ELECTRONICS and TELEVISION

TELEPHONE: CENTRAL 7400

& SHORT-WAVE WORLD

TELEGRAMS HULTONPRES, LUD, LONDON

PROPRIETORS: HULTON PRESS LTD., 43-44, SHOE LANE, LONDON, E.C.4

CURRENT ELECTRONIC LITERATURE

Screened Mercury-Electrode Clamps for Dielectric Measurements. (Rushton and Pratt.) Apparatus is described for the accurate determination of the dielectric constant and power factor of insulating materials in the form of discs about 5 cm. diameter. Mercury-electrode clamps are employed, and a guard ring is included. Complete screening is provided since the specimens have usually low capacitances of the order of 30–40 mmF. The electrodes are made of mild steel and fused silica insulation is used. The apparatus is said to be suitable for measurements at frequencies up to 1 Mc./sec. as well as for d.c. insulation measurements.—*Journal of Scientific Instruments*, October, 1940.

Portable Cathode-Ray Oscillograph. (MacFadyen.) A portable oscillograph, employing the 4081 monitor tube and an inexpensive cathode-ray tube with a 2½-in. screen, is described. Although compactness and simplicity are the principal features of the design, sufficient versatility for the majority of laboratory applications is claimed. A circuit diagram is included, and the mounting of the equipment to ensure rigidity and protection from shock is described at some length.

—*Journal of Scientific Instruments*, October, 1940.

Pulsation Welding. (Cogan and Pelton.) An outline is given of the technique of pulsation welding, and recent progress in this field is reviewed. Results of tests on pulsation welded joints are submitted and the following advantages are claimed for this method, especially with spot and projection welding: electrode life is increased; thicker material can be welded in production than was practical before; thicker material can be welded with the same equipment; any tendency of the weld metal to "spit" is reduced; and the finished appearance of the weld is improved. A description is included of a vertical, hydraulically operated, pulsation butt-welding machine, with full synchronous electronic timing control.

—*Welding Industry*, October, 1940.

The Examination of Metals by Ultrasonics (Behr.) A brief account is given of the development of the ultrasonic method of testing metals which the author claims is ready for industrial application. More particularly the author describes the instrumental side of the process and the principles underlying it. A short comparison is made with radiographic examination.

—*Metallurgia*, November, 1940.

Modern Radiography. (Woods.) A recent advance in industrial radiography is the use of metal foils for the elimination of X-ray secondaries and back-scatter. The author discusses drawbacks in the use of calcium tungstate screens. It is claimed that lead screens overcome these difficulties but with loss of intensification of the X-ray image. An effective compromise is possible, according to the author, by using a combination of lead and calcium tungstate screens.—*Iron Age*, Dec. 5, 1940.

A Decade of Progress in the Use of Electronic Tubes. (Ingram and White.) This review is divided into two sections, the first of which deals with electronic tubes used in the field of communications, and the second with tubes used in other fields. The types of tubes in the latter section are classified into groups, each having properties that make it valuable for certain industrial applications.—*Electrical Engineering*, Dec., 1940.

A Method for Detecting the Ionization Point on Electrical Apparatus. (Quinn.) Factors affecting the ionization of an overstressed air path included in the insulating circuit of electrical apparatus are discussed. It is important to know the minimum voltage at which ionization occurs, and the author describes a method for its determination. A cathode-ray oscilloscope is used to detect oscillations introduced into the charging current of the apparatus by ionization.—*Electrical Engineering*, Dec., 1940.

PRINCIPAL CONTENTS MARCH, 1941

Contributions will be promptly considered: they should be accompanied by a stamped addressed envelope for return in case of non-acceptance.

Manufacturers are invited to send details of new productions for review.

101

Electronics
AND
TELEVISION
& SHORT-WAVE WORLD

No. 157. Vol. XIV. MARCH, 1941

EDITORIAL, ADVERTISING AND PUBLISHING OFFICES
Editor : H. CORBISHLEY 43, SHOE LANE, LONDON, E.C.4 Telephone : CENTRAL 7400

Subscription Rates; Post Paid to any part of the world—3 months, 5/– ; 6 months, 10/– ; 12 months, 20/–. Monthly (published 25th of preceding month) 1/6 net. Registered for Transmission by Canadian Magazine Post.

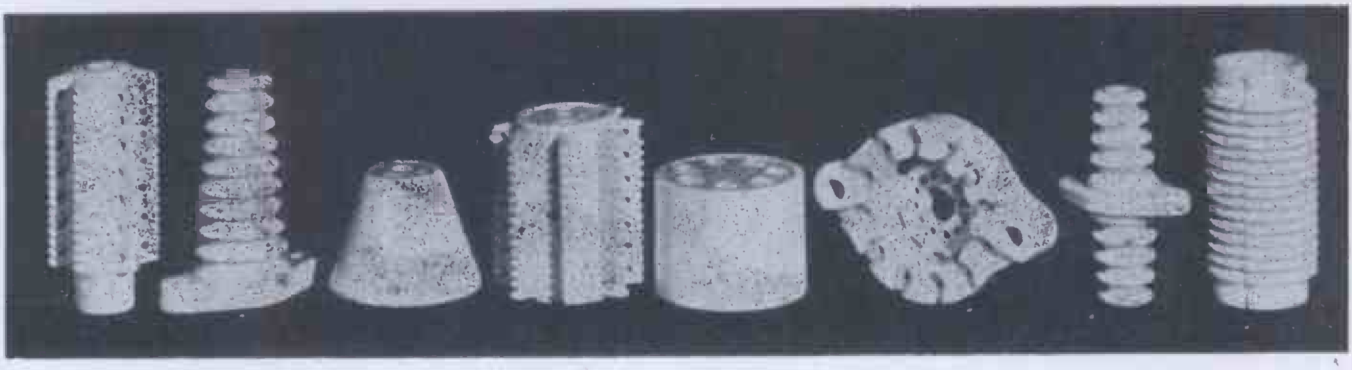
News and Views

IT will not be until after the war is over that the full story of what is taking place in the ultra-high frequency domain can be told, though it is now possible to give some idea of the numerous applications. These include high-fidelity broadcasting, aircraft communication, a variety of uses for the Services, television, facsimile and relay systems. The general advantages include reduced interference, compact and lightweight transmitting apparatus, and the use of low powers for medium ranges. The possibilities which the ultra-high frequencies are opening up are, therefore, enormous and they are the subject of most intensive research. Technique differs considerably from ordinary radio engineering and an aspect of prime importance is dealt with in this issue in the article on the development and use of low-loss ceramics. This article, which will be published in three instalments, we believe to be the most exhaustive yet publishd on this particular subject. It is hoped to follow this up with others of a similar nature in order that readers may be made acquainted with all developments which it is possible to disclose at the present time.

It is probable that shortly after this issue is published an announcement will be made regarding limitation of valve supplies. Important decisions have been made which are primarily due to the demands of the Services, and it is inevitable that some hardship will be experienced both as regards replacement valves and new receivers. These decisions which have only been arrived at after considerable discussion by all the interests concerned have been absolutely essential, and both trade and public should regard the matter in that light.

Engineers of the Du Mont company are meeting with considerable success in their efforts to cut the television picture repetition rate without objectionable flicker. As has been reported in this Journal, the basic idea is the use of the delayed-action or memory-type cathode-ray tube screen, which holds over the electronic impression from one complete scanning or frame to the next, thereby bridging the usual dark interval and minimising flicker. The memory screen, it is claimed, can be operated at 15 pictures per second with no apparent increase in flicker. Halving the picture repetitive rate means that the elbow room gained on the air may be used for increasing the number of scanning lines, or, alternatively, transmitting a picture of 325-line definition and 15 pictures per second in a band only 1,000,000 cycles wide. This, it is contended, would make medium-range television transmission feasible, and it should be possible to have television stations covering large areas which are not at present in the service area. Even medium range overseas services would be possible. From the original orange-tint delayed-action screen, Du Mont engineers have now succeeded in evolving a good white-and-black image comparable with that of the usual television screen. Also with a still greater memory or delayed action, one image is now blended so fully with the next that the dark interval is further minimised with virtually no flicker.

With relatively simple auxiliary equipment, plane-polarised light, circularly-polarised light and elliptically-polarised light may be readily demonstrated on the cathode-ray screen. It is possible to produce the patterns representing every possible state of polarisation of light (or of ultra-high frequency radio waves) by merely manipulating three controls representing the three basic elements of a generalised ellipse, i.e., the magnitude of two perpendicular components and the phase difference between them. We hope to publish some information on this matter in an early issue.



Low-loss Ceramics for Radio-frequency Use

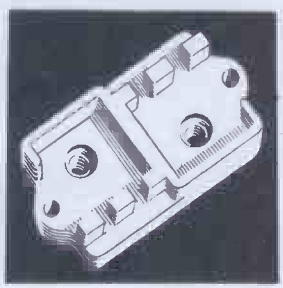
By G. P. Britton, D.I.C., A.C.G.I., D.Sc. (Eng.), of Steatite and Porcelain Products Ltd.

VAST strides have been made in radio-frequency communication technique during the past twenty years, and the pace of development is far from showing any sign of slackening. So far as the normal well-established high-frequency bands are concerned, transmitter powers are steadily increasing and frequency stability demands becoming progressively more stringent; corresponding improvements in selectivity and stability are called for in the associated receiving apparatus. Little stress is needed on the difficulties of providing similar high standards of performance in equipment for the Services, much of which may have to function under most unfavourable conditions of temperature variation and vibration.

The tendency to exploit still higher frequencies is evident from current literature, and while, owing to war restrictions, we shall inevitably learn little of the tremendous strides which will be made in this direction during the war period, we may expect them to exert a profound influence on broadcasting, television and communication in the post-war period.

Side by side with progress in circuit design and vacuum-tube technique must come improvement in materials, and among the most vital of these are insulating materials and dielectrics. Ceramics are now generally accepted as possessing to a very high degree the properties required of insulating materials in high-frequency apparatus—mechanical stability and extremely low electrical losses, with permanent maintenance of these properties. It is hoped, therefore, that the following outline of the nature of these ceramics, the methods of manufacture, coupled with a discussion of the limitations in manufacturing technique and necessary tolerances arising therefrom, will be of assistance to those designers who are as yet unfamiliar with these materials.

This is the first of a series of three articles on the development, manufacture and use of low-loss ceramics. They will provide information which hitherto has not been published but which is of vital importance to those engaged in the design of apparatus operating on ultra-high frequencies in which permittivity must be high and dielectric loss low.



The term "ceramic" is probably derived from the Greek word "keramos," meaning "burnt stuff." The term is commonly applied to those materials moulded from clay and hardened by the action of a high temperature. While clays are the major constituents of a large proportion of ceramic mixtures, many of the modern ceramic materials contain clay only as a minor constituent. This is especially true of most modern ceramic electrical insulators. The distinguishing feature of a ceramic material is the high temperature essential for its manufacture. By high temperature is meant the range from red heat up to 1,600° C. and even higher.

Ceramic materials are therefore essentially heat-resisting and so they must be inorganic in nature. This feature distinguishes ceramic materials from the organic plastic materials which are moulded at a low temperature, and are, of course, not heat-resisting. The heat-resisting and non-combustible nature of ceramic materials makes them especially suitable as electrical insulators.

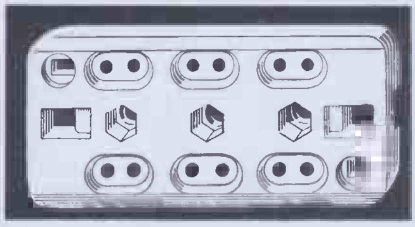
Modern commercial low-loss ceramics divide into two groups, one consisting basically of soapstone or talc, the other basically of titania. Both groups were origin-

Low-loss Ceramics for Radio-frequency Use—*Britton*

ally developed in Germany, but large-scale bulk manufacture has been established in England for some seven years and more recently in the United States.

Soapstone Base Materials

Soapstone and talc are both naturally occurring magnesium hydrosilicates of the approximate composition of $3\ MgO.4SiO_2.H_2O$.



Extensive deposits exist in many parts of the world, differing mainly in physical properties as the result of the conditions under which they are formed, varying from quite a hard, rock-like material with a close-grained structure to a very soft, soapy material. A large deposit of a natural soapstone of close and even texture occurs in Bavaria. This is very easily machined and has for many years been employed in Germany for the manufacture of gas-burner tips and later for small electrical insulators.

Small articles made from this material are mechanically quite strong when they have been fired to a temperature of 900° C., but are slightly porous. There was inevitably much waste in the machining operations and the natural steps followed of bonding the crushed waste soapstone with clay and shaping articles from the plastic mixture. This development led to the production of a non-porous compounded steatite body, manufactured by methods similar to those used for porcelain, and consisting substantially of natural soapstone with small amounts of clay and alkali flux necessary for vitrification.

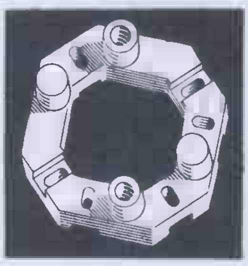
These latter constituents also affect profoundly workability of the unfired body. On firing to a temperature of some 1,400° C., the soapstone becomes a dense mass of crystals—clinoenstatite ($MgO\ SiO_2$) and cristobalite, embedded in glassy matrix. This material was, and is, extensively used for high-grade electrical insulators, as it has properties superior to the best electrical porcelain, as will be seen from the following table, which gives a comparison with "hard" (*i.e.*, high fired) porcelain:—

	Hard Porcelain	*Steatite*
Dielectric strength (Kv/mm)	25 to 30	22 to 27
Volume resistivity at 20° C.	10^3 to 10^4	10^4 to 10^5
at 300° C.	10^6 to 10^7	10^8
Tensile strength lb./in²	4,200 to 5,400	7,800 to 12,000
Compressive strength lb./in²	64,000 to 78,000	120,000 to 130,000
Impact bending strength ft. lb./in² ...	0.84 to 0.98	1.6 to 2.1

The higher mechanical strength and toughness of steatite are of particular value for insulating parts which are subject to rough treatment during assembly with metal fittings, particularly where riveting operations are involved. The higher cost of steatite mouldings is then largely offset by reduced assembly costs due to the minimizing of breakage.

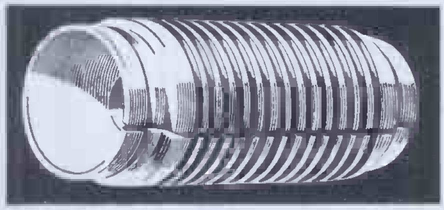
Porcelain has long been used for insulators in high-power radio transmitters, and with the advent of high-power transmission on short waves heating of these insulators often became serious. In Germany it was noticed that the temperature rise of steatite insulators was less when subjected to similar high-frequency stress, indicating considerably reduced dielectric losses. Quantitative investigation showed that the steatite body possessed a power factor of the order of 20×10^{-4} as against 60 to 100×10^{-4} for electrical porcelains.

As the technique of working steatite was already firmly estab-



lished in Germany, its use for radio-frequency purposes soon developed extensively. Further investigation showed that the power factor of this type of material could be reduced by correct selection of the fluxing constituents.

Another line of development was the enrichment of the magnesia content of the body, but the abnor-



mally high firing shrinkage of this type of material rendered it extremely difficult to manufacture, dimensional accuracy and freedom from warping being in particular difficult to attain. Some of the early German materials of exceptionally low losses were therefore not widely commercialized and little was seen of them in Germany in the two years immediately preceding the outbreak of hostilities.

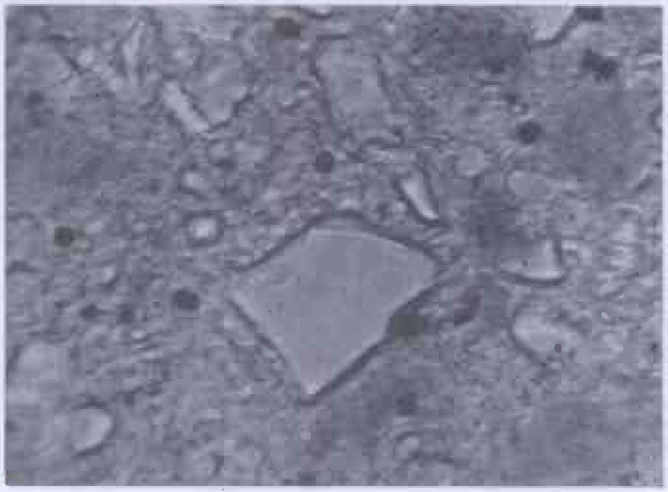
The manufacture of steatite in this country was established some twelve years ago, and development of low-loss materials followed. The electrical performance of two of these—Frequentite and Frequentite "S"—is compared in the following table with steatite and quartz, which latter material is generally accepted as the best available high-frequency insulating material. It will be seen that Frequentite "S" very closely approaches quartz in performance, and has, of course, the great advantage that it can be worked easily to give insulators of relatively complex shape, which is almost an impossibility with quartz, except at prohibitive cost.

It is of interest that the ceramic materials exhibit a slight but pro-

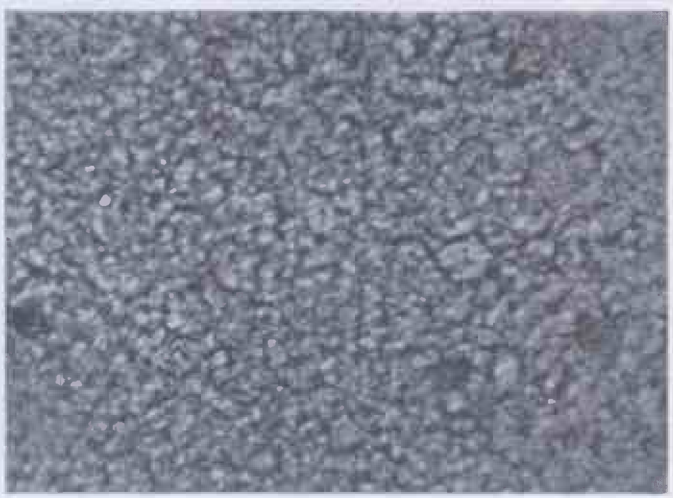
	Quartz	*Steatite*	*Frequentite*	*Frequentite "S"*
Power factor x 10^4 at 1 mc/s ...	1.0	21	6.5	1.2
at 3 mc/s ...	1.0	20	5.9	1.0
at 10 mc/s ...	1.0	17	5.5	—
Permittivity	3.9	6.1	6.1	6.0
Effective temperature coefficient of capacity (parts/million/°C.) ...	—	+ 280	+ 100 to + 130	+ 120

gressive reduction in power factor with increase in frequency, whereas most of the plastics increase in power factor, in some cases to a marked extent. Reliable measurement of power factors of the order of 1×10^{-4} is extremely difficult at frequencies of 100 mc. and above, and figures are therefore not quoted: comparative results obtained so far indicate that the low power factors are at least maintained.

The lower dielectric loss of these materials by comparison with, for example, porcelain cannot be explained either in terms of ohmic resistance or by the polarization movement of molecules, neither factor being appreciable in ceramics at room temperature; degree of homogeneity in structure is a more probable cause. Pure quartz with the low power factor of 1×10^{-4} is extremely homogeneous in structure. Frequentite contains only some 10 per cent. of glassy phase and some 20 per cent. of crystalline phase. The homogeneous crystal cluster structure is well shown in the accompanying photograph, contrasting with the much more heterogeneous structure of porcelain.



