$$
$$R_1R_2 = X_3(X_1 + X_2)$$

Notice that this last pair of equations are perfectly symmetrical; both must be satisfied to give a null.

For the general unbalanced twin-T, the six-voltage equations must be used, and on superposing, the conditions are

$$\mathbf{V}_{11} - \mathbf{V}_{21} - \mathbf{V}_{31} = 0$$
$$\mathbf{V}_{12} - \mathbf{V}_{22} + \mathbf{V}_{32} = 0$$

Multiplying throughout by $(Z_2 + Z_3)$ gives two equations in the three unknowns

$$\mathbf{i}_1Z_3(Z_6+Z_7) - \mathbf{i}_2(Z_3Z_6-Z_2Z_7) - \mathbf{i}_3[(Z_4+Z_6+Z_7)(Z_2+Z_3)+Z_4(Z_6+Z_7)]=0$$
$$\mathbf{i}_1[Z_1(Z_2+Z_3)+Z_3(Z_2-Z_7)] - \mathbf{i}_2[(Z_1+Z_5)(Z_2+Z_3)+Z_2(Z_3+Z_7)] + \mathbf{i}_3[Z_4(Z_3+Z_7)+Z_7(Z_2+Z_3)]=0$$

From the networks of Fig. 5 the output voltage will be $\mathbf{i}_3Z_4$ and the input voltage will be obtained by applying Kirchhoff's mesh-rule to one of the input meshes, say round ABE, after the three networks have been superposed. Thus the transfer function is

$$\frac{\mathbf{V}_{out}}{\mathbf{V}_{in}} = \frac{\mathbf{i}_3Z_4}{(\mathbf{i}_1-\mathbf{i}_2)[Z_1+Z_2Z_3/(Z_2+Z_3)]+\mathbf{i}_3Z_3Z_4/(Z_2+Z_3)}$$

By solving the previous pair of simultaneous equations to give say $\mathbf{i}_1$ and $\mathbf{i}_2$ in terms of $\mathbf{i}_3$, and then substituting these values in the voltage-ratio equation, the currents can be eliminated leaving the transfer function ($\mathbf{V}_{out}/\mathbf{V}_{in}$) expressed solely in terms of the impedances.

It is interesting to note that in this example, by choosing the impedances to carry the zero and unit-currents differently from the foregoing, the problem could have been solved without the use of an $\mathbf{i}_3$-network. For example to do this: in the $\mathbf{i}_1$-network, $\mathbf{i}_{AB}=\mathbf{i}_1=1$, $\mathbf{i}_{CE}=0$, the current in DA can then be calculated and the residual voltage is sited in the branch DE. In the $\mathbf{i}_2$-network, $\mathbf{i}_{AB} = 0$, $\mathbf{i}_{CE} = 1$, and currents are now required in DA and in the source AE. Due to the fact that the supply current is now a linear combination of the unknown currents $\mathbf{i}_1$ and $\mathbf{i}_2$, simplicity has been sacrificed; there appears to be no saving in computational labour.

General Remarks on the Method

In the example of Fig. 1 it was only necessary to put one fictitious generator in the i_1-network. As the network complexity increased, it was seen necessary to add more generators in different branches of the i_1-network. If G generators are needed in the i_1-network, then in general there will eventually be $(G + 1)$ networks to superimpose and involving $(G + 1)$ unknowns, $i_1i_2 \ldots i_gi_{g+1}$.

Thus to determine the G currents, $i_2 \ldots i_{g+1}$, in terms of say i_1 the current from the supply, there will be G simultaneous equations to solve, and if need be these can be related back to the supply voltage by applying Kirchhoff's mesh-rule to the supply mesh.

The merit of the method lies in the fact that G is usually much less than the number of meshes M and less than the number of independent nodes N'.

So far, no general relation between G and M,N',B has been discovered. From the examples given here, and from others, as a rough working rule the following may be used. If a planar network can be considered as comprising $m \times n$ meshes, i.e. m meshes in the X-direction and n meshes in the Y-direction, then G will usually be equal to the smaller of the integers m and n. Of course many networks cannot be so classified and it appears difficult to generalise about them.

For the examples solved here, we can summarize

	M	N'	B	G
Fig. 1	3	3	6	1
Fig. 2	5	6	11	1
Fig. 3	7	8	15	2
Fig. 4	4	4	8	2
Fig. 5	4	4	8	2

It is evident from this summary that the great advantage of this method is that the number of simultaneous equations to solve, G, is always very much less than those required in methods employing branch-currents (B), mesh-currents (M) or node-potentials (N'). It is, however, important to realize that the evaluation of high-order determinants or of inverse-matrices has *not* been avoided completely, the order has simply been depressed to G, the saving being at the expense of creating subsidiary networks with additional generators, the computational labour involved in these component networks being lengthy but very simple, requiring nothing more than a use of Ohm's law and Kirchhoff's two rules.

Finally, Tasny-Tschiassny gives some theoretical treatment of the number of unknowns to be expected by his dual process in comparison with the mesh-current and node-potential methods. He concludes that if the original circuit has few generators (all the examples here had only a single generator), then the use of the " chain-relaxation " method will be an advantage.

REFERENCES

1. OHM, G. S. Die galvanische Kette mathematisch bearbeitet. (Berlin, 1827). There is an English translation of this complete paper in TAYLOR, R. Scientific Memoirs. Volume II, Article 13, pp. 401-506. (Longmans, Green, London, 1841).
2. KIRCHHOFF, G. Ueber den Durchgang eines elektrischen Stromes durch eine Ebene, inbesondere durch eine kreisformige. *Annalen der Physik 64*, 497 (1845). There is a second statement of the two rules in *Annalen der Physik, 72*, 497 (1847).
3. MAXWELL, J. C. A treatise on Electricity and Magnetism. Vol. 1, pp. 398-400. (Oxford Clarendon Press, 1873).
4. Staff, E.E. Dept. of M.I.T. Electric Circuits. (Wiley, New York, 1943).
5. KRON, G. Tensor Analysis of Networks. (Wiley, New York, 1939).
6. LE CORBEILLER, P. Matrix Analysis of Electric Networks. (Harvard University Press, Cambridge Mass., 1950).
7. Staff, E.E. Dept. of M.I.T., loc. cit. Chapter II, pp. 121-159.
8. TASNY-TSCHIASSNY, L. Network Analysis by the Chain-Relaxation Method. *J.I.E.E. 95*, Pt. 3 (Radio) 177 (1948).
9. ROGERS, F. E. The Iterated Use of Thévenin's Theorem. *Bull. Elec. Eng. Education*, No. 4, 34 (1950).
10. HUMPHREY DAVIES, M. W. Linear Network Equations. *Bull. Elec. Eng. Education*, No. 6. p. 5 (1951).
11. See any standard book on Solid Geometry as for instance HALL, H. S. and STEVENS, F. H. A School Geometry. Parts IV, V, VI, (Macmillan, London, 1922).
12. WHITEHEAD, S. Mathematical Methods Applicable to Linear Phenomena. *J. Sci. Instrum. 21*, 73 (1944).
13. SYNGE, J. L. The Fundamental Theorem of Electrical Networks. *Quart. App. Math. 9*, 113 (1951).
14. HAGUE, B. Alternating Current Bridge Methods pp. 54-56 and 379-381. (Pitman, London, 1945).
15. HAGUE, B., ibid. Equation 9, p. 56.
16. TASNY-TSCHIASSNY, L. loc. cit. p. 178, Fig. 2, Example A.

Diode Valve Voltmeter Errors

By G. D. Morgan*, Ph.D.

DIODE valve voltmeters have been used for the measurement of alternating potentials of from the order of a volt to several kilovolts, and for a frequency range extending up to a few hundreds of megacycles a second. Optimistic estimates of the degrees of accuracy of such instruments are sometimes made, and it is well to consider the magnitudes of the errors which may be involved.

The basic circuit is shown in Fig. 1. A reservoir capacitor C is charged by the rectified current through the diode to the peak voltage of the source under measurement. The steady potential developed across the capacitor is then measured with a D.C. voltmeter of high input resistance (R) so that very little power is drawn from the source. The voltmeter may be of the direct indicating type (generally a microammeter in series with a resistor, or an electrostatic voltmeter) or it may be a potentiometer arrangement. Owing to the imperfect nature of the components the steady voltage is not exactly equal to the peak of the applied voltage and the instrument is generally calibrated at a low frequency (50c/s) by comparison with a dynamometer instrument (correct to 0·25 per cent). However, other errors depend upon the frequency and may assume considerable magnitudes at frequencies of the order of 100Mc/s. It is not possible to calculate these errors accurately and the normal procedure is to design the apparatus so that they are reduced below the tolerable limit.

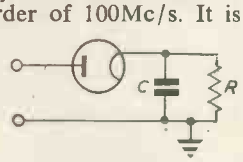

Fig. 1. The basic circuit

The author has used, over a period of years, a voltmeter of this type with a VR78 (Mazda D.1) diode. The resonant frequency of the valve and other characteristics which determine the accuracy of the instrument have been measured and the results are given in the following discussion.

Limiting Values of Capacitor (C)

The reservoir capacitor charges during part of the cycle via the valve and discharges continuously via R, so that the voltage at the cathode has a superimposed ripple at the source frequency. There is always a resistance, if only due to leakage, in parallel with C, which should be a high quality capacitor. The resistance may be found from the rate of discharge of the capacitor, for the time $t$sec for the voltage to fall from V_1 to V_2 is given by

$$t = 2{\cdot}303\, RC \log_{10} V_1/V_2 \quad \ldots\ldots\ldots\ldots \quad (1)$$

where R and C are in ohms and farads. For accuracy the fall during a cycle of the applied voltage should be of negligible amount. From (1) the value of C required so that the drop is 0·1 per cent per cycle is given very nearly by

$$C = 10^5/fR \text{ pF} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

where f is in Mc/s and R in MΩ; e.g. if R is 1MΩ (moving-coil voltmeter) and f is 10Mc/s then C must be 100pF, and 1000pF at 1Mc/s. Similarly, when calibrating the instrument at 50c/s C must be increased to 20μF.

For measurements at high frequencies it is necessary to use a physically small valve and on this account the cathode emission is generally small. Thus C should not be too large otherwise the charging time is long and the voltmeter will not follow rapid changes of the peak applied voltage. If the maximum emission current is ImA then the time taken to increase the P.D. across a capacitor CpF by V volts is $C.V/100.I$μsec. This effect would be important if a waveform of rapidly increasing amplitude were being examined with a cathode-ray oscilloscope across C.

Charging Error

In addition to the error due to the charge leaking away through R there is a further error due to the fact that the capacitor has to charge via the diode which is equivalent to a resistor. The result is that the voltage across the capacitor lags behind the applied voltage and due to the leakage it starts the next cycle at a lower value than the peak attained. In the equilibrium state C charges to a fraction k of the peak of the applied voltage $V\cos 2\pi ft$ and by equating the charge received during the conducting period to that lost by leakage over the whole period it can be shown[1] that

$$(1-k) = 2{\cdot}2(R_a/R)^{\frac{1}{3}}$$

where R_a is the resistance of the valve. R_a will vary with the voltage between the electrodes, i.e., the (I/V) characteristic will not be linear. However, since k will be very nearly unity the voltage across the diode during the conducting period will be small, so that the value of R_a as $V \to 0$ is the one required. As a reasonable approximation R_a is the value of the differential resistance at a voltage equal to half the permissible voltage error; e.g. if the error is 0·5 per cent then R_a is the value at 1/400 of the voltage to be measured. For a VR78 valve (Mazda D.1) R_a is about 500 ohms so that for the error to be <0·5 per cent R must be >5MΩ. The error depends to some extent on the voltage being measured in so far as R_a varies with the anode to cathode voltage. The effect is not serious, however, as it is independent of the frequency and is eliminated by a low frequency calibration.

Valve Characteristic Effect

Because the electrons leave the cathode with finite velocities and also due to the contact potential between the electrodes (of different materials) the diode characteristic, for small V, is

$$I = a(V_1 + V)^n$$

where a and n are constants and the two factors are included in the small voltage V_1. When the applied voltage is zero C will charge to nearly V_1 to give a constant positive error. For a VR78 valve it is found that the latter varies from 0·5 to 0·7 volts (with R 1MΩ and 5MΩ respectively and with the anode and earth connected). A 5 per cent reduction of the heater current results in a 15 per cent drop in this voltage. The effect is most serious when low voltages are being measured, but it can be eliminated by a low frequency calibration. However, the effect varies with the ageing of the valve so that the calibration should be checked at intervals.