Microstructure of hard porcelain. Magnification approx. 500 diameters.



Microstructure of Frequentite. Magnification approx. 500 diameters.

Titania Type Material

The soapstone base materials described above, while primarily intended for purely insulating parts, are used very successfully as dielectric materials in condensers for high-frequency use. For this purpose a higher dielectric constant is desirable, and in this connection the properties of rutile, one of the crystalline forms of titanium dioxide, have been turned to good account. It has long been known that this mineral has a dielectric constant of 90 and 170 approximately along the respective optical axes of the crystal. Ceramic materials have accordingly been developed consisting of a high percentage of titania with the necessary minimum of clays added to enable the material to be worked.

In the kilning process there is extensive recrystallization or crystal growth. The final product is a dense ceramic material consisting of a mass of crystals, oriented at random, with a very small amount of glassy matrix, and exhibiting an extremely homogeneous microstructure: this latter contributes largely to the low dielectric loss. The properties of Faradex, a commercial British ceramic of this class, are given in the table below:

PROPERTIES OF FARADEX.

Power factor $\times 10^4$ at 1 mc/s ...	3.3
at 3 mc/s ...	3.0
at 10 mc/s ...	2.7 to 3.0
Permittivity	80
Effective temperature coefficient of capacity (parts/million/°C.) ...	-750 to -820
Dielectric strength (Kv/mm) ...	10 to 12
Tensile strength (lb./in.)	8,400 to 11,600
Coefficient of linear expansion ...	approx. 8×10^{-6}

It will be appreciated from the above figures that this class of material is particularly suited to the production of compact condensers of small power loss. There seems little practical value in separating capacity changes due to permittivity variation from those due to dimensional changes, and the effective temperature coefficient of capacity realized from dielectrics with rigidly attached electrodes (*i.e.*, fired on silver) is therefore quoted.

The large negative temperature coefficient is often of little importance, and in some cases may be turned to good account. However, for some purposes dielectrics with very low temperature coefficients are necessary. This can be achieved by suitable variation of the titania content. In fact, a whole series of materials can be realized with temperature coefficients ranging from 800 parts/million/°C. of the high titania material to the plus 120 parts/million/°C. of the soapstone materials, the permittivities varying correspondingly.

A typical low-temperature coefficient material of this nature of this series has a permittivity of 12, power factor 15 to 30×10^{-4} and a temperature coefficient of 0 to 20 parts/million/°C. It should be noted that the intermediate materials of this series generally have much higher dielectric loss characteristics.

There are, however, certain titania-magnesia materials with small positive temperature coefficients which in contrast possess outstandingly low losses. The good properties of these materials have commonly been ascribed to the formation during firing of a high proportion of magnesium titanate. A commercial product of this class is Tempradex, figures for which are given below:—

Power factor	0.7 to 1.0×10
Permittivity	15
Effective temperature coefficient of capacity (parts/million/°C.) ...	+60 to +80

The next article in this series will deal with manufacturing methods employed in the production of ceramics, and a discussion of tolerances.

The Use of Headphones with Communication Receivers

A COMMUNICATION receiver is one in which sensitivity, selectivity, and frequency stability have been brought to a remarkably high degree of perfection. Many receivers of this nature have a switch so that the A.V.C. voltage can be cut out, and the maximum sensitivity obtained. Under these conditions the acoustic output from the receiver is approximately proportional to the aerial signal input. Thus, in tuning over the various bands, strong signals will create a large output from the loudspeaker, which, at certain times, may cause annoyance to others. The use of headphones and automatic silencing of the loudspeaker is, therefore, desirable under some conditions.

By using sensitive headphones which are light in weight and comfortable to wear for long periods, such as those made by Ericssons, the listener can tune to very weak signals which are inaudible on the average loudspeaker. No major alteration is necessary to the receiver, and as one less valve stage is necessary, background noise will be reduced.

The average communication receiver is terminated with a pentode power valve which may have negative feedback. It is not desirable to disconnect the loudspeaker from the secondary of the output transformer as the pentode would produce very high transient voltages across the primary winding, with resultant damage. The best way of silencing the loudspeaker is to short circuit the grid input load of the pentode valve, preferably by means of a jack, so that when the headphone plug is inserted the output valve is automatically static.

Fig. 1 shows one circuit arrangement which will meet these requirements.

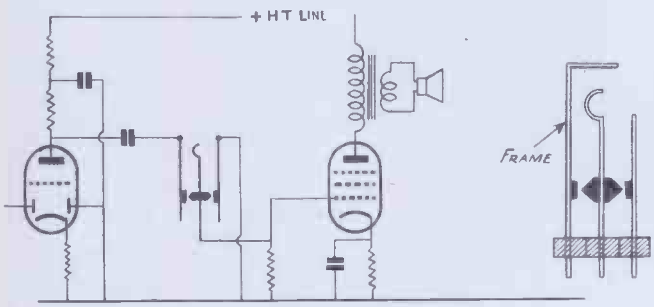


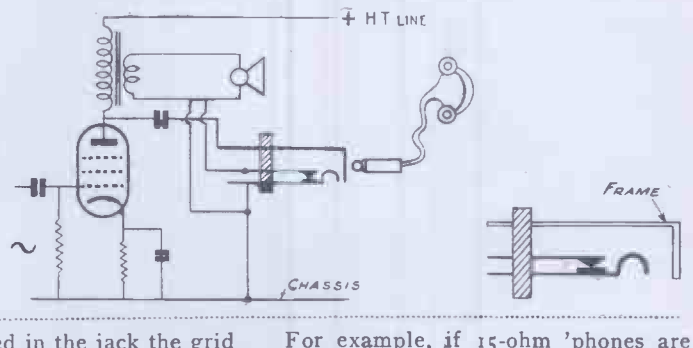
Fig. 1. In this arrangement when the phone plug is inserted the output valve is made static.

It will be seen that when the 'phone plug is not inserted in the jack the grid circuit is completed through the springy centre contact making connection with the frame. When the 'phone plug is inserted it becomes virtually the anode load of the penultimate stage, the grid resistance is short circuited and the grid of the output stage earthed. This circuit necessitates the use of high-resistance headphones and the 4,000-ohm type should be employed.

The 'phones are at earth potential so that no danger from shock can occur as is the case when they constitute part of the anode circuit.

If it is felt that the additional amplification of the output stage is desirable, the alternative scheme shown in Fig. 2 can be used. In this diagram the 'phones, which are again of the high-resistance type, are fed through a 2 μF. condenser capable of withstanding the anode voltage plus the peak signal voltage—usually a 400-volt working type will be adequate. By means of the jack contacts the loudspeaker connections are open-circuited when the phones become the load impedance of the pentode stage.

Fig. 2. This arrangement allows the additional amplification of the output stage to be used.



Suitable two-way plugs and jacks are obtainable from A. F. Bulgin and Co., Ltd., and the alterations to the wiring are quite simple. If low-resistance headphones are used then a suitable matching transformer is necessary; in the case of the first circuit a reflected load of 50,000 ohms is desirable and in the second case one of 8,000 ohms. The transformer ratio is readily calculated from the formula.

$$\sqrt{\frac{\text{Valve load}}{\text{phones impedance.}}}$$

For example, if 15-ohm 'phones are used in both cases, then the two transformer ratios are:

$$(1)\ \sqrt{\frac{50{,}000}{15}} = 57:1$$

$$(2)\ \sqrt{\frac{8{,}000}{15}} = 23:1$$

The nearest commercial values of (1) 50 : 1 and (2) 20 : 1 would be suitable for these purposes.

The Figure of Merit of H.F. Valves

By C. LOCKHART

While it is generally realized that for a given mutual conductance the lower the inter-electrode capacities and the higher the input resistance of a valve the better it performs as an H.F. voltage amplifier, there is no agreed criterion for comparing the merits of valves used in various types of circuits. It is the purpose of this article to suggest a Figure of Merit rating by means of which the most suitable valve may be chosen for a particular wide-band amplifier circuit.

THOUGH in some applications the H.T. and L.T. power required by a valve may be of paramount importance to base the Figure of Merit solely on the basis of the voltage amplification obtainable.

Figure of Merit (F.O.M.) will be derived for the following types of circuit:—

1. Video Frequency Amplifier.
2. Wide Band R.F. Amplifiers with staggered Tuned Anode coupling circuits.
3. Wide Band R.F. Amplifiers with Transformer coupled circuits.

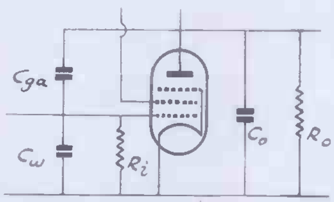


Fig. 1. Illustrating the various valve constants referred to in this article. The mutual conductance is "g." and the anode A. C. resistance "Ra."

4. The Amplifying Valve following the aerial coupling circuit (i.e., first H.F. valve of wide-band receiver).

Amplifiers consisting of similar valves in cascade are assumed in all cases.

Fig. 1 is a simplified diagram of a screened pentode and illustrates the parameters relevant in wide-band amplifier considerations, Cw is the total effective input capacity including hot capacity and space charge capacity (for a discussion of the individual components of the input capacity reference should be made to the author's article in the August issue on the Gain Control of Radio-frequency Amplifiers). Also Ri is the effective input resistance, Co is the total effective output capacity and Ro the effective (H.F.) A.C. anode resistance. Cga is the residual anode-to-grid capacity.

(b) That Ro is sufficiently large compared with Ri for its effect to be neglected. (Ro in typical high-slope H.F. pentodes is usually greater than 50,000 ohms at 45 Mc.)

It will be seen from the expressions derived below that the gain obtainable from all the circuits considered is greater

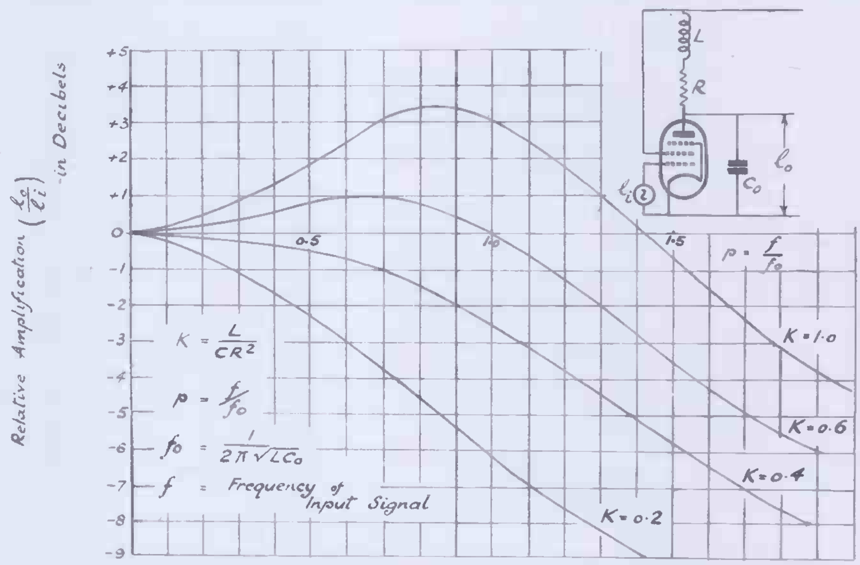


Fig. 3. Variation of the frequency response of the amplifier of Fig. 2 with the circuit parameters.

For simplicity the following assumptions which are usually justified in practice will be made:—

(a) That Cga is sufficiently small for its effect to be neglected.

the lower the capacity across the input and output coupling circuits.

For this reason, in order to maintain the highest possible gain for a given band width it will be assumed that only the electrode capacities will be used as terminations and that any trimming that may be required will be carried out by adjusting the inductance valves either by means of an adjustable low-loss iron core or an adjustable copper disc.

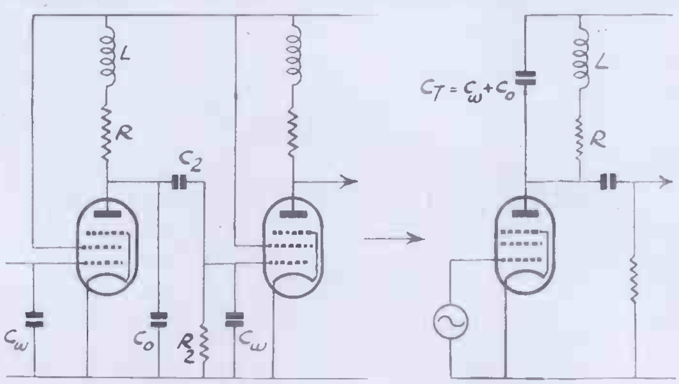


Fig. 2. Inductance compensated R.-C.-coupled video amplifier.

1. *Video Frequency Amplifiers*

Fig. 2 illustrates the simplest and most widely used circuit for television video amplifiers. With the band widths normally employed Ri is very large compared with R, and can therefore be neglected. The performance of this cir-

Figure of Merit of H.F. Valves—*Lockhart*

cuit is then completely defined by the two parameters $K = \frac{L}{C_T R^2}$ and $p = \left(\frac{f}{f_o}\right)$ where $f_o = \frac{1}{2\pi\sqrt{LC_T}}$

In Fig. 3 curves are given which show how the relative amplification varies with the parameter "p" for different assigned values of K. To obtain an absolute frequency scale we use the expression

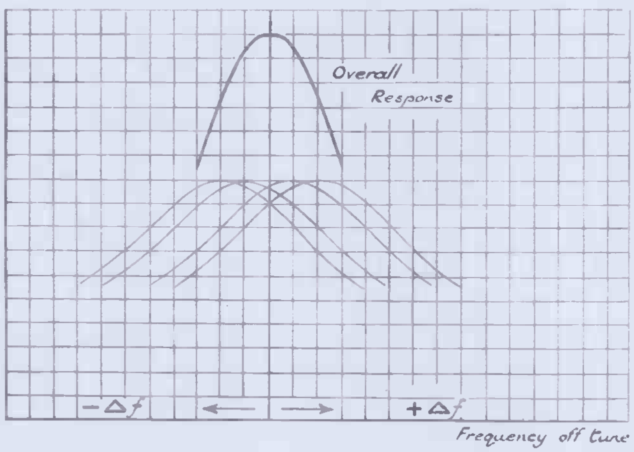


Fig. 4. Use of staggered circuits to provide level response with adequate selectivity.

$$f_o = \frac{1}{2\pi G_T R\sqrt{K}} \qquad (1)$$

which is derivable from the above equations.

As a criterion of performance we take the stage amplification at moderately low frequencies. This is

$$A = \frac{e_2}{e_1} = gR$$

$$= \frac{g}{2\pi C_T f_o\sqrt{K}} \qquad (2)$$

As far as any given performance specification is concerned, fo and $\sqrt{K}$ are fixed and we can express the F.O.M. for convenience as

$$\text{F.O.M.} = \frac{100g}{C_T}$$

$$= \frac{100g}{C_w + C_o} \qquad (3)$$

2. *Amplifiers with Single Staggered Coupling Circuits*

In television I.F. and H.F. amplifiers a reasonable level overall response is sometimes obtained by staggering the tuning of a large number of individual tuned circuits on each side of the mid-band frequency. The overall response is then the product of the individual responses of all the tuned circuits. This is shown in Fig. 4 for the case of four tuned circuits. In practice, a total of more than four tuned circuits would be employed so as to give a better overall response with adequate selectivity.

Taking first the case of an I.F. amplifier with similar tuned anode couplings between stages, as shown in Fig. 5, the band width that can be obtained from a given number of circuits and for a given permissible variation of output in the pass-band is a function of the shape of the response curve of the individual circuit.

If in the case of a single tuned circuit we express as "M" the ratio of the response at a frequency (Δf) off tune to the response at the resonant frequency (fo) of the tuned circuit. Then

$$\frac{2Af}{f_o} \doteq \frac{1}{Q}\sqrt{\frac{1-m^2}{m^2}} \doteq \frac{1}{QF} \qquad (4)$$

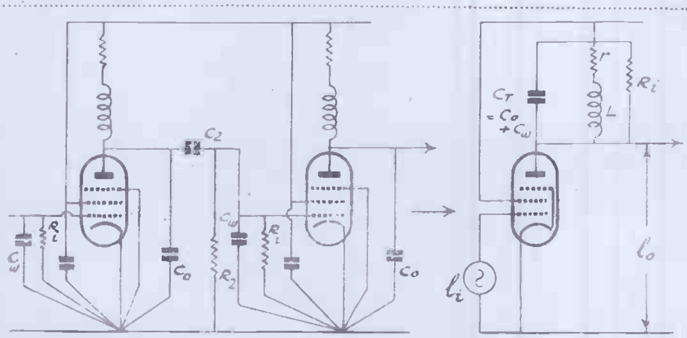


Fig. 5. Diagram showing two stages of a "staggered" tuned-anode amplifier and their equivalent circuit.

where $F = \sqrt{\frac{m^2}{1-m^2}}$

Q = Magnification of circuit
and $2\Delta f$ = Total Band Width

now Amplification $= A = \frac{e_o}{e_i} = g\,R_{dyn}$

where R_{dyn} = Working Dynamic Resistance of the tuned circuit

$$R_{dyn} = \frac{Q}{2\pi f_o C_T} = \frac{1}{2\pi F G_T(2\Delta f)} \qquad (5)$$

$$2\Delta f = \frac{1}{2\pi F C_T R_{dyn}} \qquad (6)$$

The F.O.M. in this case should be a measure of the gain obtainable for a given band width. Therefore, F can be taken as constant for a given set of conditions and

$$A \times (2\Delta f) = \frac{g}{2\pi F C_T}$$

$$\text{F.O.M.} = \frac{100g}{C_2} = \frac{100g}{C_w + C_o} \qquad (7)$$

This F.O.M. is seen to be the same as in the case of the video amplifier and is applicable to all cases where Ri is greater than the value of Rdyn required by equation (5) as is the case in television I.F. amplifiers. In the case of H.F. amplifiers or "straight" receivers employed at 45 Mc. the condition that Ri > Rdyn may not always occur and therefore a stipulation must be made that for the F.O.M. to be strictly applicable, that

$$C_T R_i > \frac{1}{2\pi F(2\Delta f)} \qquad (8)$$

if we take for example $m = 0.8$ then $F = 1.33$ for $(2\Delta f) = 4$Mc

$$C_T R_i > 30{,}000 \qquad (9)$$

where C_T is in $\mu\mu$F and Ri in ohms.

3. *Amplifiers with Transformer Coupled Circuits*

By the use of transformers with tuned primaries and secondaries as the coupling elements of a wide-band amplifier, a more level response with considerably greater gain may be obtained than in the case of staggered single circuits. This type of coupling which is shown in Fig. 6 is in effect a form of filter and it can be shown that

$$A\,(2\Delta f) \propto \frac{g}{\sqrt{C_w C_o}} \qquad (10)$$

Figure of Merit of H.F. Valves—*Lockhart*

therefore, the F.O.M. may be written as

$$\text{F.O.M.} = \frac{100g}{\sqrt{CwCo}} \qquad (11)$$

From the above equation it will be seen that for a given value of CT the highest F.O.M. with transformer coupling will be obtained when Cw and Co are as dissimilar in value as possible, the lowest F.O.M. being obtained when Cw = Co.