Transit Time Error

On account of its mass an electron takes a finite time to pass from the cathode to the anode, and consequently a fraction of the electrons emitted (i.e. the later ones) will fail to reach the anode before the field reverses. On this account the capacitor will not charge to the peak voltage and the discrepancy will be greater at higher frequencies.

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No electrons leave the cathode until the instant at which the anode voltage equals that across the capacitor. For a short interval after this electrons move to the anode in the small field. If any electrons reach the anode the capacitor voltage rises and during the next cycle the excess anode voltage is less. In the steady state the first electrons emitted just fail to reach the anode. Assuming that these electrons are not influenced by space charge and that the field is uniform, Fortescue[2] has shown that the percentage error is given by $34(2\pi f)T$, where $T = 2d/[2V(e/m)]^{\frac{1}{2}}$ is the transit time corresponding to a steady voltage V (peak applied volts) between the electrodes d centimetres apart. This may be written in terms of fMc/s, $d$cm and V volts as

$$\text{percentage error} = 7{\cdot}2\, f.d/V^{\frac{1}{2}} \qquad (3)$$

For a 955 acorn valve (grid and anode connected) Nergaard[3] gives the error as $\Delta V = 30V^{\frac{1}{2}}/\lambda$, where λ is the wavelength in cm. If this is put into the form of Equation (3), assuming $d = 0{\cdot}0135$cm, then the numerical factor is found to be 7·4.

Megaw[1] points out that for parallel plane electrodes with maximum space charge the transit time (and hence the error) is 1·5 times that for the simple case, and for cylindrical electrodes with extremely small cathode and no space charge 0·5 times. The value of the numerical factor may be determined experimentally by comparing results obtained with diodes with different electrode spacings. Using a special diode with cylindrical electrodes having a spacing of 0·008cm (anode diameter 0·023cm) Nergaard[3] gives the error as $\Delta V = 7{\cdot}5V^{\frac{1}{2}}/\lambda$. Put into the form of Equation (3) the numerical factor is 3·2. For diodes each with a hairpin-shaped filament below a hemispherical anode and with electrode spacings of from 0·13 to 2·14mm, Megaw[1] finds the numerical factor to be 11·6. As a general criterion he suggests that the error is <1 per cent if $f.d/V^{\frac{1}{2}} < 0{\cdot}1$. It is seen that the error is dependent upon the frequency and the voltage being measured, and will be considerable at low voltages. For a VR78 d is 0·015cm (anode diameter 0·15cm) so that at 100Mc/s the error is less than 1 per cent provided V is greater than 225 volts; it is 3 per cent at 25 volts and nearly 7 per cent at 5 volts. For valves designed for use at greater voltages the electrode spacing may be a few millimetres and hence the errors will be proportionally greater, i.e., about 20 times those quoted for the VR78, under similar conditions.

Impedance Errors

These errors are due to the inter-electrode capacitance of the valve and the stray capacitance and inductance of the leads and connecting wires.

Owing to its input impedance the valve voltmeter will have some effect on the source undergoing measurement, so that normally it is desirable to make this impedance as large as possible. If R is small compared with the leakage resistance between the valve electrodes then the input impedance is $R/2k^2$, i.e. very nearly $R/2$. However, at radio frequencies this is generally large compared with the reactive component due to the diode and stray capacitances.

Voltage Dividing Action

The H.F. voltage is applied to the reservoir capacitor C and the diode capacitance C_d in series, so that the voltage between the electrodes is less than the applied voltage and the percentage error is $100\, C_d/C$, C_d being much less than C. This error is not eliminated by a low frequency calibration as the reactance of C_d will be large compared with the leakage resistance; however, the error will not be accentuated. For a VR78 (C_d about 2pF) and using a 1000pF reservoir capacitor the error will be negligible.

Resonance Error

The leads from the source under measurement will have a finite inductance L and together with C_d and the stray capacitance of the wiring will form a series resonant circuit. The voltage between the electrodes will then be greater than the source voltage, depending upon the nearness of the source frequency (f) to the resonant frequency of the circuit (f_r). When f is remote from f_r, so that the resistance of the leads may be neglected, the percentage of error is given by

$$100\left\{\frac{1}{1-4\pi^2 f^2 L C_d} - 1\right\} \text{ or } 100\, f^2/f_r^2$$

where C_d includes the stray capacitance. Thus for the error to be <1 per cent f must not be greater than 10 per cent of f_r.

The resonant frequency of a VR78 valve (with the cathode and heater connected) has been determined, by measuring the resonant frequencies with different lengths of shorted transmission lines attached[3], and found to be nearly 1100Mc/s ($C_d = 1{\cdot}93$pF and $L_d = 0{\cdot}011\mu$H). Hence if the diode is used as a valve voltmeter with no added leads the resonance error would be less than 1 per cent up to about 100Mc/s and just over 3 per cent at 200Mc/s. However, if in a practical circuit the added lead inductance is, say, $0{\cdot}05\mu$H (3in. of $\frac{1}{4}$in. braid) the resonant frequency is reduced to 460Mc/s and the error is now within 1 per cent only up to 46Mc/s and at 200Mc/s it is 19 per cent.

The conditions required to reduce the resonance and transit time errors are opposed, in that the latter is reduced by small inter-electrode spacing whereas the former increases with the capacitance. The resonance error is difficult to calculate, particularly at the higher frequencies when it is most important and the inductance and capacitance cannot be considered as discrete elements. However, the resonance error is positive, i.e. the indicated voltage is greater than that of the source and so its effect is to some extent reduced by the other errors which are in the opposite sense.

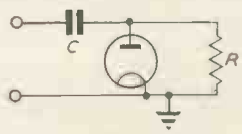


Fig. 2. The parallel circuit

Practical Circuits

The series diode circuit shown in Fig. 1 suffers from a number of disadvantages:—

(a) The heater circuit (connected to the cathode) is at a positive potential to earth, and the accumulators must be carefully insulated and screened;

(b) There must be a D.C. path for the diode current through the source under measurement or, alternatively, one must be provided by shunting with a suitable resistor; and

(c) The positive error with zero input will depend upon the resistance of the source.

These disadvantages can be overcome by using the parallel circuit shown in Fig. 2. In this case the capacitor charges so that the diode anode is negative with respect to earth and the applied waveform is reproduced across the D.C. voltmeter but with its mean level suppressed so that the positive peaks just reach zero potential, when the charge on C is maintained, and the D.C. voltmeter indicates the mean level. However, the capacitance of the resistor R now shunts that of the valve and so increases the resonance error. On this account the resistor must be placed with the valve in the probe head and be of a low capacitance type, small and not wire-wound.

Conclusion

It is seen that the errors may be divided into four classes.

(1) Errors which are eliminated by a low frequency calibration; such as those due to the resistance of the valve, the loss of charge through the D.C. voltmeter, and the fact that the diode passes current when the applied voltage is zero.

(2) Errors not eliminated by a low frequency calibra-

tion but which can be reduced to negligible magnitude by suitable design, e.g. the voltage dividing error.

(3) Errors which are not eliminated by low frequency calibration, are not negligible and must be estimated. In this case we have the transit time error, which is greater at higher frequencies and lower voltages, and also the resonance error which increases as the square of the frequency. The latter cannot be eliminated nor can it be calculated with accuracy at the highest frequencies.

(4) Finally there are those errors which though eliminated by calibration may change with time, e.g. that determined by the valve (I/V) characteristic.

In addition, there are the errors associated with the D.C. voltmeter. A high stability carbon resistor may show permanent changes up to ± 2 per cent; so that the calibration should be checked at intervals. If a direct indicating meter is used it should have a long and well graduated scale (e.g. 6in. with 100 divisions), and preferably should be provided with a non-parallax mirror. In this case the position of the pointer may be estimated to the nearest $1/5^{th}$ of a scale division, i.e. to 1/2 per cent at half-scale. Similar considerations apply to the meter used for the low frequency calibration.

REFERENCES

1. MEGAW, E. C. S. Voltage Measurement at V.H.F. *Wireless Eng.* *13*, 65, 135 and 210 (1936).
2. FORTESCUE, C. L. Thermionic Peak Voltmeter at V.H.F. *J.I.E.E.* *77*, 429 (1935).
3. NERGAARD, L. S. A Survey of U.H.F. Measurements. *R.C.A. Rev.* *3*, 156 (1938).

An Electronic Bridge-Balance

By J. W. Bayliss*, B.Sc., A.M.I.E.E.

THE problem outlined below had to be solved when a sensitive industrial metal detector caused great consternation amongst the factory maintenance staff by its continual drift out of adjustment.

The detector consisted essentially of an R.F. bridge circuit followed by a high gain amplifier and detector. In operation it was adjusted by a variable resistor and a variable capacitor to give zero output from the amplifier. Metal, when placed near a coil forming one arm of the bridge, caused a change in the effective inductance and resistance of the coil, unbalanced the bridge and indicated on the detector.

It was thought at first that the drift might be of a cyclic nature caused by temperature changes. Tests of the coil assembly in an oven showed that this was not so and attempts to correct the drift by components of an opposite thermal coefficient (e.g., negative temperature capacitors) were abandoned.

Motor-driven variable elements were then suggested, but after a few experiments, were abandoned because of their complication and expense, in favour of the method outlined below.

The circuit shows a "variable electronic reactance" V_1, and a "variable electronic resistance" V_2, controlled by phase sensitive detectors $V_3 + V_4$ and $V_5 + V_6$ respectively. Time delays R_9C_4 and $R_{10}C_7$ are inserted so that the balancing system will respond only to slow changes, e.g., thermal effects, and not to a rapid unbalance caused by the detection of a metallic body.

The reference voltage for detector V_5V_6 controlling V_2 is obtained directly from the bridge generator, which must be balanced to earth. That for V_3V_4 controlling V_1 is obtained from the generator via C_8, C_9, R_{17}, R_{18} which changes the phase by 90°. Conditions for this are that $X_C = R$ and $C_8 = C_9$, $R_{17} = R_{18}$. Conditions for satisfactory operation of V_1 are that X_{C_3} is greater than R_2, say $= 5R_2$, and similarly for V_2, $R_6 = 5R_7$ and $X_{C_6} = \frac{1}{2}R_7$.

Initial Adjustment

S_1 and S_2 are switched to "Off" and the bridge adjusted to balance by R_1 and C_1. A sensitive meter is connected between point F and earth, and the phase of the amplifier varied if necessary by detuning slightly until the movement of C_1 out of the balance position in either direction produces no reading on the meter. Movement of R_1 should give a large reading, positive when moved in one direction, negative in the other. When the meter is connected to G, operation of R_1 should give no reading and of C_1 a large reading.

With the bridge balanced close S_2. If this causes unbalance, reverse connections A and B, open S_2, close S_1; if this causes unbalance, reverse connexions C and D.

Adjustments are now completed and the automatic balance control (A.B.C.) may now be tested. Switch on S_1 and S_2. Small movements of R_1 and C_1 separately or together will be cancelled after a time interval and the balance restored.

As the coil assembly developed a permanent change in characteristics with time, it was necessary every three months or so to switch off the A.B.C., readjust R_1 and C_1 to allow for this permanent change, and switch A.B.C. on again. Apart from this attention, equipments have been in continuous operation for two years.

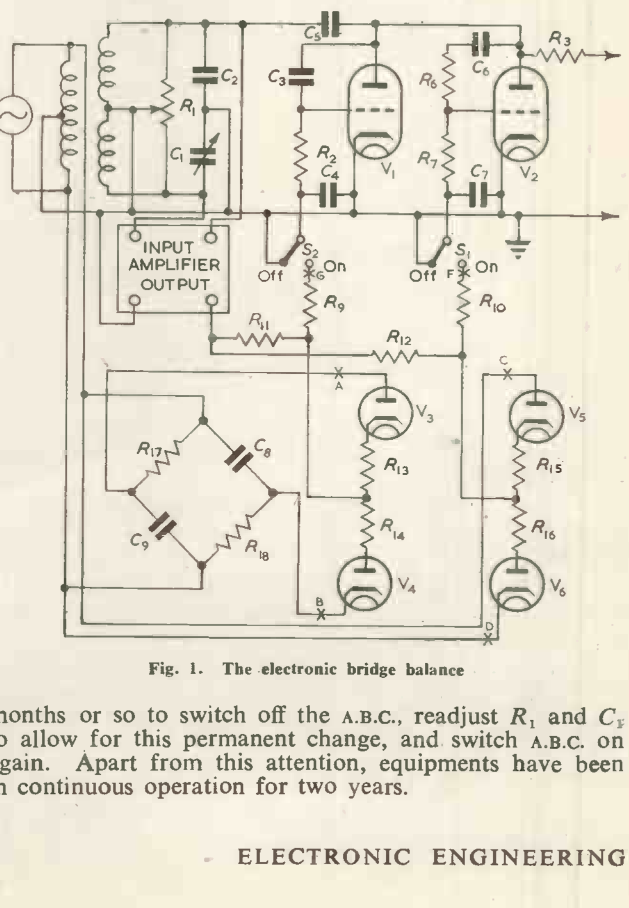


Fig. 1. The electronic bridge balance

* *Industrial Electronic Instruments, Australia.*

A Combined Timer and Cycle Counter

By P. Huggins*, A.M.Brit.I.R.E.

THE unit described combines the duty of timer and cycle counter, within the range 0·02sec to 2·64sec, the change-over from timer to counter being effected by a two-position toggle switch. It has some interesting features, including an extension of the dekatron technique[1].

The equipment is easy to operate and is similar in mechanical design and general appearance to the author's previous cycle counter[2]. The main feature is the two 12-position dekatron valves projecting through the facia panel. These "clock up" the time visually as the apparatus operates.

Advantages

The main advantages of the circuit are as follows:—

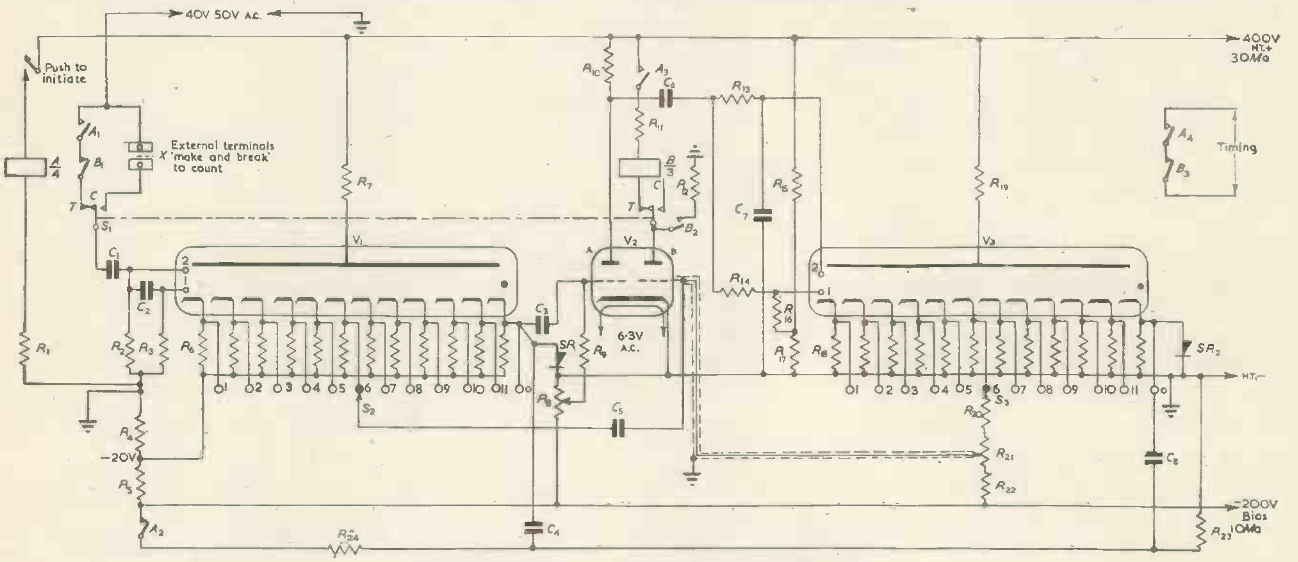


Fig. 1. Circuit of the combined timer and cycle counter

R_1, R_{11} = 33 kΩ ± 20%. R_2, R_3 = 100 kΩ ± 20%. R_4 = 22 kΩ ± 5%. R_5 = 220 kΩ ± 5%. R_6 = 220 kΩ ± 1%. R_7, R_{19} = 1·2 kΩ ± 5%. R_8, R_{21} = 100 kΩ. Wire Wound Potentiometer. R_9 = 1 MΩ ± 5%. R_{10}, R_{13}, R_{14}, R_{16}, R_{24} = 68 kΩ ± 10%. R_{12} = 15 kΩ ± 5%. R_{15} = 200 kΩ ± 5%. R_{17} = 10 kΩ ± 5%. R_{18} = 180 kΩ ± 5%. R_{20}, R_{23} = 470 kΩ ± 20%. R_{22} = 5 MΩ ± 5%. C_1, C_2 = 0·05 μF ± 20%. C_4 = 0·005 μF ± 20%. C_3, C_5 = 0·03 μF ± 20%. C_6 = 0·5 μF ± 20%. C_7 = 0·015 μF ± 20%. C_8 = 0·025 μF ± 20%. S_1 = Double Pole Double Throw Toggle Switch. S_2, S_3 = 12-Position Rotary Selector Switch. V_1, V_3 = Dekatron Valve, Type GS.12.A. V_2 = Double Triode Valve, 6.N.7. A/4 = P.O. Relay Type 3000, 3 "make" 1 "break" contact 5000 ohms coil. B/3 = P.O. Relay Type 3000, 1 "make" 2 "break" contact 5000 ohms coil. SR_1, SR_2 = Selenium Rectifiers, Type MR.4. S.T.C.

1. The function of the apparatus is two-fold. It can *control* the time of an operation, or *measure* the time taken by an operation.

2. Two 12-position rotary selector switches permit a preselection of any time from 20msec (1 cycle at mains frequency) to 2·64sec (132 cycles) in 1 cycle steps.

3. The timer is inherently independent of normal mains voltage fluctuations, even though no special stabilization circuits are employed.

4. The apparatus actually counts mains frequency cycles. Thus if the timer is set to 1 second, it counts 50 cycles of sinusoidal E.M.F. irrespective of mains frequency drift. This can be an advantage when the timed device has some connexion with the mains periodicity (e.g. resistance welding equipment), but has an adverse effect upon the timing accuracy in terms of finite time. However, since the electricity supply frequency is considerably more consistent than its potential, there is an improvement over usual unstabilized timers.

Circuit

The circuit uses three valves, but can be considered as consisting of five sections: "units" counting dekatron stage (V_1), amplifier and clipper stage (V_{2a}), "dozens" counting dekatron stage (V_3), coincidence valve (V_{2b}), timing, and reset circuit.

Power Supply

The power supply system is not shown in the diagram, but is of conventional design. Metal rectifiers are used (bridge for H.T., half-wave for bias), and no stringent smoothing requirements arise. The 40V A.C. supply may be obtained by a tap on the H.T. winding, or by a resistance chain across the complete winding.

Static Conditions

In the state of rest, relays *A* and *B* are both de-energized, their respective contacts being in the positions illustrated. The dekatrons are quiescent, since the anode-cathode P.D. is steady, and there is no input to V_1 or V_3 guides. (A_1 is open, V_{1a} is biased back beyond cut-off.)

There is an independent standing guide-cathode voltage of about +20V for each dekatron: this assists in concentrating the glow in the cathode regions of the valve in preference to the guides.

The initial surge of current charging C_4 and C_8 will cause the anode-cathode P.D. of the "O" cathodes to be higher than for any of the other cathodes. Thus the glow strikes on the "O" cathodes, and (since the minimum maintaining voltage is lower than the striking voltage), once struck, the glow remains invested there.

Action as a Timer

Switch S_1 is in position T. Timing is initiated by holding the push-button depressed throughout the sequence. Relay

* *Sciaky Electric Welding Machines Ltd.*

A closes and contact A_4 commences the timing period. Contact A_1 closing allows a continuous feed of 40V R.M.S. 50c/s sinewaves into the guides of V_1 via A_1, B_1, S_1 and C_1. Now the negative half-cycles of this sinusoidal E.M.F. will more than overcome the standing positive bias (+20V) on the guides, and consequently the guide-anode P.D. exceeds that of the anode-cathode. Hence the glow will move from the "O" cathode to the adjacent guide. By means of the phase advancing network C_2 and R_3 it is ensured that the instantaneous value of the E.M.F. on guide 2 is always slightly lagging on guide 1. Thus the glow sequentially follows the E.M.F. from guide 1 to guide 2, and thence to cathode 1, when the P.D. from guide-cathode returns below that of the anode-cathode. The explanation of the advancement of the glow from guide 2 to cathode 1, rather than back to the zero cathode, lies in the physical geometry of the tube[1].

Every succeeding negative half-cycle drives the glow around from cathode to cathode, the tube current raising the cathode upon which the glow is invested to a potential some 25V above the other cathodes.

When the glow has completed one revolution of the tube, and re-strikes on the "O" cathode, a positive pulse is fed through the coupling capacitor C_3. This in turn drives V_{2a} conductive for the duration of the pulse, causing an amplified anti-phase version of this pulse (plus some clipping) to appear across R_{10}. The amplifier pulse is resistance-capacitance coupled to guide 1, and fed via a quasi-integrating network ($R_{13}C_7$) to guide 2. This ensures sequential firing of V_3 every time the glow in V_1 passes through the "O" cathode position. Consequently the glow in V_3 dekatron moves around one position for every complete revolution of the glow in V_1 (i.e. every 12 completed input sinewaves = 0·24sec).

So far the position of the time selector switches S_2 and S_3 has been ignored. Suppose, for instance, that S_2 and S_3 are both on positions 6 (i.e. timing 78 cycles). It will be apparent that every time the glow current of tube 1 passes through the cathode resistor associated with S_2 position 6, a positive pulse will occur on V_{2b} grid. Now V_{2b} bias (as determined by $R_{18}R_{20}R_{21}R_{22}$) is some 30V below cut-off. The arrival of a 25V pulse from V_1 will not, therefore, make V_{2b} conduct.

However, when the glow discharge in V_3 rests on the cathode associated with S_3 position 6, then the level of R_{21} is raised by about 25V because of the IR drop across R_{18}. This brings V_{2b} grid voltage within a few volts of conduction, and occurs as V_1 glow goes through "O". Thus, six cycles later (when the glow alights on cathode 6 of V_1) the positive output pulse fed through C_5 does cause conduction because of the coincidence of the two positive levels: the D.C. from V_3 and the pulse from V_1. Relay B then energizes and is maintained by its holding contact B_2.

The opening of B_3 terminates the timing period. The opening of B_1 discontinues the A.C. drive into V_1, and so the dekatrons "stick" on figure 6 for each tube.

On re-opening the push-button switch relay A de-energizes. A_3 opens, de-energizing coil B. A_2 closes and sends a negative pulse through C_4 and C_8 as they charge to the −200V level. This negative pulse on V_1 and V_3 "O" cathodes brings the dekatron glow back to zero, and the apparatus is then ready for re-initiation.

Operation as a Counter

This is identical to that as a timer except that S_1 is in position C. Consequently, relay coil B is disconnected and so B cannot energize.

Instead of the 40V A.C. drive for V_1 being obtained via relay A, it is obtained when two external terminals (marked X) are shorted. Hence the connecting together of the X terminals initiates a counting period; the mains cycles being clocked upon the dekatrons in the usual manner.

When X is open-circuited, the drive is removed and the glow remains stationary, thus permitting the number of cycles counted to be read off on the facia panel engraving.

Acknowledgments

The author would like to thank Mr. T. Morrisey for his assistance during development of the prototype model and his suggestions for improving the original idea: he would also like to thank Messrs. Sciaky Electric Welding Machines Limited for permission to publish the paper.

REFERENCES

1. BACON, R. C. and POLLARD, J. R. A New Cold Cathode Counting Tube. *Electronic Engg. 22*, 173 (1950).
2. HUGGINS, P. A Mains Frequency Cycle Counter. *Electronic Engg. 24*, 276 (1952).

"Electrodynamic" Ammeter for V.H.F.

By M. Lorant

IN establishing standards for electrical circuits in the V.H.F. region it is important to extend the direct measurement techniques used at lower frequencies as far as possible. Up to 300Mc/s the current flowing in a circuit whose physical dimensions are small with respect to wavelength is essentially a uniform quantity, and the electrical characteristics of small circuit elements may be determined directly in terms of voltage and current. This fact makes possible the establishment of a standard electrodynamic ammeter for the V.H.F. ranges.

Such an electrodynamic ammeter design, employing a short-circuited ring coupled to a coaxial transmission line, has recently been the subject of a theoretical and experimental study by the U.S. National Bureau of Standards. Basically the method depends on a torque measurement on a conducting ring immersed in a field which does not change with frequency. This technique provides an absolute, broad-band measurement of high-frequency current, but several factors are critical in any actual design.