As in the case of single tuned circuits an overriding stipulation as to the minimum permissible value for Ri must be satisfied for the F.O.M. to be applicable. Thus we have that

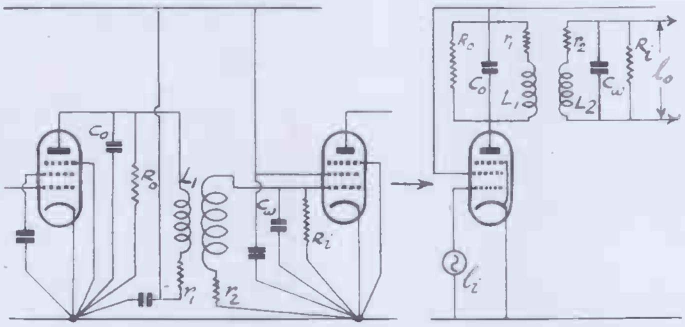


Fig. 6. Transformer-coupled wide-band amplifier and its equivalent circuit.

$$CwRi > \frac{1}{2\pi(2\Delta f)} \qquad (12)$$

In the case of the I.F. amplifiers of television superhet receivers with a pass-band of 4 Mc., the condition that

$$CwRi > 40{,}000 \qquad (13)$$

is easily satisfied, as Ri is very large even at 15 Mc.

As the H.F. stages of television superhet receivers are usually required to amplify both the sound and vision channels a larger nominal pass-band than 4 Mc. will be required and therefore a lower minimum value for Cw Ri will be satisfactory.

The majority of television superhet receivers, however, employ only one H.F. stage and in this case the following frequency changer consists of a similar valve with the heterodyne voltage injected into its grid circuit. A valve operating under these conditions with the right order of heterodyne voltage will have a working input capacity only very slightly higher than its cold capacity, and an input resistance of the order of 30% to 100% greater than when the valve is used as an amplifier. The greater the heterodyne voltage the larger the input resistance. The conversion conductance, however, is only about 30% to 40% of the mutual conductance at the normal operating point as an amplifier.

In practice, the Cw Ri product of a valve used as a frequency changer with grid injection will be higher than the Cw Ri product of the same valve used as an H.F. amplifier. The stipulation of equation (13) therefore safely covers all practical cases with something in hand.

4. *First H.F. Valve*

Fig. 7 illustrates a typical short-wave aerial coupling system and its equivalent circuit. Here we have a dipole aerial with an effective resistance of Ra ohms coupled, through a transmission line which has a characteristic impedance equal to Ra, to the primary of the aerial tuned circuit. By suitably adjusting the coupling between the primary and secondary, the transmission line can be terminated in its characteristic impedance at the mid-band frequency. The two condensers in series with the primary ensure that the termination will be purely resistive at the mid-band frequency.

This method of coupling is essential in television receivers in order to prevent reflected waves occurring in the feeder line.

When the above coupling conditions (which provide maximum energy transference) are satisfied, a voltage gain from dipole to grid of first H.F. valve of

$$A = \frac{1}{2}\sqrt{\frac{R_D}{R_A}} \qquad (14)$$

is obtained. Where RD is the effective dynamic resistance of the tuned secondary including the shunting action of the valve input resistance, but with the feeder disconnected. As before, we have

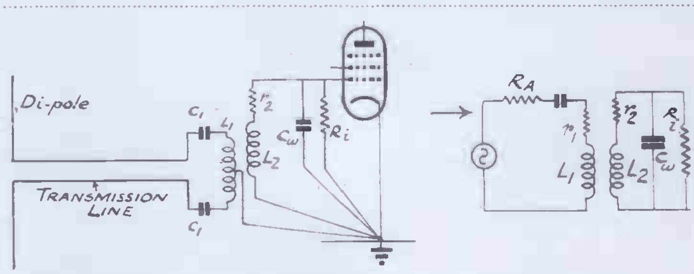


Fig. 7. Equivalent circuit of the aerial stage of a short-wave receiver.

$$2\Delta f = \frac{1}{2\pi FCwRd} \qquad (15)$$

$$Rd = \frac{1}{2\pi FCw(2\Delta f)} \qquad (6)$$

Where Rd is the effective dynamic resistance of the secondary including the damping effect of the aerial and valve, $Rd = \frac{1}{2}$ RD.

Now the total amplification is the product of the voltage step-up of the aerial tuned circuit and the amplifying ability of the first H.F. valve. As the latter factor is proportional to " g " we have

$$A(2\Delta f) \propto \frac{g}{Cw\sqrt{R_D R_A}}$$

$$\propto g\sqrt{\frac{2\Delta f}{CwR_A}} \qquad (17)$$

$$\text{F.O.M} \propto A(\sqrt{2\Delta f}) \qquad (18)$$

$$\text{F.O.M} = \frac{100g}{\sqrt{Cw}} \qquad (19)$$

with the overriding stipulation that

$$CwRi > \frac{1}{\pi FCw(2\Delta f)} \qquad (20)$$

In the case of television receivers, the aerial transformer is usually required to pass the sound channel as well as the vision band so that (2Δf) has to be made appreciably greater than 4 Mc.

If we make (2Δf) = 7 Mc. m = 0.8 F = 1.33

$$CwRi > 34{,}000 \qquad (21)$$

It should be realized that in all cases where a limitation is put on the minimum value of the product CRi there is no big advantage in making it very much greater than the limit unless low-loss coils cannot be employed. This is particularly true in cases where the improvement can only be made at the expense of a big drop in the value of the F.O.M. The main advantage of a higher value than specified for the product CRi is that it is then possible to shunt the tuned circuit with a fixed resistance and thus reduce the effect of the variation in valve characteristics on

Figure of Merit of H.F. Valves—*Lockhart*

the frequency response of the receiver. A lower value than specified for the product CRi will leave the F.O.M. unaltered, but result in a lower attenuation at the boundaries of the specified band width at the expense of lower gain.

The value of Ri at any given frequency can always be appreciably increased by operating at a lower value of mutual conductance (g). The procedure to be followed in determining the F.O.M. for any given set of operating requirements is to first calculate the necessary value for Ri to give the required CRi product. From the data on the valve we can then obtain the resulting values of Cw, Ri and g. Thus, for example, if we take the case of the SP41 for Cr Ri = 50,000, we have that Cw = 14.8 $\mu\mu$F. and g = 7.6 mA/v and Ri=2,500 w. For tuned anode operation the F.O.M. is then reduced from the value of 42 given in Table 1 to 38. If a still higher value of Ri is necessary then the F.O.M. will be further reduced.

Table 1 gives the F.O.M. and Ct Ri product for one typical medium-slope Broadcast V.M. pentode, the V.P.41 and four well-known high-slope television pentodes as well as an Acorn pentode. The first point that strikes one is the very poor showing made by the Broadcast V.M. pentode and the Acorn; the fact is that Acorns of the 954 size should not be used for frequencies as low as 45 Mc. The main advantages of the Acorn are (a) its very high input resistance and (b) its low electrode capacities. It should therefore only be used at frequencies where these two features are essential and give it a better performance than other valves. This would amount to operating the Acorn at frequencies of the order of 100 Mc. and over. The only justification for using an Acorn at lower frequencies is in the special case where it is necessary to tune the receiver over an appreciable wave range. The low capacities of the Acorn would then be of considerable help.

Actually, the picture given in Table 1 is rather favourable for the lower slope lower capacity valves, as the stray wiring and component capacities which are bound to be present in practice have not

TABLE 1

Valve Type	g MA/V	Cw $\mu\mu$F	Co $\mu\mu$F	CT $\mu\mu$F	Ri ohms at 45Mc	Figure of Merit as: Video and Tuned Anode Amplifier	Figure of Merit as: Transformer Coupled Amplifier	Figure of Merit as: First H.F. Amplifier	CT Ri ohms $\mu\mu$F at 45 Mc
Mazda VP41	2.0	7.8	11.5	19.3	10,000	10.4	21	71.5	193,000
Mazda SP41	8.5	15	5.25	20.25	2,200	42	96	220	4,4500
Mullard EF50	6.5	10	5.3	15.3	5,000	42.5	89.5	206	76,500
Osram Z62	7.5	13.5	8.0	21.5	3,150	35	72	204	68,000
R.C.A. 1851 and 1852	9.0	13.5	5	18.5	3,400	48.5	109.5	266	63,000
R.C.A. 954 Acorn	1.4	3.6	3.0	6.6	90,000	21	42.5	74	595,000

Figure of Merit ratings for different types of circuit excluding effect of stray capacities.

TABLE 2

Valve Type	Figure of Merit as: Video and Tuned Anode Amplifier	Figure of Merit as: Transformer Coupled Amplifier	Figure of Merit as: First H.F. Amplifier	(Cw+2.5) Ri ohms $\mu\mu$F at 45 Mc.
Mazda VP41	8.8	18.2	62.5	103,000
Mazda SP41	35.8	80.5	204	37,500
Mullard EF50	34.6	73.5	184	62,500
Osram Z62	30	63	188	50,500
R.C.A. 1851 and 1852	41	91	225	54,500
R.C.A. 954 Acorn	13.9	29	57	550,000

Figure of Merit ratings for different types of circuit including effect of stray capacities.

been allowed for. These strays will produce a bigger percentage gain reduction on the smaller low capacity valves. This statement should not be taken as setting a premium on high capacity as such. If we take, for example, a given valve and compare it with another having characteristics equivalent to two such valves in parallel, then the double slope, twice the capacity valve will have its performance less affected by stray wiring and component capacities.

By taking great care in the component and wiring lay-out the stray capacities can be reduced to comparatively low values. The following values have been assumed in compiling Table 2, which gives the F.O.M. after allowing for the effect of the stray capacities on the values given in Table 1:—

(a) Video Amplifier and Tuned Anode Amplifier Total Strays $3\frac{1}{2}$ $\mu\mu$F.
(b) Transformer Coupled Amplifier Anode Circuit Strays $1\frac{1}{2}$ $\mu\mu$F. Grid Circuit Strays $1\frac{1}{2}$ $\mu\mu$F.
(c) Aerial Coupling Circuit $2\frac{1}{2}\mu\mu$F.

A critical examination of the simple expressions for F.O.M. derived above might lead to criticism of the expression for the F.O.M. of the first H.F. valve. It will be seen that this expression does not contain the Co of the first valve, yet the gain of the first stage will be a function of Co. The lower Co the higher the gain. The reason for leaving out the effect of Co in expression (19) was that its effect varies with the type of anode coupling circuit employed. Only in the case of the widely used television receiver circuit having all stages coupled by tuned transformers and employing similar valves in all stages, can a F.O.M. rating for the first H.F. valve be obtained that is truly representative of the individual gain of that stage. This is

$$\text{F.O.M} \propto \frac{100g}{\sqrt{CwCo}}$$

and will be seen to be the same as the previous transformer coupled stage. The original approximate expression is still useful, as it gives a truer indication of the signal noise ratio which, with a grid injected frequency changer and high-slope valves, is not greatly affected by the usual variations in Co.

A Variable-speed Turntable Unit

It is important to rotate the turntable of a gramophone at the correct speed for proper reproduction of a record, and difficulty is sometimes experienced in adjusting an electrically driven turntable to the desired speed. This is particularly the case when the driving unit is a synchronous motor, the speed of which is dependent only on the frequency of the supply.

The following is a description of a turntable unit which is inexpensive to manufacture and which can be adjusted readily to provide the correct playing speed for the turntable. With the latter object in view the motor is coupled to the turntable unit through a gear train which can be adjusted by turning a screw to vary continuously the velocity ratio, so that accurate setting of the speed can readily be accomplished. The gear train comprises a friction wheel and a cone or cone frustum which are relatively displaceable to vary the point at which the surface of the cone or frustum engages the friction wheel and thus vary the velocity ratio.

A friction disc secured to the turntable spindle and provided with a rubber rim may constitute the driven member of the train, or the turnable itself may be provided with a rubber rim and be used for this purpose. The end of the motor shaft can be made conical and arranged to bear against the driven member and thus constitute the driving member of the train. The velocity of the train can be varied by adjusting the height of the motor.

An arrangement of this kind is shown in section. In this A is the motor board which supports a frame B that carries the turntable spindle C and the driven friction disc D. The disc D is rubber tyred and is provided with angularly spaced pins which enter rubber grummets E secured to the turntable F and thus couple the disc to the turntable.

The motor G is pivotally mounted in a cradle H on a pair of oppositely disposed pins J, and is arranged so that its centre of gravity lies to the left of the pins J. The motor, therefore, tends to rotate anti-clockwise about the pins J.

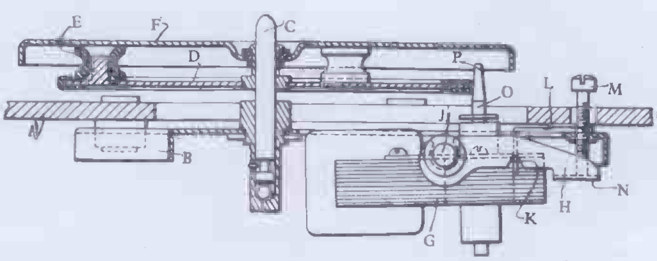
The cradle is pivotally supported upon a pair of knife edges K that are formed on brackets L and enter slots in the cradle. The brackets L are fixed to the frame B. The weight of the motor tends to rotate the cradle H anti-clockwise, and a vertical screw M engages the crossbar N of the cradle to act as a buffer and prevent rotation. The screw M is supported by the frame B and projects upwardly through the motor board A. It is adjustable to vary the angular position of the cradle.

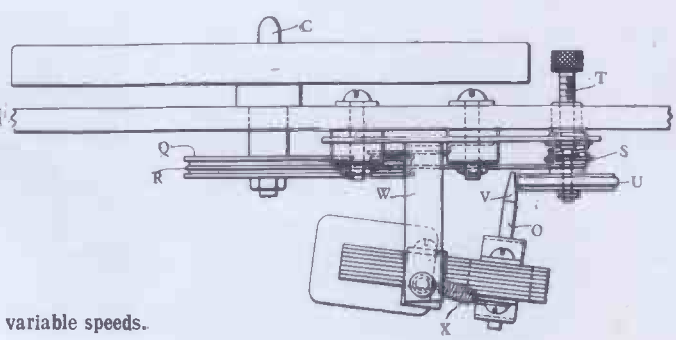
The rotor shaft O of the motor is provided with a conical tip P which bears against the rim of the driven friction disc under the action of gravity and constitutes the driving member.

To adjust the speed of the turntable it is only necessary to adjust the screw M. For example, if the screw is advanced in a downward direction the cradle H rotates clockwise about the knife edges K and raises the motor shaft O. The motor rotates anti-clockwise about the pins J under the action of gravity, and thus maintains the conical surface P in engagement with the friction disc D, but the point of engagement is lower down the conical surface, i.e., at a point of greater radius. Thus the velocity ratio is reduced and the speed of the turntable increased.

An alternative arrangement is shown in section in Fig. 2. In this case the turntable spindle C is coupled by pulley Q and belt R to a second pulley S rotatably mounted on an extension of the speed regulating screw T. The pulley S is secured to a rubber tyred friction wheel U and the conical tip V of the motor shaft drives the wheel U. The motor is pivotally mounted on a fixed bracket W and is biased in a clockwise direction by the spring X so that the end V of the shaft and the friction wheel U are maintained in engagement.

In this case the speed of the turntable can be varied by adjustment of the screw T to raise or lower the wheel U relative to the driving shaft.





Figs. 1 and 2. Two types of turntable drive providing variable speeds.

Design for

A Super-quality 12-watt Amplifier-II *(Conclusion)*

By J. C. G. GILBERT, A.M.I.R.E., A.M.I.W.T., F.R.S.A,

Senior Lecturer in Electro-communications, Northern Polytechnic, Holloway, London

In last month's issue the theoretical considerations of the output and penultimate stages of this amplifier were discussed. In this article, details of a three-valve electronic mixing stage and the following pentode tone corrector stage are given. By means of two five-position switches the frequency response curve can be tilted up or down at each end of the audible frequency spectrum.

Mains equipment compartment showing location of the various compartments. Note full-wave rectifier mounted on the side wall so as to secure good heat radiation.

THE design of audio frequency mixing systems is not simple because of the wide possible requirements. If the various outputs from microphones, radio or gramophone pickups are to be used separately and not super-imposed on each other, the circuit shown in Fig. 3a can be used. By means of the rotary switch any input can be selected and by means of the potentiometer networks approximately equal outputs can be obtained from each input. Alternatively a series of plugs terminating each output can be plugged into a jack in the input circuit of the first valve. In addition, the difficulty of adequate screening and the prevention of electrostatic and electromagnetic pickup, the system is clumsy from an engineering standpoint.

Fig. 3b shows a method of mixing by means of potentiometers, but suffers from the disadvantage that the upper one is above earth potential, and stray capacities due to potentiometer 1 tend to reduce the high-frequency response of input II. Fig. 3c shows a modification which allows each input channel to be at earth potential and the series resistor prevents short circuiting of the required input channel when the unwanted channel potentiometer slider is at the earthy end of the winding. The value of the series resistor Rs is limited by the maximum grid resistance allowable by the valve manufacturer and the Miller effect, and usually a value of 0.25 MΩ is satisfactory.

In broadcasting and recording studios a T pad or balanced H pad attenuator is used, but as the cost of these attenuators is considerable, they will not be considered here. A cheaper arrangement which gives comparable results is by the use of electronic mixing with separate valves for each input channel and a common anode load Ra. By this means no interaction takes place between the input channels and by adjustment of each input potentiometer fading in and out, and mixing can be accomplished. This simple circuit suffers from the disadvantage that each valve works into an anode load less than its own A.C. impedance, which seriously limits the output of the valve before harmonic distortion occurs. By amending the circuit to Fig. 3e to include the two resistors Rd and by suitable choice of their values, the load on the valves can be maintained at a value about twice the impedance of the valves. As medium impedance valves such as the Mullard 354v. or Mazda ACHL are suitable for this circuit, a value of 20.000Ω with an anode load Ra of 40.000Ω will produce the minimum of harmonic distortion. A common cathode resistor is used in Fig. 2, and provided that it is shunted by a condenser, the reactance of which is very small at the lowest audible frequency voltage distribution stability will be maintained.

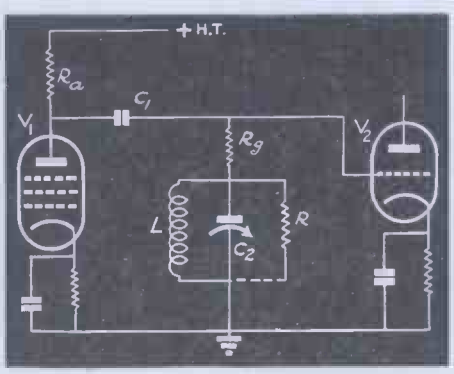


Fig. 1. Method of obtaining an increase in output voltage by means of a resonant circuit.

Tone-control Stage

The general method of increasing the output of an amplifier at any particular band of frequencies is to use a resonant circuit so that the voltage developed at the resonant frequency is greater than the response on either side of the resonance curve. Fig. 1 shows a typical type of resonant circuit in which the tuned circuit LC_2 is in series with the grid resistor Rg. At frequencies remote from the resonant frequency of L and C_2 the effect of the tuned circuit is negligible. At resonance, however, the dynamic resistance of the tuned circuit is additive to the grid resistor. If the grid resistor is smaller than Ra and the A.C. resistance of V_1 in parallel, then the stage gain at resonance will be materially increased, but if Rg is the same or greater than Ra and V_1 in

Super-quality 12-watt Amplifier—*Gilbert*

parallel, the effect of the increase of the grid-to-cathode load at resonance will be considerably less. By using a resistance-coupled pentode for V_1 its A.C. resistance can be made considerable compared with Ra and will have little effect on the effective value of Ra and V_1 in parallel.