For minimum distributed capacitance and uniform current the short-circuited ring must be only a single turn, and the ring diameter must be small with respect to wavelength. For accurate inductance calculations the ring conductor should have a small cross section, but resistance then limits the current. A ring 1cm in diameter of No. 20 copper wire is a practical size. When the ring current is small, the torque is also small, and the ring must be suspended on a delicate quartz fibre for accurate torque measurements. The coaxial line, acting as the primary current-carrying element for the electrodynamic ammeter, has several advantages over other forms of conductor. Its electromagnet field can be calculated in a straightforward manner, and the line may be readily modified for calibration work with different types of radio-frequency ammeters.

Calibration of the electrodynamic ammeter may be accomplished directly and absolutely. A section of the coaxial transmission line, one wavelength long at 300Mc/s, is arranged with short-circuited ends to form a resonant cavity, and the torque ring is placed mid-way along the section. A known value of 300Mc/s power is fed into an input loop at one end of the cavity. Under these conditions the torque ring will be at a current maximum and a voltage minimum, and the measured torque on the ring will be due almost entirely to the magnetic component of the cavity field. The measurement is then repeated at 150Mc/s where the current and voltage relations are reversed and the torque is due only to the electric component. One further measurement is needed for absolute calibration of the ammeter. The cavity resonance frequency is measured at both 300 and 150Mc/s with and without the torque ring in place. The resulting changes in frequency are then a measure of the field discontinuity introduced by the presence of the ring. After the torque and discontinuity measurements are completed the instrument will be ready for use as a standard to calibrate other ammeters at very high frequencies.

LETTERS TO THE EDITOR

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

Volume Compression and Expansion

DEAR SIR,—The design for a constant volume amplifier, described by Mr. G. J. Pope in your issue of October, 1952, utilizes a principle which occurred to me in connexion with the design of a volume expander for high fidelity reproduction. This principle consists in the use of a variable-μ cathode-follower (or equivalent feedback amplifier with gain less than unity) in order to provide a varying amount of negative feedback at some point in the amplifier circuit. The controlling grid bias is a positive or negative function of the volume level according to whether compression or expansion is desired. For expansion purposes the control bias is, of course, developed from the unexpanded signal, and the principle becomes particularly advantageous because with maximum signals the variable-gain valve can be completely cut off, thus minimizing distortion.

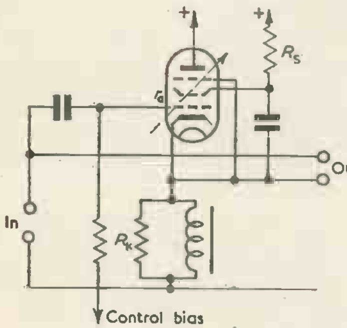


Fig. A

$R = R_K, R_S, r_a$ *in parallel*

One elementary form of this idea is shown in Fig. A. The gain of the cathode-follower is $g_mR\ (g_mR+1)$, where R is as defined in the diagram; the gain of the stage is the difference between this and unity, namely $1/(g_mR + 1)$. If the minimum value of g_m is zero then the extreme expansion or compression voltage ratio is $(G_mR + 1)$, where G_m is the maximum value of the mutual conductance.

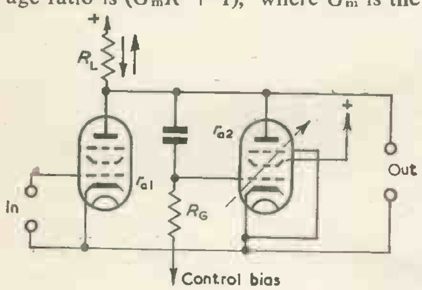


Fig. B

$R = R_L, R_G, r_{a1}, r_{a2}$ *in parallel*

Fig. A operates on the basis of two opposing signal voltages arranged in series in the output circuit. The same effect can be achieved by feeding two opposing currents through a common load, as is done in Figs. B and C. Fig. B has the advantage of dispensing with the cathode choke; its disadvantage is the very low impedance (due to Miller effect) presented to the control bias line, unless a separate grid is available for gain control as in a mixer. Fig. C is essentially Mr. Pope's circuit. All three circuits obey the same law of gain variation provided the values of R are appropriately calculated.

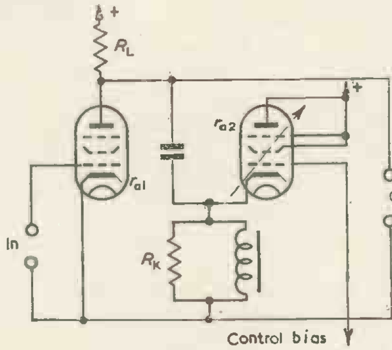


Fig. C

$R = R_L, R_G, r_{a1}, r_{a2}$ *in parallel*

The chief advantage of this principle is the linear relationship, obtainable over a wide range of volume levels, between decibel gain and control bias voltage. When used for expansion purposes there is also the possibility of modifying the response characteristic at low volume levels by means of suitable filtration in the feedback path.

Yours faithfully,

BRIAN D. CORBETT,
Department of Clinical Research,
University College Hospital,
London, W.C.1.

The author replies:

DEAR SIR,—I was very interested to read of Mr. Corbett's experiments with variable negative feedback devices. The circuits of Figs. A and B are very ingenious and combine the variable μ principle with the variable negative feedback element, both often used separately for compressors and expanders.

Since negative feedback occurs in varying amounts throughout the working range of circuits A and B, a beneficial effect is no doubt exercised on distortion production. Presumably, however, the input signal must be limited to avoid exploring too large a part of the essentially curved grid characteristic of the variable μ valve.

The advantage is mentioned for circuit B that the necessity for a choke load is eliminated, and it may not be out of place here to mention some work done in this connexion. Using a similar circuit to that shown in my article, a resistive cathode load was employed, the control bias being fed, suitably decoupled, between cathode and grid of the cathode follower. The value of this resistor must be such as to provide sufficient potential difference between anode and cathode of the valve throughout the range of anode current swing. In practice a 10k resistor is about the upper limit. The disadvantage of this arrangement, however, lies in the fact that with large input signals, the control bias drives the grid-cathode potential towards zero, when the anode current may become so great that the potential difference between anode and cathode becomes insufficient, and the g_m will fall. This fall in g_m will cause an increase in anode load impedance of the controlled amplifier which will further increase the control bias. Thus the grid will lose control of the space current which now depends on the anode to cathode potential, and expansion of the signal will commence. If the cathode resistor is reduced to avoid this trouble, less compression results.

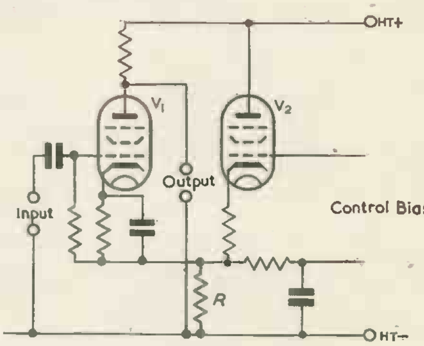


Fig. 1

A circuit overcoming this disadvantage has been produced. Here the variable element is included in the cathode circuit of an amplifier to provide negative feedback under the control of a bias voltage, a common resistor being used as cathode follower load and feedback element for the amplifier (see Fig 1).

Maximum slope and current conditions for V_2 occur when no bias is applied and the value of R may be set to preserve its g_m under these conditions. The control bias, fed as before from the compressed output must now be negative going.

The H.T. line potential should be increased above a normal 250 volt or so value by the amount of D.C. drop across R due to the space current of V_1 alone, so that both valves receive sufficient operating potential. A small bias resistor is included in the cathode-grid circuit of V_2 to limit anode current under "no signal" conditions.

Yours faithfully,

G. J. POPE,
Hillingdon, Middx.

Principles of Radio

DEAR SIR,—The purpose of my letter published in the November issue of ELECTRONIC ENGINEERING was not so much to question Mr.

Lockyer's approach to the exposition of Q (for, as he points out, mine is the same) but to question the propriety of a reviewer contradicting an author on a matter of opinion rather than of fact. It seemed only fair to point out that no less an authority than the relevant British Standard definition could be quoted against him when he states that "in fact this property (Q) can only be assigned to a resonant circuit", so the authors of the book in question ought not to be blamed too categorically for following such authority and others that could be mentioned.

But, as Mr. Lockyer points out, the difference of opinion itself is worth bringing to notice, for it affects the teaching of students at an important point. His view can be upheld by a more positive and authoritative statement[1] than mine: "Strictly speaking, Q-value should be applied only as a characteristic of a resonant system. However, the term Q is often applied to a single non-resonant element and in this case the Q-value and/or the magnification factor, given, should be understood to apply to the resonant circuit consisting of the element in question and a suitable loss-free reactance." On the other hand, these same authors base their fundamental definition of Q on the ratio of energy stored to energy dissipated, which, it seems to me, can be applied to any circuit component; and it appears to be a matter of theoretical and practical convenience that this is derived or measured in the resonant condition. It makes no difference to the steady-state working of an inductor, and to what is commonly called its Q, whether the voltage across its terminals is shared by a capacitor tuning it to resonance or is obtained straight from a generator.

One advantage of confining Q to whole resonant systems is that there is then no ambiguity about frequency, whereas the Q of a coil or other component is not definite until the frequency is specified. But it is surely no more difficult to state the frequency outright than to leave it to be inferred from the parameters of a resonant circuit? In any case, Q varies comparatively little over the working range of a coil, this fact being one of its attractions as a figure of merit. This being so, is it really important to cling firmly to the resonance basis of Q? Mr. Lockyer says that not to do so is unsound and liable to cause difficulties in subsequent work. It would be interesting if he would enlarge a little on this point.

One condition I would make to his insistence that Q should be based on circuit magnification—that the ambiguity of circuit magnification be made clear. It depends on how and where the input voltage is introduced. If in series with the coil as in the conventional Q meter it gives one figure, which disagrees with the usual expression for Q as $\omega L/R$; if induced into the coil magnetically as in many actual tuning circuits it gives another, which is the same as Q.

I am sorry that just at the moment when Mr. Lockyer appeals to me as a pillar of orthodoxy (apparently carrying more weight with him than even the British Standard!) I should show signs of wavering, but as the opener of the discussion on the paper just mentioned said: "Q is now used as a measure of the phase-angle of a two-terminal impedance, and this raises the question whether it will be possible to go back to the formal definition, or whether we should acknowledge common practice." If it can be shown that "common practice' does cause confusion disproportionate to its convenience I will stand by Mr. Lockyer on the side of orthodoxy, but if not I am prepared to move with the times.

Yours faithfully,

M. G. SCROGGIE,

Bromley, Kent.

REFERENCE

[1] A. J. BIGGS and J. E. HOULDIN, "The Development of Q-Meter Methods of Impedance Measurement," *Proc.I.E.E.*, Part III, July 1949.

Audio Frequency Spectrum Analysis

DEAR SIR,—In the first part of his article on problems in audio frequency analysis in the June issue Dr. Soanes mentions the experimentally observed "oscillations" near the beginning of the sweep.

These oscillations may be confirmed by theory when one attempts to obtain the transient response of a passive network excited by a driving force with a linear frequency change.

If the passive network with a system function

$$G(s).s = \sigma + j\omega$$

is driven by an E.M.F.

$$e(t) = k \sin (ht - at^2)$$

where k, h, a are constants, the integral expression of the corresponding transient response is

$$f(t) = \frac{1}{2\omega j} \int_{\gamma} E(s)G(s)e^{st}ds \quad \ldots\ldots (1)$$

where $E(s) = f_t[e(t)]$ — the Laplace transform.

Expressing the sine function in its exponential form and finding its Laplace transforms the following expression is obtained

$$E(s) = E_1(s) + E_2(s)$$

where

$$E_1(s) = -\frac{\sqrt{jk}}{4}\sqrt{\pi/a}\exp(-h^2/4ja$$

$$-h/2a\,s + 1/4ja\,s^2)\,.\,\text{erfc}\left[\frac{h}{2\sqrt{ja}}(s/h - j)\right]$$

and

$$E_2(s) = \frac{k}{4\sqrt{j}}\sqrt{\pi/a}\exp(h^2/4ja - h/2a\,s$$

$$+ js^2/4a)\,.\,\text{erfc}\left[\frac{-\sqrt{jh}}{2a}(s/h - j)\right]$$

Since the integration of the expression (1) is rather difficult—the approximate resolution is used.

Assuming $f^*_{(n)}(t)$ be such that

(a) $f^*_{(n)}(t)\, f(t)$ uniformly,

(b) the convergence of $f^*_{(n)}(t)$ towards f(t) is very rapid within the given interval of time—

a sequence of $f^*_{(n)}(t)$ is produced by the "pocket method", and thus an approximate solution with a wanted degree of accuracy is obtained.

Having a resonant LC circuit with $\omega_o^2 = 1/LC$ and if the current entering it is of the form

$$i(t) = k \sin (ht - at^2)$$

the solution for the output voltage in the first approximation results in the following expression:

$$V^*_1(t) = \frac{-kh}{C(\omega_o^2 - h^2)} 2 \sin [\tfrac{1}{2}(\omega_o + h)t]$$

$$\sin [\tfrac{1}{2}(h - \omega_o)t]$$

valid for $0 \leqslant t < \dfrac{\pi}{\omega_o + h}$

This expression, independent of a, shows that the transient wave increases sinusoidally with the angular velocity $\frac{1}{2}(\omega_o + h)$.

For $\omega_o \approx h$ it becomes

$$V^*_1(t) = -\frac{kh}{C(\omega_o + h)}t\,.\,\sin [\tfrac{1}{2}(\omega_o + h)t]$$

There is therefore a ratio of the pass-band to the sweep frequency at which oscillations can be observed on the scope.

Yours faithfully,

W. SAGAILLO,

Erith, Kent.

REFERENCE

[1] Guillemin, E.A., Cerrillo, M.V., Kautz. W. H.' and Lucal, H. M. Research Reports of the Electronics Laboratory, M.I.T. January, 1951.

The author replies:

DEAR SIR,—I am indeed pleased to hear of the additional reference which Mr. Sagajllo has mentioned. Although not directly stated, Fig. 1 of my paper is a purely theoretical graph and follows quite directly from the work of Barber and Ursell.

With reference to the M.I.T. solution quoted by Mr. Sagajllo, the initial applied frequency is $f_o = h/2\pi$ and the first approximation $V_1(t)$ is valid for $t < \pi/\omega_o + h$, as he says.

(a) For $h \gg \omega_o$, the solution is valid for $t \mid\gg \dfrac{\pi}{\omega_o + h}$ which is less than $\frac{1}{2}$ cycle of the applied frequency.

(b) For $h \approx \omega_o$, the solution is valid for $t \mid> \pi/h$ which is less than $\frac{1}{4}$ cycle of the applied (or resonant) frequency.

Neither (a) nor (b) can prove the existence of a noticeably oscillatory response envelope which would necessarily extend over many cycles of the applied frequency.

(c) For $h \ll \omega_o$, there is no significant response at all.

The first approximation is independent of a, but the 2nd and others are not.

The circuit selected for the example is an LC circuit with no resistance. The pass band is therefore zero, and there can be no significance to the "ratio of the pass band to the sweep frequency," which anyway does not seem to appear in the last equation.

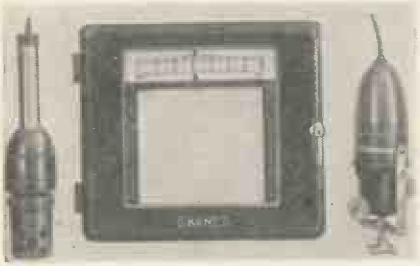
Yours faithfully,

SIDNEY V. SOANES,

Toronto, Ontario.

ELECTRONIC EQUIPMENT

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



Continuous pH Recorder/Controller

(Illustrated above)

A GLASS-ELECTRODE pH recorder for automatic control of industrial processes has recently been introduced by George Kent Ltd. It is claimed that the instrument will provide accurate and continuous records of pH over periods of several months without any maintenance attention, the addition of potassium chloride solution to the salt bridge also being unnecessary. The accuracy of the instrument is not affected by sampling under pressures up to 75 lb. per square inch. The recorder also features plug-in electrodes with quick-acting glands.

The primary element, the body of which is made of high-temperature grade material, can be used in two ways. The tank-type can be arranged so that the electrodes just dip into the liquid to be measured, or the electrode system can be completely submerged. This type has a built-in support complete with a quick-acting clamp. By simple replacement of accessories, the tank-type is converted into the flow-type, in which the liquid to be measured flows through a sealed hopper fitted on to the electrode holder. This type may be used for measurement under pressure.

The combined reference/salt bridge electrode is unbreakable and pressure compensated. The liquid junction is resistant to fouling and is a separate component which can be replaced quickly at negligible cost.

The three types of glass electrode (see Table given below) cover virtually the full theoretical range of the pH scale.

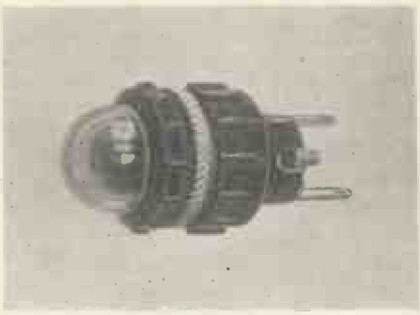
A nickel resistance thermometer, protected by a moulded rubber sheath, provides automatic temperature compensation for the electrodes.

The measuring system is in two parts —an electrometer and a recorder. The impulse-type electrometer, operated from the A.C. mains, measures the potential of the electrode system without imposing any significant load on the system. The electrometer circuit has a high input impedance which is symmetrical with respect to earth, making possible the use of robust and service-free electrodes. Only one valve is used in the electrometer. Variations in the valve characteristics do not cause zero drift, and changes in mains supply voltage do not affect the accuracy of the instrument. The electrometer can be mounted up to 100 feet away from the primary element, and the self-balancing potentiometer/recorder may be located up to 50 feet from the electrometer.

Processes involving the circulation of hot process liquor, for example, can be automatically controlled pneumatically with the aid of this universal recorder.

George Kent Ltd., Luton, Beds.

ELECTRODE	pH RANGE	TEMPERATURE RANGE	ESTIMATED CONTINUOUS-OPERATION LIFE BEFORE SERVICING
General-purpose	0–10	<50° C.	Over 6 months
High-temperature	0–14	50°–100° C.	About 2,000 hours at 100° C.
High *p*H	10–14	<75° C.	More than 2,000 hours



Arcolectric Signal Lamp Holder

(Illustrated above)

THE new Arcolectric signal lamp holder, type S.L.88, is totally enclosed and insulated so that it will withstand 1500 volts A.C. flash tests. It will accept most popular M.E.S. low voltage bulgs, and can be fixed in a single ½in. hole. The domed plastic lens allows 180° visibility, and is available in a wide range of transparent and translucent colour. The lamp is accessible from both the rear and the front.

Another version of the holder is available for use on mains voltages. A feature of this design is that the resistance is built-in, and the holder can be connected direct to the mains. No external resistor is needed for voltages up to 250 volts. The holder is designed for use with the Arcolectric neon tube type T.47, which facilitates replacement. The standard holder is for use on 200/250 volts A.C./D.C., but other models can be supplied for 80/500 volts.

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

Leland-Maury Electronic Auscultoscope

(Illustrated below)

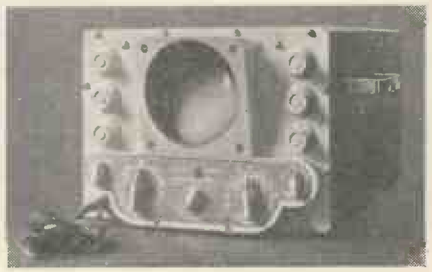
THE Leland-Maury Electronic Auscultoscope is a new instrument for the visual and aural observation of the sounds and functioning of the heart and lungs, and makes available to the medical profession generally the technical facilities of the Maury Laboratory Auscultoscope apparatus.

The auscultoscope is an electronic instrument which provides a means of listening to the sounds of the heart and lungs, and also of viewing, and if necessary recording photographically, a graphic representation of the heart and lung sounds on the screen of a cathode-ray tube. The sounds are picked up from the body by means of a small electro-dynamic transducer which, held against the patient's chest, conveys a signal through a control and amplifier unit to a loudspeaker and a visual display unit, which may be used separately or simultaneously.

The transducer, which is analogous to the microphone and which to some extent functions as such, has a response characteristic which enables it, on the one hand to avoid as far as possible such unwanted noises as those produced by surface contact between the transducer and the patient's skin, but on the other hand to accept the very low frequencies of the fundamental heart and lung movements which are verging on sub-audible frequencies, and also the sub-audible vibrations for observation on the visual display unit.

The control and amplifier unit has an over-all response characteristics which is complementary to that of the transducer, and provides a high degree of amplification of the desired body sounds, without the introduction of spurious noise, thus making audible and visible the minor and secondary sounds of heart and lung functioning. The lower fundamental pulse frequencies are also rendered audible as far as their very low frequency permits, and are clearly visible by graphic representation on the screen. The controls allow the degree of over-all amplification to be varied and permit certain parts of the frequency response band to be selected and emphasised to observe either the fundamental or the harmonic movements and sound of the organ under consideration.

The loudspeaker unit provides maximum output of very low frequencies.



While suppressing the very high frequencies associated with background noise, it gives clear and well defined reproduction of the higher frequencies associated with the heart valve movements and of the higher harmonic frequencies which give to the heart sound the "character" by which the trained ear and eye are enabled to provide a clinical interpretation of the sounds and their graphic representation on the screen.

The visual display unit is fundamentally a cathode-ray oscilloscope, having a 6in. diameter screen with low persistence coating giving a blue-green trace of high actinic value, thus permitting observation of low frequency traces, and also good strength images for photographic recording. The display unit incorporates its own time-base, which is designed to give low horizontal sweep frequencies which can be adjusted to the patient's heart beat rate, and directly calibrated to indicate the number of beats per minute. A synchronizing control enables the time-base to be locked, as far as possible, with the low and varying frequencies of the heart to give a steady graph of the organ's sounds.

The camera is an orthodox oscilloscope camera which permits records to be made on paper or film-based photographic material, the spool accepting standard 25ft rolls of 35mm film so that sufficient film for thirty individual records is available from one loading. There is also facilities for removing each record individually for immediate development if required. Single frame still records can also be made with the camera.

Leland Instruments, Ltd.,
22 Millbank,
London, S.W.1.

Truvox Tape Deck

(Illustrated below)

THE new Truvox tape deck has been designed for building into complete recorders so that they can be handled by inexperienced personnel. This is achieved by a method of tape control and tape loading which is extremely simple to operate. Push-button control enables the user to either work the machine on the record-play-back, fast-forward or fast-reverse positions. In the latter two positions of the switches a complete reel can be re-spooled in under a minute.

The braking system has a separate button which will stop the mechanism almost instantaneously when depressed. The drive is a type of silent friction mechanism to eliminate "wow" and "flutter."

The motors are normally wound for 230 volts, but other voltages are available to order. The heads are so designed that the erase head works from a low impedance circuit to ensure satisfactory erasure and minimum loss of energy in screened cables, while the record/replay head is of the high impedance type, rendering the use of transformers unnecessary. This head is, however, tapped down so that it also receives a bias voltage from a low impedance circuit.

Truvox, Ltd.,
Truvox House, Exhibition Grounds,
Wembley, Middx.

Switchboard Termination Unit

A SWITCHBOARD termination unit for use with the Pye PTC.703/4 V.H.F. fixed station has been designed and manufactured on behalf of Pye Telecommunications Ltd., by Ericsson Telephones Ltd., which permits the connexion of a duplex radio link circuit to a manual switchboard in a manner exactly similar to a line circuit.

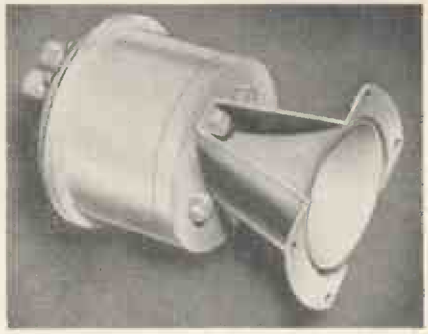
There are two types of application for this unit, either for "subscribers" circuits, or for junction circuits between two exchanges. For the former, a termination unit is used at the exchange end of the link while the remote end is connected "four wire" as a standard duplex station. For a junction link, termination units are used at each end.

The equipment comprises a panel 3½in. by 19in. which mounts in the standard fixed station cabinet and carries hybrid transformers, relays, power unit, etc. For the switchboard a standard jack and indicator are supplied loose, together with two relays and two isolating capacitors, which may be mounted in any convenient position at the back of the board. Only two wires and earth return are needed between the exchange and the V.H.F. fixed station. The length of line between the two is limited to a maximum loop resistance of 500 ohms. The unit contains its own mains power pack (100-150, 200-250 volts, 40-60 cycles), and is completely independent of, and isolated from the exchange battery. This means that the standard unit may be used with any type of switchboard.