If a triode is used for V_1, then Ra should have a value of 3 to 4 times that of the valve impedance and Rg should be small so that the tuned circuit will be effective at resonance. If Ra = V_1 then appreciable distortion will arise if the input voltage approaches the limit set by the cathode bias voltage, hence the use of a pentode type of valve. The resistor R in parallel with the tuned circuit can be used to control the limit of amplification at the resonant frequency. If the value of R is small, then the tuned circuit will be heavily damped, whilst if it is open circuited the maximum increase in stage gain is obtained.

The use of resonant circuits to produce a rise in response at a particular frequency band has several disadvantages. For example, if a very lightly damped tuned circuit is used in order to obtain a large voltage increase, then the possibility of oscillation occurring, arises. Also due to the decrement of the circuit, distortion of transient impulses is probable. The rise of voltage is restricted to a relatively small frequency band, whilst it is often desirable to produce an amplifier in which the increase or decrease is proportional to the frequency.

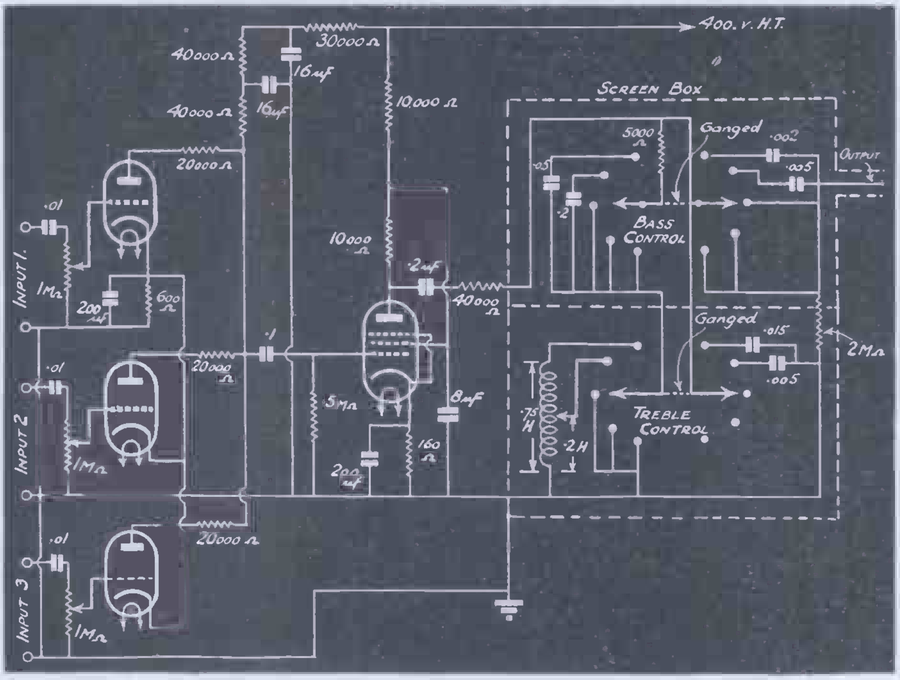


Fig. 2. Complete circuit of three-valve electronic mixing stage followed by a pentode tone-corrector valve.

Fig. 4a indicates a convenient method of increasing the bass response proportional to frequency. The response at medium frequencies is restricted by the ratio of the resistors Rp and R_2, these being necessary so that the increase of amplification at very low frequencies does not overload the valve and introduce amplitude distortion. Assuming that we desire a 20 db. increase at 50 c.p.s. compared with 500 c.p.s., then it is necessary to allow for a ten times rise in amplification at 50 c.p.s. compared with 500 c.p.s. If it is arranged that one one-tenth of the output voltage across the anode resistor is handed on to R_2 at medium frequencies then the valve will be able to amplify at 50 c.p.s. without distortion. In order to increase the amplitude at low frequencies it is necessary to include a condenser in series with R_2. As the reactance of the condenser increases with falling frequency, the impedance of the resistance R_2 and C_1 increases with falling frequency.

On the right of Fig. 4a is the equivalent electrical circuit of the valve circuit in which the resistor R_1 represents the valve A.C. resistance in parallel with the anode resistor and the series resistor Rp.

In order to calculate the gain at 50 c.p.s. compared with 250 c.p.s. let us take the values in the circuit diagram.

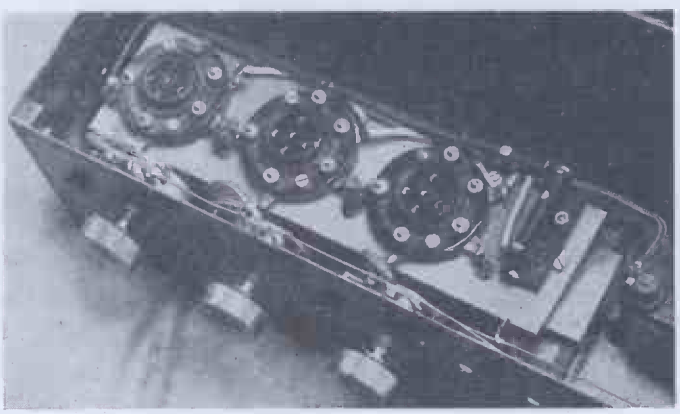
$$R_p = 40.000\Omega$$
$$R_2 = 5.000$$
$$C_1 = 0.2\mu F$$

$$\text{The reactance of } C_1 = \frac{10^6}{2\pi fC}$$

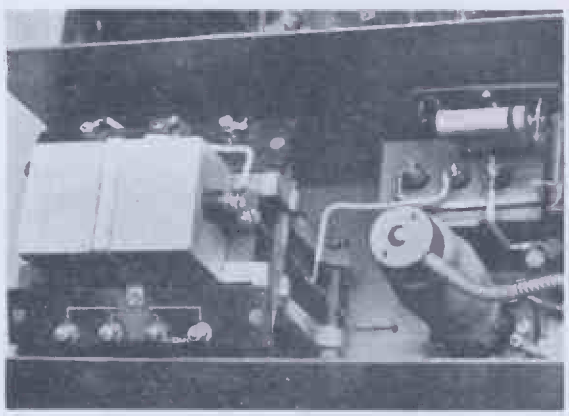
where $\pi = 3.142$.
f = frequency in cycles per sec.
C = capacity in microfarads.

At 50 c.p.s.

$$Xc = \frac{10^6}{2 \times 3.142 \times 50 \times .2}$$



Three channel input valves. Note method of suspension of the panel carrying the valveholders, very short leads and metallic screening to prevent electro-static pickup.



Tone control H.F. pentode valve together with centre-tapped output transformer.

Super-quality 12-watt Amplifier—*Gilbert*

$= 16{,}000\Omega$ (approx.)

At 250 c.p.s.

$$X_C = \frac{10^6}{2 \times 3.14 \times 250 \times .2}$$
$$= 3{,}200\Omega \text{ approx.}$$

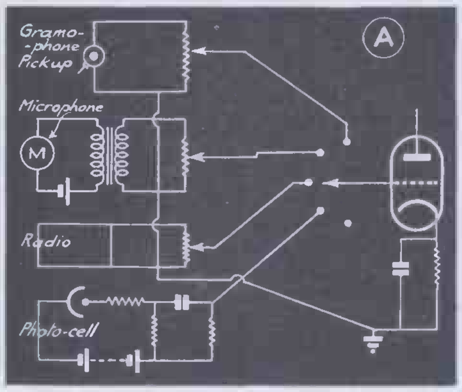


Fig. 3. (a) Switch method of selecting any required input.

The impedance of R_2 and C_1 in series is

$$Z = \sqrt{(R_2{}^2 + X_C{}^2)}$$

At 50 c.p.s.

$$Z = \sqrt{(5{,}000)^2 + (16{,}000)^2}$$
$$= \sqrt{(5 \times 10^3)^2 + (16 \times 10^3)^2}$$
$$= \sqrt{(25 \times 10^6) + (256 \times 10^6)}$$
$$= \sqrt{(281 \times 10^6)}$$
$$= 16{,}800 \text{ ohms approx.}$$

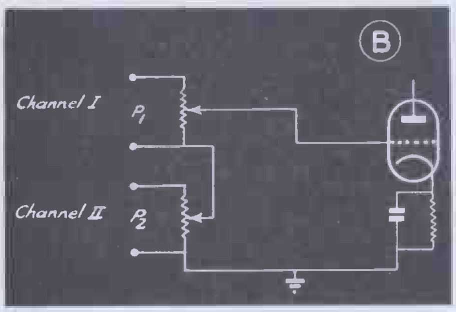


Fig. 3. (b) Twin potentiometer method. P1 is above earth potential.

At 250 c.p.s.

$$Z = \sqrt{(5{,}000)^2 + (3{,}200)^2}$$
$$= \sqrt{(5 \times 10^3)^2 + (3.2 \times 10^3)^2}$$
$$= \sqrt{(25 \times 10^6) + (10.04 \times 10^6)}$$
$$= \sqrt{35.04 \times 10^6}$$
$$= 5{,}900 \text{ ohms approx.}$$

As a percentage the voltage produced across $R_2\ C_1$ will be

$$\text{at 50 c.p.s.} = \frac{16{,}800}{16{,}800 + 40{,}000} \times 100\%$$
$$= 29.6\%$$
$$\text{at 250 c.p.s.} = \frac{5{,}900}{5{,}900 + 40{,}000} \times 100\%$$
$$= 12.9\%$$

This produces an increase in voltage at 50 c.p.s. compared with 250 c.p.s. of 2.3 times which is equivalent to 7.235 dbs on a voltage basis, and represents the intermediate bass boost, whilst the alternative condenser gives an increase of approximately 15 db.

Fig. 4b shows the method of obtaining the necessary treble boost. The inductance L_1 is in series with the resistor R_2, and as its reactance rises proportional to frequency, the impedance of the shunt arm rises with frequency. The stage gain at medium frequencies will be unaffected by the presence of either the series capacity or inductance and can be calculated from

$$\frac{e_2}{e_1} = gR_T \cdot \frac{R_2}{R_1 + R_2}$$

where e_2 = output voltage
e_1 = input voltage
g = mutual conductance in *amps.* per volt

$$R_T = \frac{Ra.Ro}{Ra + Ro}$$
$$R_1 = Rp + R_T$$

Fig 4b.

Using the suggested values in the circuit diagram it is possible to calculate the treble boost at 10.000 c.p.s. compared with the gain at 500 c.p.s.

Let $Rp = 40{,}000\Omega$
$R_2 = 5{,}000\Omega$
$L_1 = 0.75H$

Neglecting the D.C. resistance of L_1 then the reactance of the inductance.

$X_L = 2\pi fL$ where L = Henries

at 500 c.p.s.

$$= 2 \times 3.142 \times 500 \times .75$$
$$= 2{,}350 \text{ ohms approx.}$$

at 10.000 c.p.s.

$$= 2 \times 3.142 \times 10.000 \times .75$$
$$= 47{,}000 \text{ ohms approx.}$$

The impedance of R_2 and L_1 in series.

$$Z = \sqrt{(R^2 + X_L{}^2)}$$

Therefore at 500 c.p.s.

$$Z = \sqrt{(5{,}000^2) + (2{,}350)^2}$$
$$= \sqrt{(5 \times 10^3)^2 + (2.35 \times 10^3)^2}$$
$$= \sqrt{(25 \times 10^6) + (5.52 \times 10^6)}$$
$$= \sqrt{30.52 \times 10^6}$$
$$= 5{,}550 \text{ ohms approx.}$$

At 10.000 c.p.s.

$$Z = \sqrt{(5{,}000)^2 + (47{,}000)^2}$$
$$= \sqrt{(5 \times 10^3)^2 + (4.7 \times 10^4)^2}$$
$$= \sqrt{(25 \times 10^6) + (22.1 \times 10^8)}$$
$$= \sqrt{22.35 \times 10^8}$$
$$= 47{,}200 \text{ ohms approx.}$$

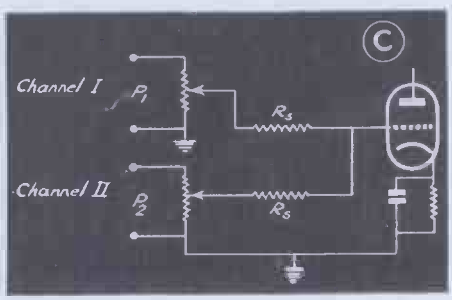


Fig. 3. (c) Twin potentiometer method with isolating resistors Rs.

As a percentage the voltage produced across R_2L_2 will be :—

At 500 c.p.s.

$$= \frac{5{,}550}{5{,}550 + 40{,}000} \times 100$$
$$= 12.2\%$$

At 10.000 c.p.s.

$$= \frac{47{,}200}{47{,}200 + 40{,}000} \times 100$$
$$= 54.2\%$$

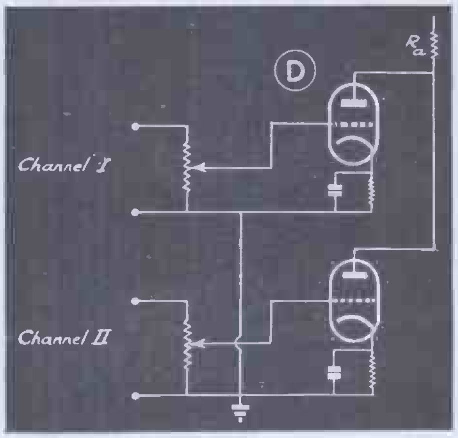


Fig. 3. (d) Separate valve input to each channel with common anode load.

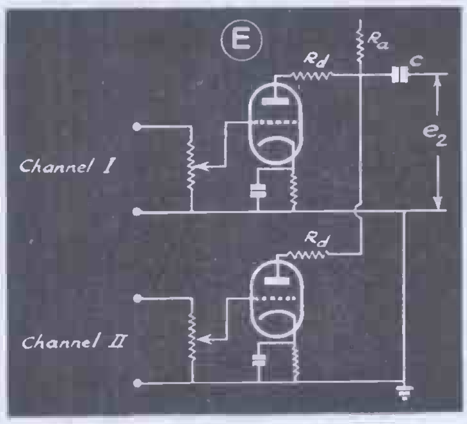


Fig. 3. (e) Separate valve input to each channel with isolating resistor Rd.

The voltage step-up is, therefore $\frac{54.2}{12.2}$ or 4.5 times which is equivalent to a rise of 13.06 dbs.

In order to decrease the output in the bass register the circuit shown in Fig. 4c is utilised. Under normal conditions the condenser C offers negligible reactance and the full signal voltage is produced across R_2 and R_p in series. If, however, a condenser C_1 of suitable value is inserted in series with R_p, the shunt impedance rises with decreasing frequency, and as the output voltage circuit does not include the condenser

Super-quality 12-watt Amplifier—*Gilbert*

portion of the potentiometer network the ratio of voltage e_2 falls with decreasing frequency.

In order to calculate the bass cut at 50 c.p.s. compared with 250 c.p.s. the

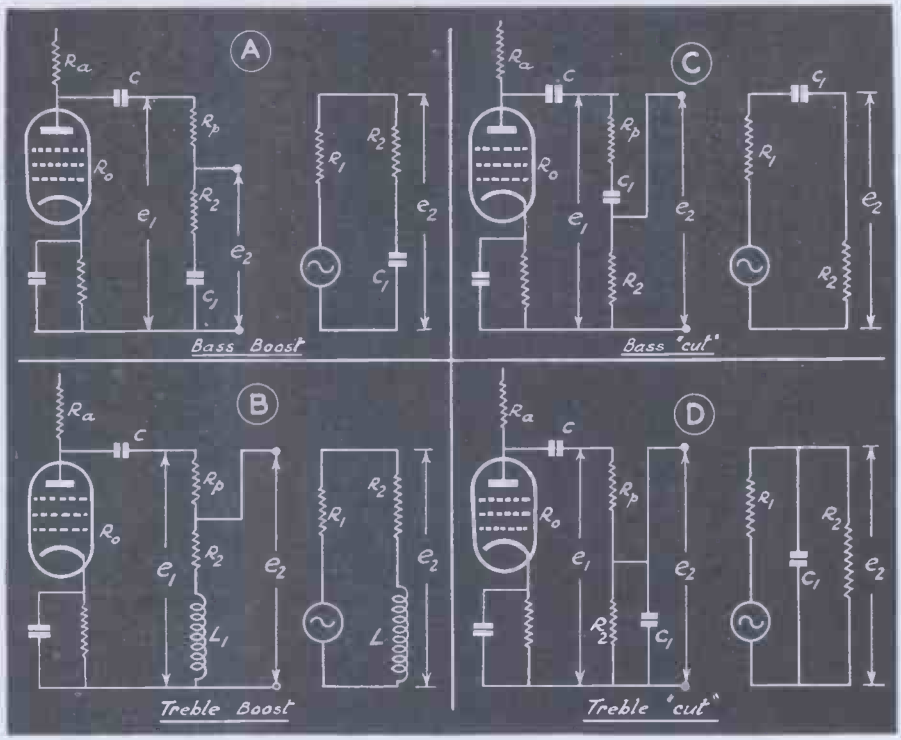


Fig. 4. Showing "breakdown" of complete tone-control stage. (A) Bass boost. (B) Bass cut. (C) Treble boost. (D) Treble cut. On the right of each circuit is shown the equivalent A.C. circuit.

values given in the circuit are :—

$R_p = 40{,}000\Omega$

$C_1 = .005\mu F$

$R_2 = 5{,}000\Omega$

The reactance of $C_1 = \dfrac{1}{\omega C} = \dfrac{10^6}{2\pi fC}$

where $C = \mu F$

Therefore at 50 c.p.s.

$$= \frac{10^6}{2 \times 3.142 \times 50 \times .005}$$

$= 637.000\Omega$

at 250 c.p.s.

$$= \frac{10^6}{2 \times 3.142 \times 250 \times .005}$$

$= 127.000\Omega$

The impedance of R_p and C_1 in series is

$$Z = \sqrt{R^2 + X^2}$$

At 50 c.p.s.

$$\begin{aligned} Z &= \sqrt{(40.000)^2 + (637.000)^2} \\ &= \sqrt{(4 \times 10^4)^2 + (6.37 \times 10^5)^2} \\ &= \sqrt{(16 \times 10^8) + (40.58 \times 10^{10})} \\ &= \sqrt{40.74 \times 10^{10}} \\ &= 639.000 \text{ ohms approx.} \end{aligned}$$

At 250 c.p.s.

$$\begin{aligned} Z &= \sqrt{(40.000)^2 + (127.000)^2} \\ &= \sqrt{(4 \times 10^4)^2 + (1.27 \times 10^5)^2} \\ &= \sqrt{(16 \times 10^8) + (1.61 \times 10^{10})} \\ &= \sqrt{1.77 \times 10^{10}} \\ &= 133.000 \text{ ohms approx.} \end{aligned}$$

As a percentage the ratio of the signal appeared across R_2 compared with that across R_p and C_1 in series :—

At 50 c.p.s.

$$= \frac{5{,}000}{5{,}000 + 639.000} \times 100$$

$= .78\%$

At 250 c.p.s.

$$= \frac{5{,}000}{5{,}000 + 133.000} \times 100$$

$= 3.62\%$

The ratio of bass cut at 50 c.p.s. is $\dfrac{3.62}{.78}$ which is a voltage reduction ratio of 4.65 times which is equivalent to a bass cut of 13db.

In a similar way by placing condenser C_1 in parallel with R_2 as shown in Fig. 4d the total impedance of the circuit decreases with rising frequency and smaller voltages are produced across e_2.

In order to calculate the amount of treble attenuation and using the values given in the circuit diagram.

$R_p = 40.000\Omega$

$R_2 = 5.000\Omega$

$C_1 = .005\mu F$

Comparing the attenuation at 5,000 c.p.s. with the output at 500 c.p.s. :—

Reactance of $C_1 = \dfrac{1}{\omega C} = \dfrac{10^6}{2\pi fC}$ where

$C = \mu F$

At 500 c.p.s.

$$Xc = \frac{10^6}{2 \times 3.142 \times 500 \times .005}$$

$= 63{,}700$ ohms

At 5,000 c.p.s.

$$Xc = \frac{10^6}{2 \times 3.142 \times 5000 \times .005}$$

$= 6{,}370$ ohms

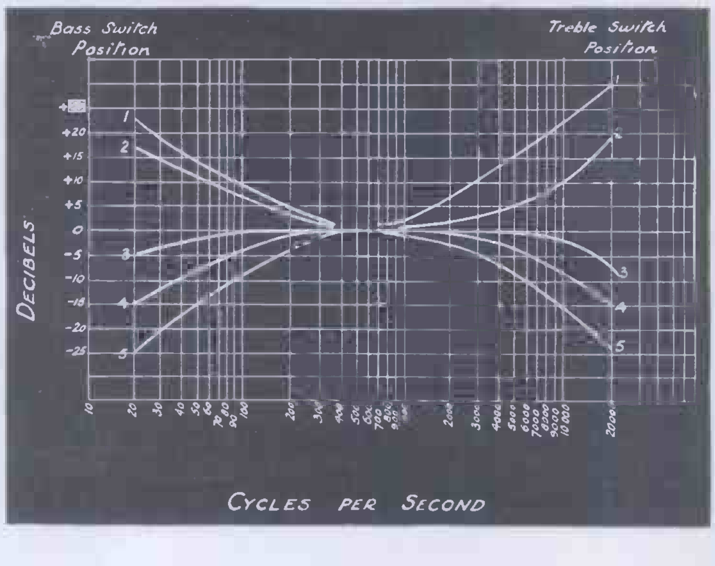
The impedance of C_1 and R_2 in parallel.

$$Z = \frac{RXc}{\sqrt{R^2 + Xc^2}}$$

At 500 c.p.s.

$$Z = \frac{50{,}000 \times 63{,}000}{\sqrt{(5{,}000)^2 + 6.37 \times 10^4)^2}}$$

Fig. 5. This group of curves indicates that by the various combinations of the treble and bass switches, twenty-five different response curves can be obtained.



Super-quality 12-watt Amplifier—*Gilbert*

$$= \frac{(5 \times 10^3) \times (6.37 \times 10^4)^2}{\sqrt{(5 \times 10^3)^2 + (6.37 \times 10^4)^2}}$$

$$= \frac{31.85 \times 10^7}{(25 \times 10^6 3) + (40.58 \times 10^8)}$$

$$= \frac{31.85 \times 10^7}{6.39 \times 10^4}$$

$= 4980\Omega$ approx.

At 5000 c.p.s.

$$Z = \frac{5{,}000 \times 6370}{\sqrt{5{,}000^2 + 6370^2}}$$

$$= \frac{(5 \times 10^3) \times (6.37 \times 10^4)}{\sqrt{(5 \times 10^3)^2 + (6.37 \times 10^3)^2}}$$

$$= \frac{31.85 \times 10^7}{\sqrt{(25 \times 10^6) + (40.58 \times 10^6)}}$$

$$= \frac{31.85 \times 10^7}{81.0 \times 10^3}$$

$= 3{,}940$ ohms approx.

Therefore the ratio of attenuation at 5,000 c.p.s. compared with 500 c.p.s.

$$= \frac{4{,}980}{3{,}940} = 1.25 = 2.25 \text{ db.}$$

By amalgamating these four individual characteristics by switches as was shown by W. T. Cocking in the *Wireless World*, June 8, 1939, an amplifier can be produced in which either rising or falling characteristics at each end of the frequency range can be obtained. The complete circuit is shown in Fig. 2 in conjunction with the 3-way mixing unit. A Mazda TSP4 high-frequency pentode in conjunction with an anode load of 10.000Ω gives a stage gain of approximately 50 times with a total harmonic distortion of $1\frac{1}{2}$ per cent. The fixed potentiometer R_p R_2 gives a voltage reduction of nine times at medium frequencies and Fig. 5 shows the tone control response curves at all settings of the two rotary switches. By correct settings any one of twenty-five separate response curves can be obtained which should fill every requirement for various inputs and acoustic conditions.

Construction of Complete Equipment

From the various photographs shown in this and last month's issue of this journal, it will be seen that the complete amplifier has been built in a sectionalised manner. In the February issue, p. 72, will be seen the method of assembly and as it was designed for hard usage it had to be ruggedly constructed.

The complete cabinet which measures 20 in. by 12 in. by 7 in. is constructed from $\frac{1}{8}$ in. hard brass with brass angle $\frac{1}{2}$ in. by $\frac{1}{2}$ in. by $\frac{1}{8}$ in. as the means of supporting the side panels. A lid of similar design completely encloses the equipment for transport purposes, but it should not be operated with the lid in position because of the temperature rise from the total anode and filament dissipation of the valves.

In the left-hand section is housed the three input valves together with the input potentiometers. In order to prevent any possibility of microphony, the three valve holders are rigidly mounted on a brass panel which is held between sorbo rubber supports. Flexible wiring must be used to this panel so that complete freedom of movement is possible. On the front panel is the potentiometer controlling the input to the tone control stage which is located in the second section. All the condensers, resistors, switches and tapped inductances are contained within a screened box in order to reduce the possibility of electro-magnetic pickup. Behind the TSP4 R.F. pentode is the output transformer which is orientated in order to prevent pickup from any alternating magnetic field.

The third section contains the two penultimate valves and the output valves. The 50.000Ω paraphasing potentiometer is mounted on the shallow baseboard, below which is the whole of the filament wiring and H.T. circuits.

On each side of the R.H. PX25 valve will be seen the biasing resistors of the output stages. In the fourth section is the mains equipment, with the rectifier valve mounted above the mains transformer in order to provide free heat radiation.

Although the positioning of the component parts is not critical, a few notes will enable the constructor to obtain the maximum signal to background noise ratio. In high stage gain amplifiers the magnetic field due to the mains transformer and smoothing chokes, particularly the constant inductance type, are prone to induce a 50 c.p.s. signal into the output transformer and any other inductance. Obviously signals of this nature will be transmitted to the loudspeaker in the form of a 50 cycle hum, and no quantity of smoothing will eliminate it. The only method of preventing magnetic pickup is to place the offending component in such a position that the radiating magnetic fields do not induce a current into the output and other transformers.

In order to determine these exacting positions, the position of the mains transformer should be decided and wired into circuit with a resistance load connected across the filament windings so that a considerable magnetising current flows. The 25H smoothing choke should then be positioned by connecting a pair of high-resistance telephones across the choke and the position found where no 50 c.p.s. hum is heard. Usually this is not at right angles to the core of the mains transformer and it will be noted from the photograph that the core of the 25H choke is at an angle less than 90° in respect of the mains transformer. The complete wiring of the mains equipment can be completed and a dummy resistance load dissipating the total anode wattage connected across the H.T. supply.

The output transformer can be positioned next, and again a pair of phones connected across the primary winding will enable the position of zero pickup to be found. As will be seen from the photographs of the layout of the various components, the output transformer could not be placed nearer than 12 in. from the mains transformer, and therefore it was placed in the second section. Care in placing the components will be repaid by the hum-free background. Screened wiring should be used for all grid wiring, earthing the screening to the chassis.

Balancing the Paraphase Stages

There are several methods of balancing the valves of the penultimate and output stages. It is desirable that the output valves should be matched so that equal anode currents flow with given anode and grid volts in order that the magnetic fields in the split primary of the output transformer cancel one another. One method is to place an output transformer in the lead between the output transformer centre tap and the milliameter or H.T. line. Across the secondary of the output transformer (a small loudspeaker type can be used) is connected a pair of phones. A medium frequency signal of about 200 to 2,000 c.p.s. is applied to one of the input channels, and the input level kept as low as possible. A signal should be heard in the phones, but on rotation of the paraphasing potentiometer, a position will be found where zero or practically zero signal will be heard. This is the correct position and confirms that the alternating current of each output valve is of the same amplitude and 180° out of phase with its counterpart.

An alternative method is to place two equal resistors of about 50.000Ω in series from anode to anode of the output stage and from the junction of the resistors a connection is taken to a $1\mu F$ 600v. working condenser. A pair of phones are connected to the other side of the condenser and to earth. The balancing is made in the same way as described above.

Probably a 50 c.p.s. signal will be heard in the phones and this is due to an out-of-balance component in the filament circuit of the output stages. By rotating the filament balancing potentiometers to their mid-point, approximate balancing is obtained and the careful rotation of either will give a position of complete cancellation.

Throughout the whole amplifier adequate precaution has been taken to reduce the 100 c.p.s. A.C. ripple superposed on the H.T. line voltage by the generous use of high-value electrolytic condensers. It is very advisable to wire the filament circuits with twisted insulated wire so that the electro-magnetic field is cancelled. Particular care should be taken to keep all grid wiring and resistors as far as possible from the filament circuits.

Since the first part of this article was published experiments have been conducted in connection with the method of producing the necessary paraphase signal. It will be seen from the circuit in the first article, that the signal is fed back to the Mullard 354v. from the grid potentiometer of the Osram PX25. Due to the Miller effect of the PX25 an attenuation of the extreme high frequencies takes place and therefore a restricted frequency range is fed to the paraphase valve and its output valve. In order to overcome this small degree of attenuation, an alternative circuit is being tested and if the results merit any amendment in the original circuit, full details will be given in the next issue.

In part I, mention was made of the "Coil" and Voigt moving coil pickups. The opportunity of taking measurements on these instruments has been afforded and these curves together with details of a wide-range acoustic labyrinth loudspeaker system will be given.

COMPONENTS FOR 12-WATT SUPER-QUALITY AMPLIFIER

Since the amplifier was constructed certain components have become unobtainable due to war conditions. Other components of similar characteristics have been substituted and are available at the time of issue of this journal.

POWER UNIT AND AMPLIFIER. (February issue.)

CONDENSERS
2—50 μF. electrolytic 12 v. working (T.C.C.).
2—50 μF. electrolytic 50 v. working (T.C.C.).
2—8 μF. electrolytic 500 v. working (T.C.C.).
2—16 μF. electrolytic 550 v. working (T.C.C.).
1—4 μF. paper 600 v. working (T.C.C.).
2—.1 μF. tubular 1,500 v. working (T.C.C.).

CHOKES, SMOOTHING
1—25 H. at 150 mA. (Partridge).
1—15 H. at 150 mA. (Partridge).

RESISTANCES
2—power variable resistors 0-1,500 ohms. (Varley).
2—30-ohm potentiometers. Kabe. (Lechner).
1—1 megohm ½ watt (Erie).
2—1,000 ohm 1-watt (Erie).
2—50,000 ohm 1-watt (Erie).
2—10,000 ohm 1-watt (Erie).
2—5,000 ohm ½-watt (Erie).
2—50 ohm 1-watt (Erie).
1—200,000 ohm ½-watt (Erie).
1—250,000 ohm ½-watt (Erie).
1—potentiometer 50,000 ohms (Erie).

SUNDRIES
5—5-pin valve holders, baseboard mounting.
1—0-150 mA. milliammeter (Ferranti).
2—closed circuit jacks J6 (Bulgin).
1—plug P15 (Bulgin).
1—twin fuse holder (Bulgin).

TRANSFORMER, MAINS
Primary: 0-200/220/240 v.
Secondaries:
500-0-500 v. at 150 mA.
2-0-2 v. at 3 amps.
4 v. at 2 amps.
4 v. at 2 amps.
2-0-2 v. at 8 amps. (Partridge).

TRANSFORMER, OUTPUT
Primary: 2 x PX24 in P.P.
Secondaries:
(a) 15 ohms.
(b) 35 ohms. (Partridge).

VALVES
2—354v triodes (Mullard).
2—PX25 triodes (Osram).
1—MU14 full-wave rectifier (Osram).

3-WAY MIXER AND TONE CONTROL STAGE. (March issue.)

CONDENSERS
3—.01 μF. 1,500 v. working (T.C.C.).
2—200 μF. 12 v. working electrolytic (T.C.C.).
1—.1 μF. 1,500 v. working (T.C.C.).
3—8 μF. 500 v. working electrolytic (T.C.C).
2—.2 μF. 1,500 v. working (T.C.C.).
1—.05 μF. 500 v. working (T.C.C.).
1—.002 μF. 200 v. working, mica (T.C.C.).
2—.005 μF. 200 v. working (T.C.C.).
1—.015 μF. 200 v. working (T.C.C.).

RESISTORS
3—potentiometers 1 megohm (Erie).
1—potentiometer 0.5 megohm (Erie).
1—600 ohm 1-watt (Erie).
1—350 }
1—300 } =150 ohm. 1-watt (Erie).
3—20,000 ohm ½-watt (Erie).
2—40,000 ohm 3-watt (Erie).
1—40,000 ohm 1-watt (Erie).
1—30,000 ohm 3-watt (Erie).
2—10,000 ohm 2-watt (Erie).
1—5,000 ohm 1-watt (Erie).
1—2 megohm ½-watt (Erie).

SUNDRIES
1—tone correction choke .75 H. tapped at .25 H. (Postlethwaite).
3—5-pin baseboard valveholders.
1—7-pin baseboard valveholder.
3—screened input plugs and sockets, P 158 (Bulgin).
2—5-pole two-way rotary switches (Bulgin).

VALVES
3—MH4 (Osram).
1—TSP4 (Mazda).

Colour Television

THE first public demonstration in the U.S.A. of *direct* pick up of colour television was recently given by Dr. Peter C. Goldmark, chief television engineer of the Columbia Broadcasting System. The colour pictures were picked up by a television camera in the C.B.S. television laboratories, transmitted by coaxial cable and reproduced on different models of colour receivers in a studio in the new C.B.S. building.

Three receivers were used in the demonstration, a standard black-and-white receiver, adjusted to receive colour pictures in black and white, a standard black-and-white receiver that has been adapted for colour, and a compact table model colour receiver specially designed and constructed in the C.B.S. laboratories.

An Orthicon tube was used in the demonstration, and it is stated that special Orthicon tubes are now being developed which are expected to require even less light than was necessary on this occasion.

The small, specially designed and constructed receiver incorporates two important features not previously demonstrated to the public. The first of these is a method of synchronising the colour disc in the receiver with the colour disc in the studio by the synchronising impulses ordinarily transmitted. This means it is no longer necessary to rely upon the 60-cycle mains for synchronising the discs, and permits reception of colour pictures when a receiver and transmitter operate from different power supplies.

This receiver also has a simple, ingenious method of phasing the colour disc so that the colours shown at the receiver can be "locked" to the colours being picked up. The viewer has only to push a button until the picture appears in its proper colours, then release the button. The colours remain properly synchronised with those of the original scenes.

One impressive part of the demonstration emphasised dramatic effects possible with colour television. Merely by turning a dial, Dr. Goldmark was able to create the effect of moonlight on the face of a girl.

During the demonstration, a new system of low intensity fluorescent lighting was used. These lights eliminate most of the glare from the eyes of the person in front of the camera and are "cold light" so that no discomfort is experienced. They have been developed in the Columbia Broadcasting System television laboratories.

News Brevities—

Commercial and Technical

IT was recently announced that Dr. A. P. M. Fleming has been awarded the 1940 Faraday Medal—the highest award of the Institute of Electrical Engineers. Dr. Fleming is a director of Metropolitan Vickers Electrical Co., Ltd.

* * *

Seven radio monitoring stations of the Federal Communications Commission (America) are being used to track down the illegal use of radio transmitters. It is claimed that the monitor service can actually trace a transmitter which is being operated from a fast-moving car.

* * *

At the twentieth anniversary of broadcasting in the United States, the guest of honour was Dr. Frank Conrad, a Westinghouse engineer, who is credited with initiating the first successful broadcasting in America.

* * *

Oliver Pell Control Limited announce with regret the recent death of Charles Oliver, M.I.E.E., founder of the firm bearing his name. Mr. Oliver was one of the pioneers of the electrical industry. As long ago as 1898 he started to make arc lamps in a small factory, and was later joined by W. M. D. Pell.

* * *

At the annual general meeting of the Radio Manufacturers' Association, Mr. M. M. Macqueen of the G.E.C. took the chair. Mr. Macqueen revealed the considerable amount of work which the executive council and the various committees of the Association carried out during 1940; the year was undoubtedly the busiest in the history of the Association. The Government were relying on trade organisations for help in various directions.

Referring to the technical work of the Association, Mr. Macqueen paid tribute to the efforts of E. M. Lee, chairman of the Technical Committee, who has been very active, and whose committee is an organisation of immense value.

* * *

Westinghouse (America) engineers believe that it may be possible to light homes with power picked up from the air, in the same way that we now receive broadcast programmes, by the use of a new tube recently developed in their laboratories. This tube, the Klystron, is capable of transmitting power through space. The energy from such a tube might be focused like a light beam with the aid of a copper-lined horn. Other possible uses suggested by Westinghouse engineers for the Klystron include the supply of ultra shortwave heat for medical treatment, increasing the number of television channels, and utilising it to serve as a radio beacon for guiding aircraft.

* * *

According to the latest reports from the Federal Communications Commission, fifteen radio stations have been authorised to engage in frequency modulation broadcasting on a commercial basis as soon as they are equipped to do so. Thirty-five other applications are being considered.

* * *

A gala television broadcast was staged by the National Broadcasting Company when President Roosevelt and Mr. Wendell Willkie delivered their final addresses prior to the election of the new President. On election night, a comprehensive coverage of election returns was transmitted, commencing at 6.45 and continuing until 1 a.m. N.B.C.'s largest studio was set up, not only to put the election returns on the air as quickly as possible, but also to serve as a reception room for some 2,500 members of the radio industry and other guests.

* * *

The Federal Communications Commission recently conducted a hearing on medical short-wave radio apparatus used to generate artificial fevers and to produce local heat. It seems probable that certain frequency bands will be specifically allotted for the use of such apparatus, as such equipment often causes interference with radio receivers if it is permitted to operate on an adjacent channel.

* * *

The schedule of broadcasts of the B.B.C.'s Overseas Service has been completely revised and the services have been increased from four to six, operating for a total of $54\frac{1}{4}$ hours daily—an increase in the transmitting time of nearly twenty-five per cent. This new grouping will give a World service in English for 21 hours daily; a second World service in other languages of the Empire—such as Hindustani, Afrikaans, Burmese, Maltese—for two hours daily; a main European service in German, French, Italian, Dutch, Flemish and the languages of Central Europe for twenty hours daily; a second European service of five hours daily for broadcasts to Spain, Portugal, the Scandinavian countries and the Balkans; a Latin-American service; four hours daily, for broadcasts in Spanish and Portuguese; and the Near East service, $2\frac{1}{4}$ hours daily, for broadcasts in Arabic, Persian and Turkish.