The receiver and transmitter heaters are continuously switched on. When the operator plugs into the line, the local transmitter is switched on and the indicator operates. The arrival of the signal at the remote receiver operates its mute relay, and this in turn operates the indicator on the remote switchboard. When the operator plugs in to answer, the remote transmitter is switched on, communication is established and both indicators return to normal. A reverse sequence is followed when the line is cleared at the end of the call.

The unit can equally well be wired to a cordless type of switchboard. As will be seen from the sequence of operations above, the V.H.F. radio "lines" are presented to the operator in exactly the same form as normal wire lines.

Pye Telecommunications Ltd.,
Radio Works, Cambridge.



Stentorian Pressure Type Tweeter Unit

(Illustrated above)

The Stentorian Pressure Type Tweeter Unit is of the moving coil pressure type and is similar to that used in the 10in. and 12in. concentric Duplex loudspeakers. The speech coil is of aluminium wire, wound on an aluminium former which is rigidly fixed to an aluminium diaphragm, and these are situated at the rear of the magnet. The centre pole is hollowed out to form the commencement of the horn, in the centre of which is the phase equalizer.

The speech coil impedance is 15 or 30 ohms, and the flux density is 14 000 gauss. The unit has a response of 2 000-14 000c/s and a power handling capacity of 3 watts.

Whiteley Electrical Radio Co. Ltd.,
Radio Works, Mansfield.

Electrothermal Rubber Sheeting

A NEW method of applying heat in defined areas of any size has now been developed and produced by Electrothermal Engineering Ltd. under the name E.R.S.

E.R.S. appears to be a flat sheet of black rubber, but it is electrically wired at 150, 250 or 400 watts per square foot for mains operation to radiate heat at controlled temperatures up to 200°C sufficient for applications of heat to flat, shaped or curved surfaces, where the required heat input does not exceed 400 watts per square foot. E.R.S. ratings provide for a temperature of 100°C (Type 1), 150°C (Type 2) and 200°C (Type 3) respectively. For general applications Types 1 and 2 have become the standard choice, but Type 3 should be selected if rapid temperature rise is necessary. Thermostats and pyrometers are available for use with the sheeting.

Standard sizes of E.R.S. range from 12in. by 6in. to 96in. by 48in. heated area, but the product can be supplied to individual temperature and production requirements. The sheeting operates on 100/200 or 200/250 volt mains A.C. or D.C., but can be supplied for all other voltages. It has been tested satisfactorily for resistance, completely immersed in water at 2 500 volts A.C. and 1 000 D.C. The standard E.R.S. will take approximately 20/30 minutes to reach maximum temperature, and will thereafter maintain that temperature, but it can be designed to reach a specified temperature in a shorter period.

Electrothermal Engineering Ltd.,
270 Neville Road,
London, E.7.

BOOK REVIEWS

Automatic Feedback Control

By W. R. Ahrendt and J. F. Taplin. 412 pp. McGraw Hill Publishing Co. Ltd. 1951. Price 64s.

THIS book presents its subject in a very readable and in general a logical manner. The mathematician will learn nothing more on his own subject, but on the other hand, he may benefit from the chapters dealing with the more practical or engineering aspects of control.

A balance has been achieved such as might be expected from the authors representing as they do, the academic and the instrument manufacturing and application viewpoints.

The subject is introduced in a gentle manner which should give the beginner a good concept of its scope and purpose, but he will not get far without a fair working knowledge of differential equations and transforms.

The outlook of the practical experimentalist is apparent and reveals itself particularly in such an expression as (p. 263) "In order to determine whether or not a servo will meet its performance requirements, there is no better test than the actual application for which it is intended"; and also (p. 106) the practical advice is given, not to design for the apparent ideal, but to limit the frequency of response of the system being designed to that required to deal with the genuine control motion required, thereby rendering the system more safe against resonance, noise and spurious signals which often appear in the frequency spectrum outside the operative range.

On p. 185 it is admitted that it is often necessary to resort to a graphical solution, but that (p. 199) "mathematics act as a guide to design".

Chapters 1 to 6 deal with the fundamentals and theory, 7 and 8 with design problems and Chapters 9 to 13 with practical applications.

After a general introduction in Chapter 1, the second chapter describes the method of forming differential equations for simple systems and introduces the Laplace transform method of solving them. Chapter 3 follows with the classical approach to the analysis of physical systems beginning with the simple servo-mechanism, while Chapter 4 presents the frequency response approach. Applications of this method to simple closed loop systems are covered in Chapter 5 and multiple loop systems in Chapter 6. There follow two very useful chapters on design considerations, and non-linearity and discontinuity. Chapter 9, opening the practical section, deals with servo-mechanisms and is informative on many practical details. Chapters 10 to 13 illustrate the applications of feedback control in pneumatic controllers, temperature regulators, speed governors and fluid control.

It is pleasing to see the Laplace transform used in the early chapters of the book on the simpler differential equations, which enables the new reader to grasp more easily the operational method.

It is a little unfortunate that Chapter 2 tends to be rather jumbled. It opens with formation of differential equations, continues with the Laplace transform, in the middle of which there appears a section on complex numbers and this is followed by paragraphs on transfer functions and frequency spectra, before including a paragraph on the geometrical interpretation of the Laplace transform.

The book appears to be designed for the student in training to be a feedback engineer and for engineers and others who wish to improve their knowledge of this aspect of engineering. If this is the purpose of the book then the reviewer considers that the object has been achieved.

J. BELL.

Electricity Meters and Instrument Transformers

By S. James. xi + 467 pp. 205 figs. Chapman and Hall Ltd. 1952. Price 50s.

AT the outset it should be pointed out that the title of the book is strictly correct: it deals with electricity integrating meters and excludes voltmeters, ammeters and other indicating and recording measuring instruments.

Unlike most books on metering, this book is written by a designer with a lifetime of commercial manufacturing experience; consequently it is both authoritative and of practical use in connexion with meter manufacture, instead of only to installation, etc., from the users' aspect. Except in one respect, and this is forecast in the Appendix, the book is well up to date. The exception is due to the inevitable delay between writing and publication, references to B.S.37: 1937, being not now fully correct owing to the recent revised issue of B.S.37, Parts 1 and 2, 1952 (with other parts to follow). But all interested readers will have the revised British Standard and will refer to it as necessary.

Chapter 1 deals with specifications and legal requirements. This is an admirable start, since meter performance has to comply with various statutory requirements in addition to the British Standard. Most of the chapters deal with a particular class of meter—D.C., single-phase, polyphase, prepayment, etc., each including both early types such as electrolytic meters as well as modern types. After these chapters follow others on more specialized and recent aspects—reactive meters, both internally compensated and for use with special transformers; kVA meters, with a discussion on power factor implications and ambiguities in the measurement of the quantity, which is not always a simple conception; maximum demand indicators, with a note on tariffs and details of thermal bimetallic and integrating types of meter, the latter as used for Grid metering. Summation metering, an important subject due to the Grid network and to comprehensive industrial installations, is dealt with very fully and the various mechanisms and meters described. The last two chapters deal with current and voltage transformers, both in theory and as applied to practical metering.

Meter testing is excluded from this book, but a companion volume dealing with this subject is stated to be in course of preparation.

The book shows signs of careful editing, terminology and symbols being correct throughout. The many line diagrams and other illustrations are extremely clear and the book is well printed and very readable. It contains no higher mathematics.

There is little application of electronics in the metering of industrial and domestic energy, chiefly because the few volt-amperes needed by the meter can be provided from the circuit without affecting it, unlike the case of the valve voltmeter where the instrument energy is specifically not taken from the (low-power) source being measured.

The book can be thoroughly recommended to engineers of all grades who are concerned with integrating meters, both in manufacture and use.

E. H. W. BANNER

Theoretische Elektrotechnik III. Grundzüge der Theorie Elektrischer Maschinen

(Theoretical Electrical Engineering III. Fundamentals of the Theory of Electrical Machines)

By Dr. Ing. Karl Kuhlmann. 547 pp. 328 illustrations. Verlag Birkhäuser, Basel. 1951. Price sf.74.90.

THIS book constitutes the third volume of a comprehensive work on the theory of electrical engineering of which so far only the second volume, dealing with the theory of the A.C. circuit and the single-phase transformer, was published in 1947, while the first and fourth volumes are still in preparation. The present volume deals in great detail with the principles of the theory of electrical machines, their mode of action, the winding techniques, the production of magnetic fields, some induction calculations including stray leakage reactance, the E.M.F.s induced and the magnetic energy and torque of electrical machines. In an appendix additional phenomena and problems are dealt with, viz. the commutation process, particularly that encountered in A.C. machines, special losses in electrical machines, e.g. those due to skin effect, to eddy currents and to magnetic reversal, and the iron losses in machines with circular and elliptical rotating fields. About 90 pages are devoted to the theory of symmetrical components, a special feature being that they deal briefly also with arc-suppression coils.

The treatment is mathematical throughout without neglecting physical considerations and practical aspects, as may be expected from an author who, before being called to the chair of Electrical Engineering at the ETH Zurich, held for many years responsible positions in various departments of one of the largest electrical firms of the Continent.

It is perhaps somewhat surprising that no mention is made of D.C. crossfield generators as the author was, in 1905, the first (together with Hahnemann) to try to give a mathematical theory for this type of machine. There may be two reasons for this omission. The first is that the crossfield generator, although at present still used to some extent as an arc welding generator may not be important enough to justify its treatment in a book already rather voluminous, though it should be remembered that the modern metadynes and amplidynes are based on the crossfield principle. The other reason may be that this type of machine, working as it does with high saturation and strong strayfields, does not lend itself easily to exact mathematical treatment (cf. Rosenberg, Die Gleichstrom-Querfeldmaschine 1928, p. 84).

The old practical system of units is used in this volume, while the more modern M.K.S. system will apparently be dealt with to some extent in the first volume not yet published. The book is a valuable contribution to the literature of electrical theory and is particularly suitable for the advanced student who is able to read a German text. It is well produced, and a brief subject matter index facilitates its use.

R. NEUMANN

Radio Interference Suppression

By G. L. Stephens, A.M.I.E.E. 2nd edition. 132 pp. 65 diagrams and photographs. Iliffe and Sons Ltd. August 1952. Price 10s. 6d.

THIS handbook is a guide to the various methods of suppressing electrical interference with radio and television reception. The author describes the origins of interference and the theory of suppression technique. He then gives practical applications. Typical interfering appliances discussed include: engine ignition systems, switches, thermostats and contactors, electric motors and generators, rotary convertors, lifts, neon signs, fluorescent and other types of discharge lighting, radio-frequency heating apparatus, television receivers, spectrographic equipment and valve rectifiers.

Attention has been paid to the problem of interference at television frequencies and to suppression arrangements on motor vehicles and on board ships. Other chapters deal with the design and choice of suppressor components, methods of locating the source of interference, and suppression at the receiver itself. Reference data is provided in the appendices.

Electric Fuses

By H. Läpple, Dipl.Ing. 173 pp. Butterworths Scientific Publications. August 1952. Price 25s.

THIS book is a critical review of the literature on electric fuses by Herr Läpple, the Chief Engineer of the H.V. Fuse and Overvoltage Protection Department of Siemens-Schuckertwerke, Berlin. In the course of the work he includes his own views and observations as a commercial fusegear engineer, so that the accent is on practical and commercial engineering aspects.

In his introduction the author states that in his opinion "... the fuse not only is a device of former and present importance but will in the future play a notable part in the general field of electrical engineering". Reasons for this opinion are outlined in the brief historical development of fuses given, and the author's resumé of the literature on the subject is influenced by this viewpoint.

Discussions on physical aspects, circuit breaking problems, breaking-capacity, deterioration, time/current characteristics, and an examination of several basic formulae from various sources are covered. Practical aspects outlined include: discrimination, co-ordination, the necessary qualities for satisfactory fuse performance, and their applications to special functions. An interesting feature is the comparison of rules and ratings and the standards used in various countries, while an extensive list of references in which each item includes an abstract of the subject matter covered adds to the value of the book.

The Practical Electrician's Pocket Book

Edited by R. C. Norris. 1953 edition. 551 pp. Odhams Press Ltd. 1952. Price 5s.

THE 1953 edition of this pocket book includes five new chapters and extensive revision of four other sections, while the wiring systems, refrigeration, water heating and reference tables are also brought up-to-date.

The five new chapters cover: a review and explanation of types of switches with diagrams showing their use in lighting, motor and appliance control circuits; time switches and their applications; power factor correction for economy; efficient domestic lighting, and details of the insulation material fibreglass.

Previously dealt with together, the subjects of motors and generators now have separate sections. The latest developments in the radio and television interference suppression field are included in the appropriate chapter. Practical guidance for contractors on installations for farms, dairies, market gardens, poultry houses, etc., is given in "Electricity on the Land", while the section on illumination now includes application data for latest fluorescent tubes and fittings.