Overseas English New bulletins may be heard at the following times (B.S.T.) on the wavelengths given:—

7.15 a.m. (W): 31.55, 31.32, 31.25, 25.53, 19.66 m.
9.0 a.m. (W): 31.55, 25.53, 19.82, 19.66, 19.60, 16.84 m.
9.0 a.m. (E): 49.59, 30.96 m.
12.0 noon (W): 31.25, 25.53, 19.82, 16.86, 13.97, 13.92 m.
2.0 p.m. (W): 25.53, 19.82, 16.86, 16.84, 13.97, 13.92 m.
2.15 p.m. (E): 49.59, 25.38, 25.29 m.
5.0 p.m. (W): 31.75, 31.55, 25.53, 16.84, 13.93, 13.92 m.
5.15 p.m. (E): 49.59, 25.38, 25.29 m.
7.0 p.m. (W): 41.96, 31.25, 25.53, 19.82, 19.66 m.
9.45 p.m. (W.): 31.25, 25.53, 25.38, 25.29, 19.82 m.
11.0 p.m. (E): 373.1, 285.7, 261.1, 49.59, 30.96 m.
1.0 a.m., 11.45 p.m., 2.45 a.m., 5.30 a.m. (W): 49.10, 31.32, 31.25, 25.53 m.

(E) European Service (W) World Service.

* * *

A signal generator, designed primarily for laboratory research involving the study and development of new radio equipment to operate on 3,000 to 4,000 Mc., has been developed in America. This new instrument generates test signals of adjustable intensity and frequency, and operates at unusually high frequencies.

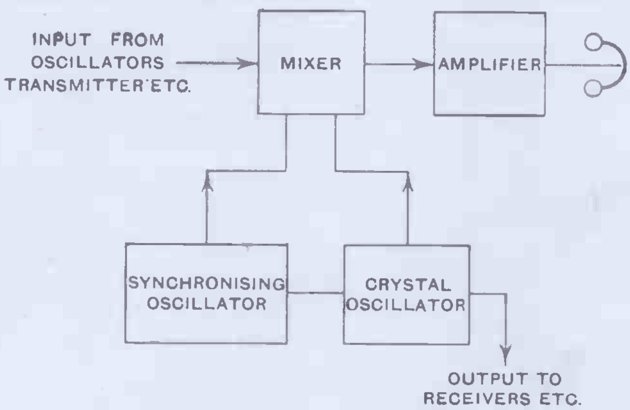
The oscillator is an end-plate magnetron tube, having a maximum output of about 1 watt in the frequency region of 3,000 Mc., and feeds an aerial from which the generated signal is radiated. Control over the signal strength is obtained by means of an attenuator, and a thermocouple is used to indicate the output.

A New Crvstal Calibrator

For R.F. Measurements on Oscillators, Receivers and Transmitters

Photograph of the new G.E.C. crystal calibrator.

AN addition to a range of high-frequency measuring instruments produced by the General Electric Co., Ltd., is a new crystal calibrator for radio-frequency measurements on oscillators, receivers, transmitters and similar apparatus. As the schematic diagram shows, the instrument comprises essentially a crystal oscillator followed by a mixing stage and listening post. In addition, a synchronizing oscillator is incorporated which enables very



Schematic diagram showing the operating principle of the crystal calibrator.

high frequencies to be determined.

The crystal oscillator contains a circuit whereby crystals operating at a frequency of 100 kc/s 1 mc/s or 5 mc/s may be selected by means of a convenient panel switch. When this switch, which can be seen in the photograph, is operated the associated circuit components are also changed over. The crystals used are made by the G.E.C. at their Crystal Works from the finest quality quartz, and the special apparatus used ensures the optimum cut for each particular frequency, thus ensuring an extremely low temperature coefficient, precise accuracy and freedom from oscillation in spurious modes.

The heterodyne method of detection is employed when the apparatus requiring calibration produces an output voltage. The unknown signal is fed to the mixer valve grid, while the oscillation voltage is fed into an independent circuit which is arranged to accentuate the harmonics of the oscillator. The two frequencies mix, and as the unknown source is varied, an audio frequency beat note will be heard which can be reduced to zero when the precise frequency is obtained. The beat note is amplified by a detector following conventional lines, the output of which is fed into high-resistance headphones.

When operated at very high frequencies the higher harmonics of the 5-Mc oscillator is used and confusion may arise in differentiating, for example, the 30th or 31st harmonic. In order to avoid this difficulty, an additional synchronizing oscillator is employed, using an acorn triode. This oscillator is capable of being switched to either 25 or 50 Mc, which can be brought into tune with the corresponding harmonic of the 5-Mc crystal oscillator by means of a trimmer adjustment on the panel of the instrument. Thus a "milestone" is provided making it impossible for errors to be made.

In order that receivers, wavemeters and similar apparatus can be calibrated, a socket is provided whereby an output voltage from the crystal oscillator is made available. By taking a feed from this point, apparatus can be calibrated, using the receiver output meter or the wavemeter indicator as a detecting device. All the controls are conveniently placed on a grey lacquered panel which is engraved to show the functions of each control, as can be seen in the photograph.

The instrument is mains-operated, rectifiers and smoothing equipment being incorporated to enable it to be operated from alternating-current mains, 230 volts 50 cycles.

Novel Suppressor Grid-cathode Connection for Receiving Valves

IN certain valves a grid or an auxiliary screen electrode is provided to which there is no direct external connection. One particular case is the pentode valve in which the suppressor grid is electrically connected to the filament by a conducting strip. It sometimes happens that this connection is not properly made, or is effected to the wrong electrode. Tests suitable for line production seldom reveal these defective valves.

The following arrangement of connections which has been developed in the laboratories of the Radio Corporation will ensure that faulty connections of this kind in filamentary valves will be detected during manufacture when the filament is energised: the legs of the filament are joined to lugs sealed in the pinch and only connection from one lug is taken out to a pin in the base. The suppressor grid is provided with a lead which extends through the pinch to a pin in the base and functions as the other filament lead, a connection being made inside the valve from the suppressor lead to the other otherwise free filament leg. Thus the filament circuit includes the connection between the filament and the suppressor, and if the connection has not been made, no filament current will flow and the fault will be revealed at once.

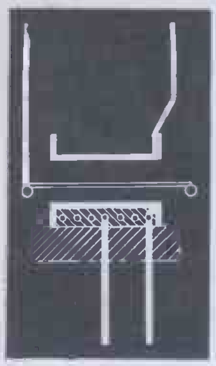


Fig. 1. Cross-section showing the structure of the small triode.

Valves for Extremely High Frequencies

Considerations of Design

By B. J. THOMPSON and G. M. ROSE, Jr.
Research and Engineering Department, RCA Manufacturing Company

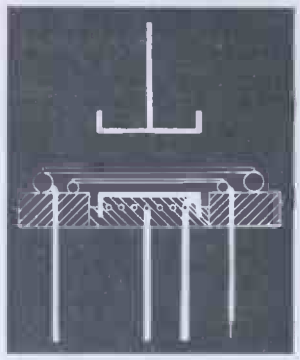


Fig. 2. Cross-section showing the structure of the small screen-grid valve.

This article, which is an abstract from a paper published by the Institute of Radio Engineers, outlines the fundamental principles in the design and construction of valves for reception down to 60 centimetres with conventional circuits. The valves referred to provided the basis of design of the 953 and 956 Acorn types. These valves represent nearly a tenfold reduction in dimensions as compared with conventional receiving valves, but compare favourably with them in transconductance and amplification factor. The inter-electrode capacities are only a fraction of those obtained in the larger ones.

THE limitations imposed on tuned radio-frequency reception at the lower wavelength limit are due to a number of factors. These are:

(1) The inter-electrode capacities of the valve are so great that, with the addition of the tuning capacity, the L/C ratio is too low for a value of impedance sufficient to afford appreciable amplification.

(2) The lead inductances of the valve are so great that much of the output voltage appears inside the bulb, where it is unavailable.

(3) The interelectrode capacities and lead inductances form a tuned circuit at a wavelength well above the limit desired.

(4) The time of transit of the electrons across the space between electrodes becomes an appreciable part of a period which results in a reduction in the effective amplification of the valve.

(5) As the wavelength is reduced the radio-frequency resistance of the circuit is increased with a consequent reduction in resonant impedance.

It will be seen that most of these limitations are associated with too large a ratio of some fixed characteristic of the valve to those characteristics of the circuit which vary with frequency. These characteristics of the valve are fixed only for a given design; however, any change in design which results in lower transconductance, as would be the case with increased interelectrode spacing to reduce capacity, cannot be considered a genuine improvement.

In a valve, if all linear physical dimensions are kept in a fixed ratio to each other there will be no change in transconductance, anode current, or amplification factor at fixed operating voltages, no matter what changes are made in the actual magnitude of the linear dimensions. On the other hand, the values of interelectrode capacity, lead inductance, and time of electron transit are in direct proportion to the magnitude of the linear dimensions.

For optimum design at any wavelength, all valve and circuit linear dimensions should be in proportion to the wavelength. This principle is modified in practice since, at longer wavelengths, there is no advantage in making the valve of large size and it becomes inconvenient to make the tuned circuit of optimum dimensions.

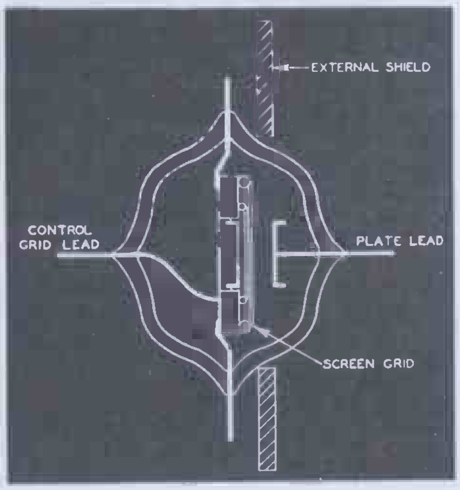


Fig. 3. Screening arrangement used with the small screen-grid valve.

Unfortunately, it is not to be expected that this proportionality of dimension will result in constant amplification, since the resonant impedance, L/RC, is reduced as the wavelength becomes shorter. However, on the basis of Butterworth's formulæ for high-frequency resistance, and Coffin's formula for inductance, with Rosa's correction, a coil having the following dimensions:

length = $\frac{1}{8}$ in.
diameter = $\frac{1}{8}$ in.
turns = 5
wire diameter = 0.020 in.

should have a resistance of 0.37 ohm and an inductance of 5.87×10^{-8} henry at 50 centimetres wavelength. A capacity of 1.2×10^{-12} farad would be required for resonance, giving an impedance, L/RC of 132,000 ohms. This high value in comparison to those obtained at longer wavelengths may be accounted for in part by the fact that the coil is of much more nearly optimum design than those used at the longer wavelengths.

Since valves of conventional size have been found to have a lower wavelength limit of about five metres, the principle of proportionality requires a tenfold reduction of linear dimension to produce a valve capable of amplification at a wavelength of 50 centimetres.

Valve Structure

The largest dimension of either of these small valves is less than three quarters of an inch, and the elements themselves are correspondingly small. Both types of valve are of parallel plane construction and have indirectly-heated cathodes.

In the triode the parts are sufficiently light in weight to permit supporting them on their lead wires alone. This has resulted in the elimination of capacities which would otherwise be present between the various elements and the support structure. Both anode and cathode are of the same shape, consisting of two small metal cups placed back to back with the grid interposed between them. The cathode cup has within it a small heater; its outer surface is coated with the emitting material. The grid is of mesh fastened on a support ring. The inter-electrode spacings are only a few thousandths of an inch. Fig. 1 shows the general construction.

The assembly scheme of the triode cannot be satisfactorily applied to the screen-grid valve because of mechanical complications arising from the presence of the second grid. A different method is used which is productive of a stronger and more rigid assembly. The tetrode parts, however, are of the same size and shape as those of the triode, with the addition of the screen grid which is similar to the control grid, though somewhat larger.

A small ceramic disc serves as a foundation upon which the valve parts,

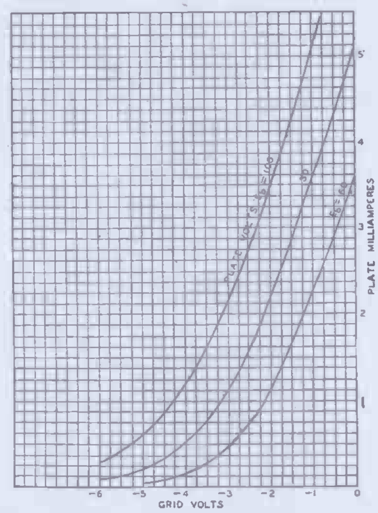


Fig. 4. Mutual characteristics of the small triode.

with the exception of the anode, are assembled. It, therefore, acts as a common supporting insulator. The correct spacings between the grids and the cathode are obtained by adjusting their individual distances from this insulator. As the distance from the anode to the screen grid is not so critical, the anode is supported by its lead wire from the glass bulb. The spacings between the other parts are again only a few thousandths of an inch. The general construction is shown in Fig. 2.

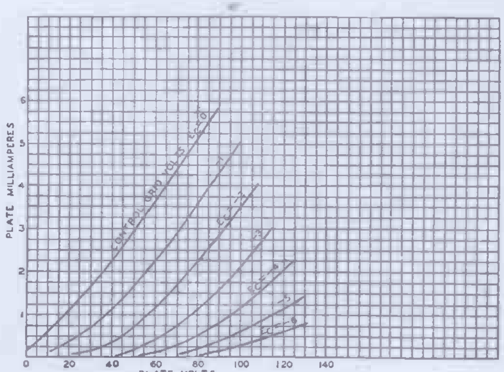


Fig. 5. Anode characteristics of the small triode.

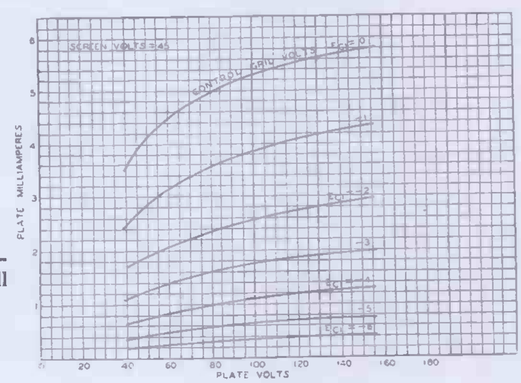


Fig. 6. Anode characteristics of the small screen-grid valve.

The bulbs used to enclose both types of valve are in two parts which are more or less hemispherical in shape. These two parts are placed together with the mount inside, and a seal is made between them. All of the triode leads pass through this seal, thereby eliminating the need for a stem as ordinarily used. In the tetrode separate seals are made at opposite ends of the bulb for the anode and control grid leads, while all of the remaining leads come out through the main seal.

In the case of the tetrode this general arrangement has advantages from the standpoint of screening. The mount is so placed in the bulb that its screen grid lies just above the plane of the seal and extends almost to the glass. It is readily seen from Fig. 3 that when the external shield is placed as indicated the anode is effectively isolated from the control grid. The screen grid lead is quite short and comes out adjacent to the external shield where it can be readily earthed, thus minimising screen-lead impedance. This holds true for the heater and cathode leads likewise.

Electrical Characteristics

From an examination of the static characteristics of the triode which are shown in Figs. 4 and 5 it is readily seen that these characteristics are directly comparable both as to magnitude and shape with those of an ordinary triode. Under the operating conditions, anode voltage = 67.5 volts and grid voltage = −2 volts, the values of the various parameters are as follows :—

Anode current = 4 mA.
Anode resistance = 9,500 ohms.
Transconductance = 1,150 ua/v.
Amplification factor = 14.7.

The inter-electrode capacitances for these values have been measured as follows :—

Grid-cathode capacitance = 0.7 $\mu\mu$f.
Anode-cathode capacitance = 0.07 $\mu\mu$f.
Anode-grid capacitance = 0.8 $\mu\mu$f.

As might be predicted from the results of the measurements on the triode, the tetrode characteristics are likewise similar to those of the larger valves of this sort. A family of anode current-anode voltage curves is shown in Fig. 6. Points were not taken for the lower values of the anode voltage because of the excessive values of screen-grid current. The mutual family of curves is given in Fig. 7. Under the operating conditions, control-grid voltage = 0.5 volt, screen-grid voltage = 67.5 volts, and anode voltage = 135 volts.

Anode current = 4.0 mA.
Transconductance = 1,100 ua/v.
Anode resistance = 336,000 ohms.
Amplification factor = 400.

The values of the inter-electrode capacitances are :

Input capacitance = 2.5 $\mu\mu$f.
Output capacitance = 0.5 $\mu\mu$f.
Anode-grid capacitance = 0.015 $\mu\mu$f.

Operation

Tests have been made upon both the triodes and the screen-grid valves to determine how well they will perform in conventional circuits at wavelengths much lower than the minimum at which ordinary valves will function.

The minimum wavelength at which a triode will generate oscillations offers a means for comparing it with ordinary valves. The value of this minimum wavelength of oscillation is of particular interest here because it shows how closely a normal feed-back oscillator can approach those wavelengths generated almost solely by Barkhausen tubes and circuits.

An inductive feed-back oscillator was set up whose inductance consisted of several turns of small copper wire wound in a solenoid about one-eighth of an inch in diameter tuned only by the tube inter-electrode capacitances. The circuit is given in Fig. 8. With a coil of six turns very stable 65-centimetre oscillations were produced with as low as 45 volts on the anode of the valve. Smaller coils gave shorter wavelengths with continued stability until a minimum wavelength of slightly below 30 centimetres was reached with a coil of only one turn. Oscillations at this wavelength could be sustained with as low as 115 volts on the anode of the tube and with an anode current of approximately 3 milliamperes.

Owing to the difficulty of making quantitative measurements of radio-frequency amplification at wavelengths of one metre and less, the gain realisable by the use of the screen-grid valves was determined by their operation in actual receiving sets. The first set consisted of two stages of tuned radio-frequency amplification, a detector, and one stage of audio-frequency amplification. The screen-grid valves were used in the radio-frequency amplifier stages and the small triodes as the detector and audio amplifier. The whole set was enclosed in a brass box seven inches long, three inches high, and three inches wide. Small coils, such as those used in the oscillator, tuned by almost equally small variable condensers constituted the tuned circuits. In order to prevent any signal pick-up except through the aerial, the batteries and all external leads were enclosed in metal shielding. With the set so shielded there was no trace of oscillation in any of the circuits. The tuning range of the receiver was from about 95 to about 110 centimetres.

An oscillator operating at a wavelength of 100 centimetres, consisting of one of the small triodes modulated by broadcast receiver and loosely coupled to a half-wave radiator, was set up in an open area. The total anode power supplied to the oscillator was 68 milliwatts.

With the receiver located about 200 feet from the transmitter, signals of good strength were received with the half-wave receiving aerial coupled to the input of the first radio-frequency stage, but none could be heard with the aerial coupled directly to the detector. From other listening tests it was estimated that the gain per stage was of the order of four.

The second receiver was constructed to operate at 75 centimetres or thereabouts. This set was not so elaborate as the one previously described, but more care was taken in placing the valves so that all circuit connections would be shorter than before. It consisted of one stage of radio-frequency amplification

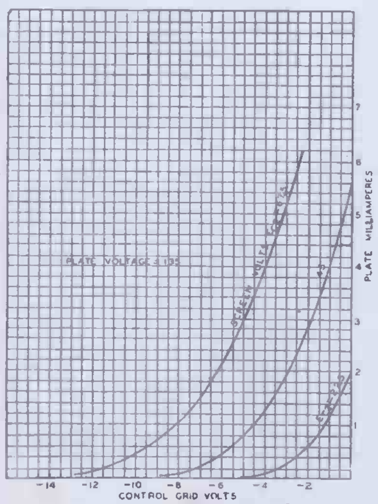


Fig. 7. Mutual characteristics of the small screen-grid valve.

and a grid-leak detector. As before, the set was enclosed in a small brass box and completely shielded.

Inasmuch as it was desired to obtain as high a value of input and coupling circuit impedances as possible, tuning condensers were elimated and use was made of the valve interelectrode capacities only. This made it necessary to fix the tuning of the set. The initial tuning necessary to line up the amplifier and detector circuits at approximately 75 centimetres was accomplished by changing the turn spacing of the tuning coils, thereby varying their inductances. The frequency of the transmitter was adjusted by means of a variable condenser to bring it into tune with the receiver.

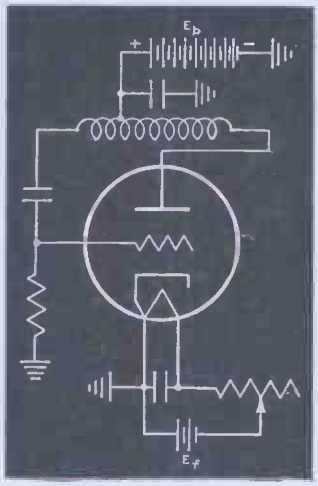


Fig. 8. Circuit diagram of the ultra-high frequency oscillator using the small triode.

Following much the same procedure as before, except that the distance from oscillator to receiver was less and the tests were carried out in a large shielded room, the receiver output when the aerial was coupled to the radio-frequency stage was compared to its output when the aerial was coupled directly to the detector. Again these comparisons were qualitative rather than quantitative. While the contribution of the radio-frequency stage was found to be small, it did furnish some gain as evidenced by the increase in output when it was operating.

New Books

The Meter at Work, by John F. Rider (John F. Rider Publishing Inc., 404, Fourth Avenue, New York.) Price $1.25.

The construction of this book is unique. Actually, it consists of two books, one above the other, in the same binding, the upper part containing all the diagrams and the lower the text. The great advantage of this arrangement is that the reader is always able to have the appropriate diagram in view to correspond with the text being read.

The book itself is of great practical utility, as it shows how each type of meter works, how each is used, and information on the selection of suitable meters. In all, there are 138 diagrams which, together with the text, provide a guide to the use of meters for practically all requirements. The range covered includes moving-iron, moving-coil, dynamometer, electrostatic, thermal meters, rectifiers and thermocouples, and their practical application for all types of circuits.

Understanding Radio, by Herbert M. Watson, Herbert E. Welch and George S. Eby. (McGraw-Hill Publishing Co., Ltd., Aldwych House, London, W.C.) Price 20s.

We believe that this book is the first which teaches radio from the practical standpoint rather than from the theoretical. The general scheme adopted is to describe the purpose of a piece of apparatus and its construction in the simplest manner possible, explain its operation, and finally why it works. The lessons are progressive, commencing with simple experiments with the valve, followed by tuning and examples of various types of receiver, from the single-valve to the superhet, and oscillators and transmitters.

Such matters that cannot well be explained by practical methods—as, for example, wave travel and form—are given very lucid treatment with the aid of many diagrams.

The student who will take the trouble to wire up the simple lay-outs provided and carry out the experiments described cannot fail to obtain a thorough grounding in radio. We can thoroughly recommend this book to those who have entered or are about to enter branches of the Services in which a sound knowledge of radio is essential.

Television Broadcasting, by Lenox R. Lohr. (McGraw-Hill Publishing Co., Ltd., Aldwych House, London, W.C.2.)

It might be inferred from the title of this book that it would only be of interest to those engaged or likely to be engaged in television broadcasting. Actually, this is by no means the case, and it can be read with interest and profit by all those who have followed television development. In the first place, it provides an exceedingly clear exposition of modern television at both the transmitting and receiving ends, and explains its limitations and possibilities with the technical reasons for these as they occur in practice. The book, therefore, is of value to the technician and research worker. It also deals with production in great detail and shows how this is affected by technique; the relevant matter given is essential to producers, operatives and artists. The information in this respect is based largely upon a large number of tests and experiments made by the National Broadcasting Company and is, therefore, of American origin, a fact which, however, does not detract from the value of the work in this country, because during the whole period of experimental development a close watch was kept upon British methods. The price of the book is 21s.; it contains 274 pages and is well illustrated.

Gas-filled Triodes and their Practical Use—VII (Conclusion)

Methods of Testing Gas-filled Triodes

By G. Windred, A.M.I.E.E.

Efficient application of the gas-filled triode relies upon accurate knowledge of its properties under given conditions of operation. In this article, the last of the present series, consideration is given to methods of testing which may be employed for determining the characteristics of principal importance in practice.

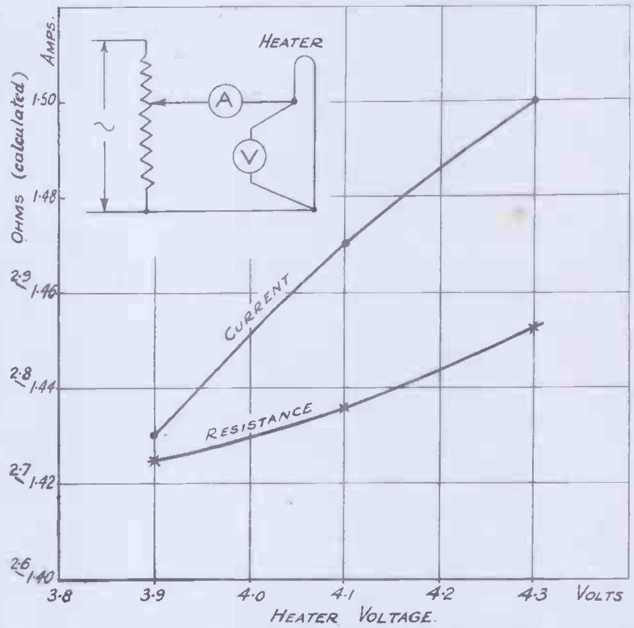


Fig. 1. Relationship of heater current and resistance to heater voltage (GT1C tube).

ALTHOUGH it would be difficult to exaggerate the practical importance of accurate information concerning the properties of gas-filled triodes, there is no doubt that the subject of testing has hitherto received very little attention from the user. It has been implicitly assumed that the question of testing was one for the manufacturer, and that the average user had neither the facilities nor the knowledge of electronics required for making test measurements.

While it is true that some aspects of testing are appropriate only to the manufacturer with laboratory equipment developed specially for the purpose, it must not be overlooked that some knowledge of testing, even if comparatively elementary, is desirable on the part of the user. It is evident that a great step forward in the industrial application of the gas-filled triode will have been made when it can be subjected to routine tests and examination independently of the manufacturer. Tests of this kind may be classified under two headings, according to whether their object is to determine the suitability of the device for a particular condition of service, or simply to check its conformity with the manufacturer's data.

Much of the difficulty arising in connection with applications of the gas-filled triode in the past may be ascribed to incorrect use, and it seems reasonable to suppose that trouble from this cause would be considerably reduced by a more general understanding, not only of the factors affecting operation, but also of the means which may be employed for the experimental determination of characteristics. One of the primary objects of such measurement must be to ensure that the intended conditions of operation are within the capabilities of the tube, so as to avoid premature failure due to an insufficiently conservative rating.

It cannot be too strongly emphasised that the question of testing gas-filled triodes calls for the closest possible collaboration between the user and manufacturer. It is evident that in all matters of this kind the final authority must rest with the manufacturer, but it is very desirable for the user to be able to investigate matters pertaining to specific applications, especially where the conditions are unusual.

The following discussion of methods of testing deals with those properties of the gas-filled triode which are of chief importance in practice.

1. *Heater Voltage and Current*

It is of the greatest importance that the heater should be operated always within a small percentage of its nominal current and voltage. In this connection it is useful to have advice from the manufacturer concerning the maximum permissible variation of heater voltage, since a variation of some kind, however, slight, is to be expected under all normal conditions of operation.

It will be realised that this question is related to the accuracy with which the heater itself is proportioned, and some small variation of resistance is to be expected among tubes of the same type. For this reason it is not practicable to apply very close limits to the heater current at a specified voltage, but some check upon the current taken by the heater is necessary. A convenient method is to apply an alternating or direct voltage to the heater, of approximately the nominal value, and to adjust the value exactly after the heater has reached a stable temperature. The heater current measured under these conditions should be within a few per cent. of the same value for all tubes of a given type.

An alternative method is to employ a source of fixed voltage such that when the heater has warmed up, its terminal voltage lies between fixed limits in the region of the nominal value. The permissible limits of heater current in this case will naturally be wider than with the first method. It must be remembered that the heater will have a temperature coefficient of resistance, so that the range of current will be smaller than the corresponding range of voltage. As an approximate general rule, it is usually permissible to work on the assumption of a 5 per cent. variation of voltage.

The curves in Fig. 1 show the relationship of heater current and resistance to the applied voltage in the case of a typical tube. It will be seen that the resistance rises at a faster rate than the voltage, owing to the increasing temperature, while the current curve shows a corresponding drop. The temperature coefficient thus has the desirable effect

Methods of Testing Gas-filled Triodes—*Windred*

of stabilising heater current changes with change of voltage.

The design of transformers for heater supply should take into consideration the load variation with voltage as represented in Fig. 1 so as to obtain close regulation of terminal voltage with variation of line voltage. It may also be necessary to allow for voltage drop in the conductors and connections between the transformer terminals and the actual heater pins of the tube. This voltage drop varies according to the circuit arrangements, but an allowance of about 0.05 volt is generally sufficient. In the case of testing equipment, heater voltage variation for adjustment purposes may be obtained by means of a Variac, for example, or a suitable potentiometer.

Care must be exercised in the selection of suitable instruments in order to avoid errors of measurement. The best arrangement is represented by the diagram in Fig. 1, where the voltmeter is a high-resistance instrument. If a low-resistance voltmeter is used it must be connected to the supply side of the ammeter, so that its current is not added to the ammeter reading. In this case the ammeter resistance must be small in comparison with the heater resistance.

2. *Grid Control Characteristic*

This property is more important in some applications than others, accord- to the required accuracy of control. In the case of so-called " trigger " applications, for example, where the flow of anode current is established and interrupted without being controlled closely in magnitude, precise information regarding the grid control characteristic is not so important as in cases where close control over the anode current is required. In cases where anode current variation is obtained by changing the phase of the grid voltage it is often necessary to have accurate knowledge of the grid control characteristic, which determines the corresponding changes of anode current.

Another aspect of the subject arises when it is necessary that performance should not be altered by changing the tube for another of the same type. Otherwise it is necessary to incorporate in the circuit some means of obtaining the required operation with a tube having a different characteristic. A certain amount of change is inevitable during the normal life of the tube, and for some applications it is necessary to determine the amount of this change as well as the characteristics of a new tube. Cases may also arise where two or more tubes employed in the same apparatus require to have closely similar characteristics. In all these instances it is necessary to be able to measure the grid control characteristic with reasonable accuracy.

One method of measurement is to apply an adjustable direct voltage to the anode, while the grid is maintained sufficiently negative to prevent conduction, and then to reduce the grid voltage gradually until conduction occurs. The particular value of grid voltage at the instant of striking is the critical value for the selected anode voltage. Repetition of this process for several values of anode voltage allows of the complete characteristic being drawn. As a variation of this method, the anode voltage may be increased to the striking value while the grid voltage is kept fixed. This process yields similar results, but one or other method should be adhered to so as to avoid confusion in comparing when the tube strikes is the critical value for the particular value of V_a selected. It should be noted that if grid circuit resistance (R_g) is present, the reading of the meter giving V_g will drop when the tube strikes, owing to the flow of grid current through this resistance. It is the value of V_g *just prior* to this drop which represents the critical value.

The drop, which may be confusing, may be reduced by short-circuiting the resistance during the tests, but a slight drop will still occur owing to the flow of grid current through the potentiometer itself. This effect may be minimised by using the lowest resistance potentiometer consistent with sufficiently smooth regulation and reasonable drain upon the battery. It will be

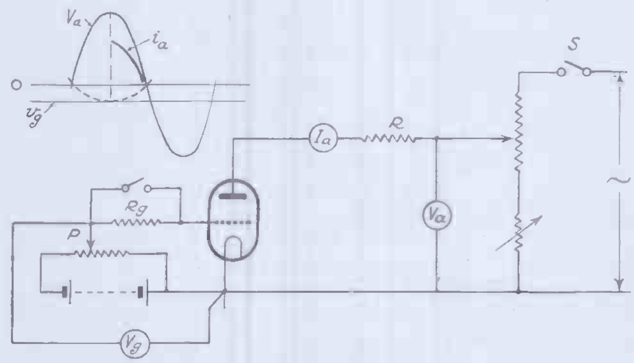


Fig. 2. Principle of operation and circuit used for determination of grid control characteristic.

measurements. In either case it is very important that the respective voltages should be entirely free from ripples or variations, since otherwise errors of measurement will arise. Battery supplies are preferable for this reason, and if mains are used it is necessary to ensure absence of ripple by means of suitable smoothing arrangements.

A rather serious disadvantage of the use of D.C. for the anode circuit is that when the tube strikes there is no further control by the grid, and the anode circuit has to be broken and reclosed before another reading can be taken. A better method is to employ A.C. on the anode and D.C. on the grid, so as to obtain continuous control over the anode current. The basic circuit diagram and conditions of operation are shown in Fig. 2.

In operation, the tube is allowed to warm up on its appropriate heater supply before making the test. The grid circuit potentiometer is adjusted so as to give a high negative grid potential, and the alternating anode voltage is then applied by closing the switch S. The main potentiometer is then adjusted to give some arbitrary value of anode voltage, say 50 volts, and the grid potentiometer then moved very slowly in the direction to reduce the grid voltage. The value of grid voltage (V_g) understood that if grid circuit resistance is used, its value should be the same for all tests on the same type of tube, since variations will change the critical voltage reading.

It must be carefully noted that with this method of test the critical voltage applies not to the voltage indicated by the A.C. voltmeter, but to the peak value corresponding to its r.m.s. reading, as will be seen from Fig. 2. If, for example, the r.m.s. value of V_a, as shown by the meter, is 50 volts, then the corresponding critical grid voltage will apply to $V_a = 1.41 \times 50$, or 70.5 volts, provided that the alternating voltage wave approximates closely to sine shape. The plotting of curve ordinates is facilitated by taking meter readings of V_a which correspond to convenient values of peak voltage. For example, to fix points of the curve corresponding to 50, 100, 150 and 200, the meter readings selected would be 35.5, 71, 106.5 and 142 approximately, to the nearest scale divisions.

If desired, crest values of V_a may be measured directly by a peak voltmeter, but it may be questioned whether this method offers any great advantage. The fixed resistance shown in the anode circuit in Fig. 2 serves simply to limit the anode current to a value within the tube rating. The adjustable

Methods of Testing Gas-filled Triodes—*Windred*

resistance in series with the main potentiometer serves as a vernier to allow close adjustment of anode voltage.

In the case of mercury-vapour tubes it is necessary to carry out the tests at a constant temperature, which should be specified in the test results owing to the considerable effect of temperature upon the grid control characteristic. Tubes with inert gas filling do not show the effect to such a marked degree, but it is always present to some extent, and represents a possible source of error in comparing tube characteristics.

It may be mentioned that the use of a transformer in the anode circuit is to be avoided, owing to its tendency to cause distortion of wave form in the anode voltage. This distortion destroys the relationship between r.m.s. and peak value of voltage upon which the measurements are based, and is caused by the unidirectional pulsating flux in the transformer core resulting from the flow of anode current during positive half-cycles. If the distortion ceased immediately upon cessation of anode current, the effect would not be important, since the critical stage is when anode current is just on the verge of flowing, as it were; but in practice the distortion may take several cycles to die away, and thus upsets the conditions at the point of striking. The difficulty is overcome by using a non-inductive potentiometer for the anode voltage adjustment.

When making tests it is necessary to repeat each determination of critical grid voltage two or three times. If the values obtained vary by more than a few per cent. there may be disturbances in the circuit conditions, caused possibly by insufficiently smooth voltages. With suitable conditions it is possible to reproduce readings with considerable accuracy, but it is always advisable to make allowances for experimental error. This tolerance can be decided on the basis of a series of measurements with tubes of a given type.

3. *Grid Current*

The amount of grid current drawn by a gas-filled triode is of varying importance, according to the conditions of operation, and in some cases it must be known for purposes of circuit design. Since the current to be measured may be only a few microamperes it is necessary to employ sensitive methods under carefully controlled conditions in order to ensure reasonable accuracy.

Since with a given tube the grid current depends upon whether anode current is flowing or not, the amount of grid circuit resistance, the grid voltage, and its polarity relative to the cathode, it will be realised that the conditions of test must be completely specified before any significance can be attached to the results of grid current measurements. Owing to the complication caused by the use of alternating voltages on anode and grid it is usual to employ direct voltages for these measurements. There are three principal methods, as follows:—

(a) A sufficiently sensitive meter is connected in the grid circuit so as to give a direct indication of the current flowing.

(b) The grid current is calculated by observing with a high-resistance voltmeter, such as a thermionic voltmeter, the drop across a resistance of known value included in the grid circuit.

(c) The grid current is calculated by observing the change of applied grid voltage to cause striking of the tube when the grid circuit resistance is changed by a known amount.

This last method has the advantage of requiring only standard apparatus, and is the one employed in general practice. It is based on the principle that the flow of grid current will cause a definite drop in a known resistance, so that when the resistance is incorporated in the circuit the applied grid voltage corresponding to the critical voltage on the grid must be greater than the corresponding voltage without the resistance by an amount equal to the drop in the resistance. In short, a greater grid voltage must be applied in order to compensate for the drop in the resistance caused by the grid current.

The circuit shown by Fig. 2 may be used for measuring grid current in this way. The grid circuit resistance R_g is for convenience made 1 megohm, and may be short-circuited by a switch as shown in order to give the condition $R_g = 0$. With this condition, and a predetermined value of anode voltage, a note is made of the critical grid voltage at which conduction occurs. This voltage may be denoted by V_{g1}. With the same value of anode voltage, the determination is repeated with the switch open, so that R_g is in circuit. It will be found that the striking condition is now reached with a higher voltage, which we can call V_{g2}. The grid current is accordingly given by $i_g = (V_{g2} - V_{g1})/R_g$. The convenience of making $R_g = 1$ megohm arises from the fact that the numerical value of voltage difference is then equal to the grid current in microamperes.

In order to ensure consistent results it is necessary to arrange the circuit with protection against outside disturbances, such as pick-up, and to adopt a definite test procedure. Short connections and suitably earthed screened wiring are highly desirable. The readings are also influenced by tube temperature, so that a test carried out immediately after the lapse of the normal preheating time will yield results differing from those obtained with the same tube at a higher temperature, due either to a long preheating time or a definite period of anode current flow. Actual conditions are best represented by allowing a moderate anode current to flow before making the measurements. There is then the added advantage that the tube temperature remains more constant during the tests. These are matters which may readily be settled as experience is gained, but it is important that they should be appreciated at the outset.