Antennas: Theory and Practice

By S. A. Schelkunoff and H. T. Friis. xxii + 639 pp. John Wiley & Sons, Inc., New York, and Chapman & Hall Ltd., London. 1952. Price 80s.

SINCE the beginning of this century considerable attention has been focused on the subject of radio aerials and an extensive theory has been built up. Unfortunately the average radio engineer, who is most concerned with aerial design, has found much of it beyond him because of its highly mathematical nature. The textbook under review makes a welcome attempt to describe the basic principles of aerial behaviour in a way which will be intelligible to such engineers. The authors, the one a well-known mathematical physicist who has made many important contributions to aerial theory and the

BOOK REVIEWS (Continued)

other the director of high frequency research at the Bell Telephone Laboratories, are admirably equipped for writing a book of this kind. They have replaced purely mathematical proofs by a combination of physical pictures and sufficient mathematical work to justify the results presented. The physical background is built up from the concepts of electrostatics and transmission lines both of which will already be familiar to the engineer. Aerial theory is thus developed as an extension of the radio engineer's normal background and not as a totally new subject. What are in so many textbooks mere mathematical equations take on a physical shape which emphasizes their relevance to practical design problems. Much of the apparent simplicity of the treatment has been achieved by dispensing almost entirely with the notation of vector calculus: the operators, curl, div and grad, which so often prove a stumbling block to students, make very rare appearances in this book.

As may be seen from an outline of the contents, almost the whole field of aerial theory is surveyed. An introductory summarizing chapter is followed by three chapters on Maxwell's equations and their solutions for plane and spherical waves. Aerials themselves are then treated from two standpoints: firstly the directive nature of the radiation and secondly the internal working conditions. At this point the current distribution in linear radiators, such as the half-wave dipole, is considered in detail with particular attention to the frequently-used assumption of a sinusoidal distribution. Complete details of the methods used to calculate the current distribution are not given but there are adequate descriptions of the general lines of attack. As elsewhere in the book there are copious references to enable the reader to pursue the subject further if he desires. The remaining chapters are devoted to specific aerials such as self-resonant systems, rhombic arrays, slots, horns, and reflectors.

It is very easy to find fault with a book of this type because pet subjects of the reader have been omitted. The reviewer, for example, feels that the question of bandwidth has not been treated as fully as it deserves and that Yagi and helical aerials merit more than a passing mention. The chapters on the aperture radiators, horns, reflectors and lenses, are not so extensive as those on current radiators and it might have been preferable to expand them into a separate volume. However, these criticisms do not detract from the value of the information which is given.

Since the primary concern of this book is with basic principles, the bulk of it is of a theoretical nature but experimental results are frequently referred to in corroboration and amplification of approximate theories. On the whole it is very readable, but there are a few places where compression would be advantageous. This is partly due to the arrangement of the chapters which is such that there are frequent forward references. The book is well produced and has many admirable diagrams. Few misprints have been detected, the most serious being in Appendix III: the half-power beamwidth of a wire of length l should read $50{\cdot}7\lambda/l$ degrees. Most of the chapters contain examples and the answers are supplied. These examples are not of the all-too-common type which merely involve the substitution of numerical data in formulae from the text, and any reader who works through them conscientiously can claim a good knowledge of aerial theory.

J. BROWN

Electrical Engineering—Theory and Practice

By W. H. Erickson and N. H. Bryant. 523 pp. John Wiley & Son Inc., New York, and Chapman & Hall Ltd., London. August 1952. Price 48s.

THIS book is based on lectures delivered to students of civil, mechanical and chemical engineering at Cornell University.

The subjects covered can be divided into three main parts: circuits, machines and electronics. The first ten chapters deal with the fundamentals of D.C. and A.C. circuits including magnetism, electromagnetic induction, inductance, capacitance, single and polyphase A.C. systems, instruments and measurements. Chapters 11 to 19 are devoted to machines. D.C. generators and motors, together with motor control, are first considered, then transformers, induction motors, synchronous machines, single-phase motors and control equipment are presented. Chapters 20 to 28 are concerned with electronics. In addition to chapters covering electronic emission, valves and their characteristics, considerable space is devoted to rectification, amplification, communications and electronic control.

Since the book is not intended for electrical engineering students it is surprising to find that its scope is so wide. It is debatable whether topics such as the following should have been included: modulation, detection, receivers, generator voltage regulators, resistance-welding control, repulsion motors, synchros, rotating amplifiers, amplidyne, rototrol and regulex. On the other hand, one would have expected a detailed treatment of the cathode-ray oscillograph whereas this important piece of equipment is dealt with in less than $4\frac{1}{2}$ pages in an Appendix.

Physical analysis of the operation of circuits and machines has been stressed, although mathematical analysis is not avoided. Certain methods of treatment, such as the equivalent-circuit concept for A.C. machines, have, therefore, been omitted. Further, A.C. circuit theory is presented without introducing complex quantities, although an Appendix is included explaining complex notation.

At the end of all but four chapters between 5 and 25 problems are found with answers, giving a total of 269 problems. The authors have endeavoured to compile problems which do not require the mere substitution of numbers in formulæ. There are not many illustrative problems in the text.

The book seems to be free from serious errors. The authors use kv, mv, ma, μa and kw instead of the usually accepted kV, mV, mA, μA and kW. The amplitude of an alternating quantity is defined as the maximum value of the quantity which seems unusual and one-half of a cycle is called an alternation which is uncommon. It is stated that voltage-regulator tubes are available for voltages between 75 and 150V and for currents up to 200mA. Tubes are available for voltages between 50 and 160V, however, and for currents up to 200mA. Regulation over the current range of a voltage-regulator tube is often greatly in excess of the 1 or 2V quoted. It is a pity that the authors have found it necessary to use so many words ending in "tron", e.g., phanotron, kenotron and pliotron. They also use both the words anode and plate. No references are given for further reading although some topics are dealt with very briefly. For example, references to some of the literature on strain gauges would have been useful since the subject is treated in three short paragraphs only. It would have been helpful if some suggestions had been given for experimental work. Readers may be surprised to find that the M.K.S. system of units is not used throughout.

The book is easy to read and understand, the diagrams are clear and the explanations have been carefully thought out. The printing and general production of the book are first rate and it can be warmly recommended as a most useful addition to any library.

F. A. BENSON

Scientific Research in British Universities 1951-52

485 pp. H.M. Stationery Office. October 1952. Price 8s. 6d.

"SCIENTIFIC Research in British Universities 1951-2" published by H.M.S.O. for D.S.I.R. gives details of research work in universities and colleges in the United Kingdom. The object of this volume is to provide brief notes outlining the projects being undertaken by British universities and colleges during the 1951-2 session. It covers all fields of science and technology including agriculture and medicine.

The material has been collected by the British Council from the heads of departments of the universities concerned, who are alone responsible for the entries, and it is issued by the Department of Scientific and Industrial Research with the collaboration of the Agricultural and Medical Research Councils.

Details of the research work are arranged according to the universities at which they are being undertaken and include the names of the staff working on the problem, together with a description to show the scope of the research. The names are indexed so that it is possible to find out where a particular person is and the research upon which he is engaged.

Notes from the Industry

City and Guilds of London Institute Insignia Award. The City and Guilds of London considers that in certain branches of industry additional encouragement and recognition could usefully be given at a higher level than that represented by its Full Technological Certificates to those engaged in industry who continue to pursue their studies and to broaden their knowledge. Therefore the Institute proposes to establish under its Royal Charter an Insignia Award in Technology which will lay emphasis upon technical training based primarily upon practical experience, supplemented by theoretical study, as distinct from the more academic approach to training.

The standard of the Insignia Award will be considerably above that of the existing Full Technological Certificate. The Institute intends to introduce the Award gradually as opportunity offers in the main branches of the chemical, constructional, electrical, mechanical and textile industries. The co-operation and assistance of representatives of industry will be sought in judging the eligibility of candidates.

Copies of the general regulations governing the Insignia Award scheme, together with notes for the guidance of candidates, and an application form for the registration of candidates (Form CGIA/1) will be sent to any applicant on receipt of a stamped addressed foolscap envelope. Inquiries should be addressed to the Director, Department of Technology (I.A.), 31 Brechin Place, South Kensington, London, S.W.7.

New Broadcasting Equipment for the Union of South Africa. An order for what is probably the largest number of high-power broadcasting transmitters ever placed in Britain has been awarded to Marconi's Wireless Telegraph Co. Ltd. These transmitters will give complete broadcasting coverage to the Union of South Africa. The network of nine 20kW high-frequency transmitters will be centred on one site.

The new system will be additional to the existing medium frequency transmitters already serving part of the Union, and the new transmitters will be installed at Paradys near Bloemfontein.

Temporary Low-Power Television Stations at Pontop Pike and near Belfast are to be installed by the BBC so that viewers in the neighbourhood of these stations may have television in time for the Coronation.

The BBC's plan for five permanent medium-power stations is still deferred by the Government, but at the two sites mentioned it is proposed to use temporary transmitters having a power of 1kW (vision) and $\frac{1}{4}$kW (sound). These will not give the range and quality that will be ultimately obtained from the permanent medium-power stations and there will be a greater risk of breakdown.

The transmitter at Pontop Pike is expected to serve about one million people within a radius of some 20 miles, which includes Tyneside; the transmitter in Northern Ireland will serve about half a million people in the city of Belfast and its immediate surroundings.

The transmissions from both stations will be horizontally polarized. The frequencies used at Pontop Pike will be vision 66·75Mc/s, sound 63·25Mc/s shared with Wenvoe, and those at Belfast; vision 45·0Mc/s and sound 41.5-Mc/s, shared with Alexandra Palace.

1953 Radio Show. The Radio Industry Council announces that provisional dates for the 20th National Radio Show, 1953, to be held at Earls Court, London, are September 1 to 12.

The Aeronautical Division of Marconi's Wireless Telegraph Co., Ltd. has been expanded and reorganized to meet the increasing demand for specialized wireless equipment for aeronautics. The division was formed as a separate unit for the company after the first World War, and Mr. L. A. Sweeny has been its manager since 1936, with the exception of service with the Fleet Air Arm between 1942 and 1946.

To meet with the reorganization, Dr. B. J. O'Kane, Ph.D., B.Eng., A.M.I.E.E., was appointed Chief Air Radio Engineer as from October 1st.

Erratum. Mr. D. A. Levell, the author of "A Hard Valve Pulse Generator", published in the November issue of ELECTRONIC ENGINEERING, has called our attention to an error in the component values given on p. 508. These should be:—

$C_2 = 300$pF. $C_3 = 0{\cdot}01\mu$F.

BINDING OF VOLUMES

Arrangements for the binding service started last year are being continued this year, and the 1952 volume can be bound at an inclusive charge of £1.

Copies will be bound, complete with index and with advertising pages removed, in a good quality red cloth covered case blocked in gold on the spine.

Home and Overseas readers who wish to have their copies bound are asked to comply with the following instructions :—

(1) Tie the twelve issues (January to December, 1952) securely together before parcelling.

(2) Enclose a remittance for £1 and a gummed label bearing the sender's name and address.

(3) Enclose the copies, remittance and label in a closed parcel and address to :— The Circulation Dept. (E.E. Binding), 28, Essex Street, Strand, London, W.C.2.

(No other correspondence is necessary.)

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The following are also available from our Circulation Dept. :—

A limited number of Bound Volumes for 1951 and 1952. Price, Two Guineas, post free.

Binding Cases for twelve issues. Price 5s., postage 6d.

The Index for Volume XXIV (1952) free.

"Easibind" Cases for binding current issues, or complete volumes. Price 12s. 6d., postage 6d.

PUBLICATIONS RECEIVED

VALVES AND TUBES FOR INDUSTRY AND COMMUNICATIONS is designed to provide electronic equipment designers with a convenient guide to the range of Mullard valves and tubes recommended for communication and industrial applications. In addition to abridged data on all available types, this publication includes general operational recommendations and a guide to American (R.T.M.A.) types replaceable by Mullard valves. It is issued by the Communications and Industrial Valve Department of Mullard Ltd., Century House, Shaftesbury Avenue, London, W.C.2.

TYPES AND GRADES OF BAKELITE PHENOLIC MOULDING MATERIALS has been rewritten to provide a second edition. The booklet mentions new materials introduced during the last five years, and describes selected moulding materials of each grade. Of interest to our readers are the sections on : crack resistant materials for moulding round metal inserts; low loss and tropical electrical moulding materials, and those for electrically conducting or heat resistant mouldings. The booklet is well presented and includes an index. Bakelite Ltd., 18, Grosvenor Gardens, London, S.W.1.

A.C. TO D.C. BY WESTINGHOUSE METAL RECTIFICATION is a publication giving general instructions for the use and application of Westinghouse copper-oxide and Westalite rectifiers. This book is well produced and presented, and contains a great deal of useful information on performance, circuits, loads, classification of applications of metal rectifiers, etc. Westinghouse Brake and Signal Co. Ltd., 82, York Way, King's Cross, London N.1.

TECHNOGRAPH PRINTED CIRCUITS is an attractively produced booklet describing the Technograph system for manufacturing printed circuits. It outlines various applications and compares the printed circuit with conventional methods. A brief review of the economic advantages is also given. Technograph Printed Circuits Ltd., 32 Shaftesbury Avenue, London, W.1.

RADIO INDUSTRY COUNCIL'S SPECIFICATIONS. The third sections of three R.I.C. Specifications have been issued recently, and are for the following: resistors, fixed, wirewound, non-insulated (Specification No. RIC/111); resistors, rotary variable, composition track, with or without switches (Specification No. RIC/122), and capacitors, fixed, ceramic dielectric, grade 1 (Specification No. RIC/133). These sections can be obtained from the R.I.C., 59, Russell Square, London W.C.1., price 5s., 1s. 6d., and 3s. respectively, post free. Complete specifications are also available price 9s., 5s., and 8s. respectively.

PHOTOCELLS FOR INDUSTRIAL APPLICATIONS has recently been issued by the Communications and Industrial Valve Department of Mullard Ltd., Century House, Shaftesbury Avenue, London W.C.2. This booklet has been enlarged and revised, and serves as a guide to the use of the Mullard range of emission photocells developed for industrial applications. It contains notes on the principles of operation of both vacuum and gas-filled cells, together with characteristics, data and suggested applications.

THE MANCHESTER PUBLIC LIBRARIES PATENTS is a brief guide to the patents collection in the Technical Library, compiled by J. E. Wild. It contains: information on patents searches, including U.S., German and Commonwealth patents; instructions to applicants for patents: a list of periodicals containing patents abstracts, and of books on patent law.

W. EDWARDS AND CO. (LONDON) LTD. LEAFLETS NO. E.112/1, C.103/1, AND A.116/1, on vacuum pumping units, "Circseal" glass joints and "Speedivac" educational equipment respectively, were issued recently. The first describes three main types of outfit; mobile units, combined rotary and diffusion pump units and cabinet mounted pumping systems. The second deals with a range of precision ground glass joints incorporating several new design features, while the third details a specially designed schools apparatus enabling twenty basic physics experiments to be demonstrated. W. Edwards and Co. (London) Ltd., Worsley Bridge Road, Lower Sydenham, London, S.E.26.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: December 10. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Lecture: The Production of Television Receivers.
By: Frank Allen.

Scottish Section

Date: December 4. Time: 7 p.m.
Held at: Department of Natural Philosophy, University, Edinburgh.
Lecture: The London to Kirk o'Shotts Television Cable/Radio Link System.
By: J. H. H. Merriman, M.Sc.

West Midlands Section

Date: December 9. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire Technical College, Sulfruna Street, Wolverhampton.
A programme of Technical Films.

North-Eastern Section

Date: December 10. Time: 6 p.m.
Held at: Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: The Development of the Radio and Electronics Industry in India.
By: G. D. Clifford.

Merseyside Section

Date: December 18. Time: 7 p.m.
Held at: The Merseyside and North Wales Electricity Board Service Centre, Whitechapel Liverpool.
Lecture: The Technique of Frequency Measurement in Telecommunications. Part II.
By: H. Hipple, A. Phipps.

BRITISH KINEMATOGRAPH SOCIETY

Date: December 10. Time: 7.15 p.m.
Held at: Gaumont-British Theatre, Film House, Wardour Street, London, W.1.
Lecture: The Quality of Television and Kinematograph Pictures.
By: L. C. Jesty, B.Sc., M.I.E.E., and N. R. Phelp.

BRITISH SOUND RECORDING ASSOCIATION

Date: December 19. Time: 7 p.m.
Held at: Royal Society of Arts, John Adam Street, London, W.C.2.
Lecture: Equalisers, Filters and Tone Control Systems.
By: N. H. Crowhurst, A.M.I.E.E.
Date: December 15. Time: 7.15 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Lecture: Mechanics of Hearing.
By: J. E. J. John.

THE INSTITUTE OF PHYSICS

Electronics Group

Date: December 9. Time: 5.30 p.m.
Held at: 47 Belgrave Square, London, S.W.1.
Lecture: Xero-Radiography.
By: W. D. Oliphant, F.Inst.P.

Industrial Radiology Group

Date: December 12. Time: 6.30 p.m.
Held at: 47 Belgrave Square, London, S.W.1.
Lecture: Science in Industry.
By: R. Carr, M.P.

South Wales Branch

Date: December 6. Time: 2 p.m.
Held at: University College, Swansea.
Branch Annual General Meeting followed by Lecture: Some Properties of Electrical Contacts.
By: Dr. A. Fairweather.

North-Eastern Branch

Date: December 2. Time: 6.15 p.m.
Held at: King's College, Newcastle-upon-Tyne.
Lecture: Some Aspects of Transistor Physics.
By: Dr. H. K. Henisch, A.Inst.P.

THE INSTITUTION OF ELECTRICAL ENGINEERS

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

Informal Meeting

Date: December 8.
Discussion: The Maintenance of Electrical Equipment.
Opened by: R. H. Cobbold, B.A.

Radio Section

Date: December 15.
Discussion: How to Plan a Radio Project.
Opened by: J. Thomson, M.A., Ph.D., D.Sc.
Date: December 3.
Lecture: A Survey of Present Knowledge of Thermionic Emitters.
By: D. A. Wright, B.Sc.

Measurements Section

Date: December 2.
Discussion: Recording Instruments.
Opened by: W. Bamford, B Sc.(Eng.).

Education Discussion Circle

Date: December 10.
Discussion: The BBC School, Evesham, and The Post Office School, Stone.
Opened by: K. R. Sturley, Ph.D., B.Sc., and H. R. Harbottle, O.B.E., B.Sc.(Eng.).

East Midland Centre

Date: December 5. Time: 6.30 p.m.
Held at: Albert Hall, Nottingham.
Faraday Lecture: Light from the Dark Ages, or the Evolution of Electricity Supply.
By: A. R. Cooper.

Cambridge Radio Group

Date: December 9. Time: 8.15 p.m.
Held at: Cavendish Laboratory, Cambridge.
By: R. Theile, Dr. Phil.

Merseyside and North Wales Centre

Date: December 1. Time: 6.30 p.m.
Held at: Liverpool Royal Institution, Colquitt Street, Liverpool.
Discussion: The Impact of Television on Sound Broadcasting.
Opened by: G. Parr, B.Sc.

North-Eastern Radio and Measurements Group

Date: December 1. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: Harmonic Response Testing Apparatus for Linear Systems.
By: D. O. Burns, B.Sc.(Eng.), and C. W. Cooper, B Sc.(Eng.).

Tees-Side Sub-Centre

Date: December 3. Time: 6.30 p.m.
Held at: Cleveland Scientific and Technical Institute, Middlesbrough.
Lecture: Post-Graduate Activities in Electrical Engineering.
By: W. J. Gibbs, M.Sc.(Eng.), D. Edmundson, B.Sc., R. G. A. Dimmick, B.Sc., and G. S. C. Lucas, O.B.E.

North Midland Centre

Date: December 2. Time: 6.30 p.m.
Held at: Huddersfield Training College, Queen Street South, Huddersfield.
Discussion: Technique of Teaching.
Opened by: A. Machennan.
(Education Discussion Circle Meeting.)
Date: December 9. Time: 6.30 p.m.
Held at: Offices of the British Electricity Authority, 1 Whitehall Road, Leeds.
Lecture: Post-Graduate Activities in Electrical Engineering
By: W. J. Gibbs, M.Sc.(Eng.), D. Edmundson, B.Sc., R. G. A. Dimmick, B.Sc., and G. S. C. Lucas, O.B.E.

North-Western Centre

Date: December 3. Time: 6.15 p.m.
Held at: Engineers' Club, 17 Albert Square, Manchester, 2.
Lecture: Electronic Telephone Exchanges.
By: T. H. Flowers, M.B.E., B.Sc.(Eng.).
(Joint Meeting with the North-Western Centre of the Institution of Post Office Electrical Engineers.)

North Scotland Sub-Centre

Date: December 3. Time: 7.30 p.m.
Held at: The Caledonian Hotel, Inverness.
Chairman's Address.
By: L. B. Perkins, B.Sc.

South-East Scotland Sub-Centre

Date: December 17. Time: 7 p.m.
Held at: Heriot-Watt College, Edinburgh.
Lecture: Microwave Radio Links.
By: A. T. Starr, M A., Ph.D., and T. H. Walker, B.Sc.Tech.

South-West Scotland Sub-Centre

Date: December 2. Time: 6 p.m.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Lecture: An Account of, and Discussion on, the CIGRE Conference, Paris, 1952.
By: A. S. Husbands, B Sc., and W. L. Kidd, B.Sc.
Date: December 16. Time: 6 p.m.
Held at: Institution of Engineers and Shipbuilders, Glasgow.
Lecture: Microwave Radio Links.
By: A. T. Starr, M.A., Ph.D., and T. H. Walker, B.Sc.Tech.

South Midland Centre

Date: December 1. Time: 6 p.m.
Held at: James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Colour Television: Some Subjective and Objective Aspects of Colour Rendering.
By: G. T. Winch.
(Joint Meeting with the South Midland Radio Group.)

South Midland Radio Group

Date: December 1. Time: 6 p.m.
See South Midland Centre Lecture.

Rugby Sub-Centre

Date: December 3. Time: 6.30 p.m.
Held at: Rugby College of Technology and Arts, Rugby.
Lecture: Electricity in Newspaper Printing.
By: A. T. Robertson.
Date: December 5. Time: 6.30 p.m.
Held at: Rugby College of Technology and Arts, Rugby.
Faraday Lecture: Light from the Dark Ages, or the Evolution of Electricity Supply.
By: A. R. Cooper.

Southern Centre

Date: December 10. Time: 6.30 p.m.
Held at: The Dorset Technical College, Weymouth.
Lecture: Principles of Colour Television.
By: J. H. Mole, Ph.D.

West Wales (Swansea) Sub-Centre

Date: December 11. Time: 6 p.m.
Held at: Central Public Library, Swansea.
Lecture: Standardization and Simplification in the Electrical Industry.
By: J. T. Moore, B.Sc.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: December 5. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Victoria Embankment, London, W.C.2.
Lecture: The London-Wenvoe Experimental Radio Link for Television.
By: W. J. Bray, M.Sc.(Eng.), A.M.I.E.E., R. L. Corke, A.M.I.E.E., and R. W. White, B.Sc., F.Inst.P., A.M.I.E.E.

THE PHYSICAL SOCIETY

Low Temperature Group

Date: December 4. Time: 5 p.m.
Held at: The Science Museum, South Kensington, London, S.W.7.
Annual General Meeting, and
Lecture: Low Temperature Physics.
By: Professor F. E. Simon.

Acoustics Group

Date: December 8. Time: 5 p.m.
Held at: The Science Museum, South Kensington, London, S.W.7.
Lecture: Methods of Measurement of Elasticity of Solids.
By: G. Bradfield.

PRESENTATION OF TECHNICAL INFORMATION DISCUSSION GROUP

Date: December 12. Time: 6 p.m.
Held at: Anatomy Theatre, University College, London.
Annual General Meeting and Conversazione.

RADIO SOCIETY OF GREAT BRITAIN

Date: December 19. 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: December 11. Time: 7 p.m.
Held at: C.E.A., 164 Shaftesbury Avenue, London, W.C.2.
Lecture: The Television Society's Transmitter—Part 1.
By: H. Banting, H. Fairhurst, and C. Banthorpe.
Date: December 17. Time: 7 p.m.
Held at: C.E.A., 164 Shaftesbury Avenue, London, W.C.2.
Lecture: The Television Society's Transmitter—Part 2.
By: C. Banthorpe, D. N. Cornfield, and E. A. Dedman.

Leicester Centre

Date: December 8. Time: 7 p.m.
Held at: Leicester College of Technology.
Lecture: Electrostatics and Their Applications to the Field of Television.
By: A. Chadfield.

Bristol and South-Western Centre

Date: December 2.
Held at: Carwardines, Baldwin Street, Bristol.
Lecture: The Television Society's 405 Line Transmitter.
By: H. de Laistre Banting.