Grid current measurements may be made with alternating or direct anode voltages, and the results will naturally differ in the two cases. For purposes of comparison and record it is preferable to make the measurements with a D.C. anode supply at a given voltage, and with a definite value of anode current limited by a fixed resistor.

4. *Arc Drop*

The voltage drop between anode and cathode in the conducting state is important for two reasons. Firstly, it has a bearing on the design of the circuit, and secondly, it provides some indication of the state of the tube. The measurement is usually made with a D.C. anode circuit supply.

It is not possible to measure the arc drop with an ordinary voltmeter, owing to the disturbing effect of its relatively low resistance, but approximate measurements are generally possible with an instrument having a resistance of not less than one thousand ohms per volt. Better results are obtained with a thermionic voltmeter. As an alternative a wattmeter method may be employed, the procedure being as follows:

The current coil of the wattmeter is connected in the anode circuit, in series with a moving-coil ammeter indicating the average value of rectified anode current, an alternating anode voltage being used. The voltage coil of the wattmeter is connected from anode to cathode, so as to receive the arc drop voltage. The grid is normally connected to cathode or made slightly positive in order to ensure full conduction. Under these conditions the wattmeter reading represents the power dissipated in the arc and the instrument itself. For a given anode current the meter loss can be measured by connecting the voltage coil across the current coil and observing the meter reading. If the normal meter reading with a given anode current I_a is denoted by W

Methods of Testing Gas-filled Triodes—*Windred*

and the meter loss is W_L, the voltage drop in the arc is given by

$$V = \frac{W - W_L}{I_a} \text{ (volts)}$$

In cases where tests on different tubes are being made, or where the arc drop is measured with varying values of anode current, it is convenient firstly to plot a curve of meter loss W_l in relation to various values of I_a throughout the range of anode current to be observed. There is sometimes difficulty in applying this method owing to the smallness of the wattmeter reading in the case of small gas-filled triodes. The method is useful as a means of measuring the arc drop with an alternating anode voltage, as distinct from the value corresponding to D.C. operation.

It is generally assumed that the arc drop is independent of anode current, but in practice it will be found that this is not so. Even in the case of inert gas-filled tubes, there is a variation of drop with variation of anode current, and within a given range of current the drop may rise or fall as the current is increased. In a batch of new tubes of a given type the arc drop with a given anode current may differ appreciably from tube to tube, although the average arc drop over a range of anode current may be approximately constant. As a general rule, for tubes with inert gas or mercury vapour filling, an arc drop of more than 20 to 25 volts results in damage to coated cathodes owing to excessive positive ion bombardment. Limiting values of arc drop are usually specified by the manufacturer for each type of tube. Owing to inevitable variations in manufacture, some tolerance must be allowed, and this should preferably be made as large as possible consistent with satisfactory behaviour under working conditions.

Considerable evidence regarding the conditions in a gas-filled triode may be obtained by oscillographic observation of the arc drop when the tube is acting as a rectifier. Two typical conditions are represented by the diagrams in Fig. 3, obtained by connecting the cathode-ray oscillograph between anode and cathode of the tube. In both cases the conduction period was the same; somewhat less than a half-cycle. In the case of record (a) the arc drop was approximately constant throughout the entire period of conduction, whereas record (b) shows a rise of voltage which indicates considerable variation of arc drop with the current. This property is undesirable, and is often an indication of a faulty tube with little further useful life. More evidence on this point is forthcoming if the oscillograph scale is previously calibrated in volts. The curves in Fig. 3 correspond to a non-inductive anode circuit, which is advisable in these tests in order to eliminate the uncertainties caused by differing values of inductance.

5. *Inter-Electrode Resistance*

The possibility of unduly low insulation resistance between the tube electrodes must be guarded against, and some means of checking this factor is desirable before the tubes are put into service. The insulation resistance between the grid and the remaining electrodes is of particular importance, especially in cases where high grid circuit impedances are normally employed. In the case of the gas-filled tetrode, for example, the grid circuit may have an impedance of ten megohms in some cases, so that an insulation resistance of several times this amount is necessary for the control grid in respect to the other electrodes.

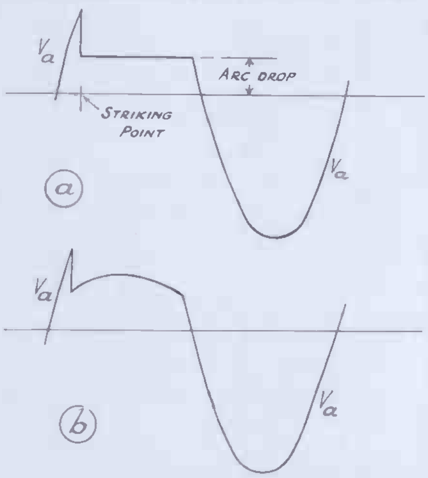


Fig. 3. Typical oscillograms of arc drop in gas-filled triodes.

Ordinary methods of measuring the insulation resistance introduce two main difficulties, as follows :—

(a) If the voltage used exceeds the ionisation potential of the gas filling, which may be as low as approximately 10 volts, there is a possibility of discharge occurring in the tube which will upset the measurements.

(b) Insulation resistance should be measured with the tube at operating temperature, so as to detect any faults which may not exist with a cold tube. At this temperature, however, there is a possibility of thermionic emission from the cathode which will facilitate current flow under some conditions of test.

The first difficulty may be overcome by using any of the forms of instruments for resistance measurement which employ small voltages and have an adequate range of measurement. The second difficulty may be avoided by ensuring that when the cathode forms one electrode during the measurement it is connected to the positive terminal of the measuring voltage so as to prevent conduction due to thermionic emission. An alternative is to allow the cathode to cool down well below emission temperature instead of applying the test immediately after disconnection of the heater.

6. *Life Tests*

The life of a gas-filled triode cannot be accurately predicted, but is very largely influenced by the treatment it receives in use. In some cases it is difficult to predetermine such important factors as average and peak values of anode current, although they may be measured quite readily under operating conditions. Any doubts concerning tube life may be checked by carrying out life tests, but it is essential that the test conditions should approximate as closely as possible to those existing in practice.

If tests are made with a particular type of tube to determine its life when operated at the maximum specified rating, care should be taken that the average and peak anode currents do not exceed the specified maximum. Average current may be measured on an accurate moving-coil meter, but the determination of peak current cannot be made directly. A suitably calibrated cathode-ray oscillograph connected across a known resistance in the anode circuit provides a convenient means of measuring peak values of current, or alternatively a peak voltmeter may be used in a similar manner.

The importance of peak current ratings arises mainly in connection with control of anode current by phase displacement of the grid voltage. In some cases the peak current is very much greater than the average value; perhaps ten or fifteen times as great, and it then becomes necessary to rate the tube purely on a basis of peak current, since the ratio of peak to average anode current rating rarely exceeds a factor of 6, and is often considerably less.

One of the most important rules in the application of gas-filled triodes is that the rating shall be conservative. If this condition is met, and the heater is always preheated for the appropriate period as well as operated as nearly as possible on the correct nominal voltage, the main requirements for maximum life will be fulfilled.

A RECORD OF PATENTS AND PROGRESS

RECENT DEVELOPMENTS

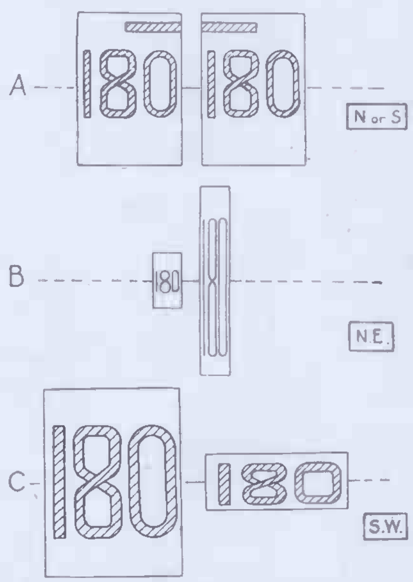
PATENTEES

Marconi's Wireless Telegraph Co., Ltd., and R. J. Kemp :: Radio-Akt. D. S. Loewe :: A. F. Pearce :: D. Jackson and Pye Ltd. :: Philips Lamps Ltd. :: Photoswitch Ltd. :: B. Chance

Navigation by Television

(Patent No. 527,841.)

IN wireless navigational systems it is common practice to radiate two beams from a beacon station so that they overlap slightly along their length. One beam carries a series of Morse "dots," and the other a series of Morse "dashes," the two sets of signals being so timed that they alternate or "sandwich" with each other. The navigator of a distant vessel can then tell when he is on the centre-line between the two beams by noting when the dots and dashes merge together into a continuous note. If he is to one side or other of that centre-line, the navigator will hear either a series of separate dots or a series of separate dashes, and so know that he is either to port or starboard of the continuous signal or correct bearing line.



Using television signals for navigation. Patent No. 527,841

Instead of using Morse signals, the inventors propose to replace them by television signals which give a picture of the actual compass bearing of the distant beacon station, relative to the point at which the signals are received.

The arrangement is such that the bearings are duplicated and appear of equal size as at A only along the equi-signal line joining the beacon station to the receiver. At "off course" points, one picture will appear to be either longer than the other as at B, or broader than the other as at C.—*Marconi's Wireless Telegraph Co., Ltd., and R. J. Kemp.*

Luminescent Screens

(Patent No. 527,843.)

Although a certain amount of "afterglow" is desirable on the fluorescent screen of a television receiver, since it helps the persistence-of-vision effect and so tends to prevent flicker, the same considerations do not apply to a television transmitter of the kind in which the luminescent scanning-screen of a cathode-ray tube is used to energize a photo-electric cell.

Here the presence of afterglow is found to give rise to "blurring," and a noticeable lack of definition, particularly in the brighter parts of the reproduced picture. In order to avoid such distortion, the luminescent screen of the cathode-ray transmitter is backed by a hollow copper support through which a continual stream of cooling water is passed. The water is maintained in constant circulation by a feed pipe and a discharge pipe, both sealed into the glass wall of the cathode-ray tube. When so cooled, the luminescent screen is found to be practically free from afterglow. —*Radio-Akt D.S. Loewe.*

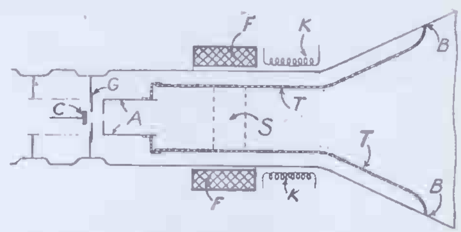
Cathode-ray Tubes

(Patent No. 527,980.)

Owing to the high operating voltages commonly used in a cathode-ray tube, it is important to prevent any leakage or discharge over the glass surface of the tube, say, from the anode (which may be carrying anything from 10,000 to 50,000 volts) to the other electrodes, or to the external scanning coils.

With this object in view, the anode A is supported, as shown in the Figure, at one end of a long tube T of glass which is fused at the other end B to the glass bulb of the cathode-ray tube, at a point well removed from the other electrodes. The two ends of the support may be made of different kinds of glass, one suitable for welding to the glass bulb, and the other for sealing to the metal anode, both being joined to a common section S as indicated by the dotted lines. This facilitates the process of assembly.

The tube T may be coated with an inside film of metal, so that, in effect, it forms a prolongation of the anode A. Its point of contact B with the glass bulb of the CR tube is, however, sufficiently far removed from the cathode C, grid G, focusing coil F, and scanning coil K to minimize any danger of high-voltage leakage from the anode to these parts via the glass surface of the bulb.—*A. F. Pearce.*



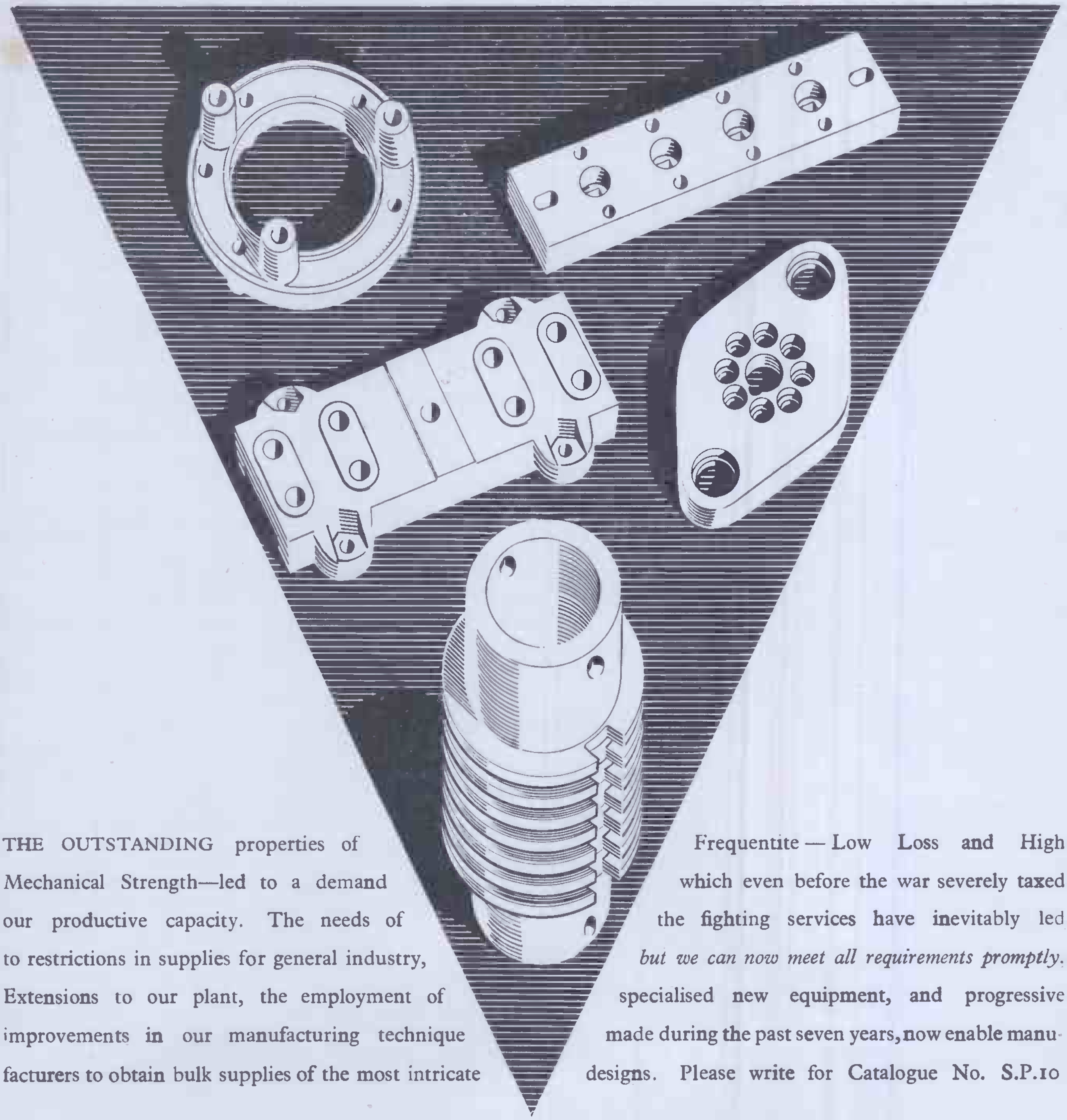
Preventing leakage in cathode-ray tubes. Patent No. 527,980

Television Receivers

(Patent No. 528,198.)

To facilitate the carrying out of necessary repairs or renewals, the chassis S of a television receiver is so mounted that it can be moved bodily from inside the cabinet, and then turned into an inverted position, as shown in dotted lines, or, if necessary, completely detached, without having to use any tool.

FREQUENTITE



THE OUTSTANDING properties of Frequentite — Low Loss and High Mechanical Strength—led to a demand which even before the war severely taxed our productive capacity. The needs of the fighting services have inevitably led to restrictions in supplies for general industry, *but we can now meet all requirements promptly.* Extensions to our plant, the employment of specialised new equipment, and progressive improvements in our manufacturing technique made during the past seven years, now enable manufacturers to obtain bulk supplies of the most intricate designs. Please write for Catalogue No. S.P.10

STEATITE & PORCELAIN PRODUCTS LTD.

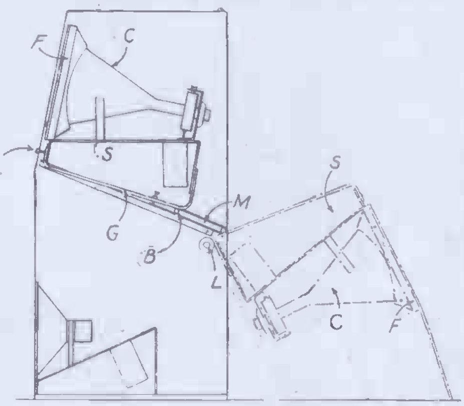
Pioneer Manufacturers of British Low Loss Ceramics

Head Office and Works :
STOURPORT - ON - SEVERN, WORCS.

Telephone: Stourport 111. Telegrams: Steatain, Stourport.

SP.3

The upper surface of the chassis supports the cathode-ray tube C in a frame F, whilst the underside carries the other usual circuit components (not shown). By first loosening a screw knob A on the front panel, the chassis will slide bodily along a pair of inclined guide rails G, secured to the sides of the



Accessible receiver construction. Patent No. 528,198

cabinet. At the rear end of the guide rails are two U-shaped members M which enclose the ends of a crossbar B supporting the chassis, so that the latter can be turned upside down once it is brought outside the cabinet. To remove it completely, one of a pair of buffers L is moved to one side, so as to provide sufficient play to allow first one end and then the other of the crossbar B to be slipped out of the U-shaped guides M.—*D. Jackson and Pye, Ltd.*

A Studio Problem

(Patent No. 528,354.)

The television studio in which a play is being performed for broadcasting is usually lighted by powerful gas-filled discharge lamps of the high-pressure mercury-vapour type. The lamps are generally fed from A.C. supply mains, and since they have a comparatively low " time constant " they tend to fluctuate in brilliancy at the frequency of the A.C. supply. The mean brightness of the image screen in the television transmitter tube is naturally affected, the result being that the image reproduced at the receiving end shows undesirable light and dark bands.

In order to overcome this defect, an auxiliary photo-electric cell is inserted in circuit with the television transmitter, and is exposed either directly to the light from the studio lamps or to the light from the image screen of the transmission tube. It thus develops a fluctuating voltage, which is then combined with the image or signal currents in such a way as to oppose or neutralize the original undesirable effect.—*Philips Lamps, Ltd.*

Controlling Textile Machinery

(Patent No. 528,543.)

Relates to a photo-electric device for supervising the movement of a series of threads, such as are handled in beam-warping and other textile machinery. Stop-motions are already known for automatically stopping the machine, or for sounding an alarm, if any one of the series of threads should break. They are, however, somewhat complicated, and are not always reliable if the threads are very light or frail and therefore unable to apply only a reasonable motivating force.

According to the invention, if any of the threads to be supervised either breaks or sags, it is made to obstruct, either directly or indirectly, the passage of a ray of light on to a photo-electric cell. The interruption of the light automatically actuates a relay, which either stops the machine at once or else sounds an alarm to draw the attention of the operator to the faulty thread.—*Photoswitch Inc.*

Automatic Steering

(Patent No. 528,562.)

The Figure shows, in skeleton, the control elements by which a ship or other navigable vessel can be kept to a set course, any tendency to stray either to port or starboard being automatically corrected.

A ray of light from a source S (Fig. A) falls on to a prism P mounted at the centre of a com-

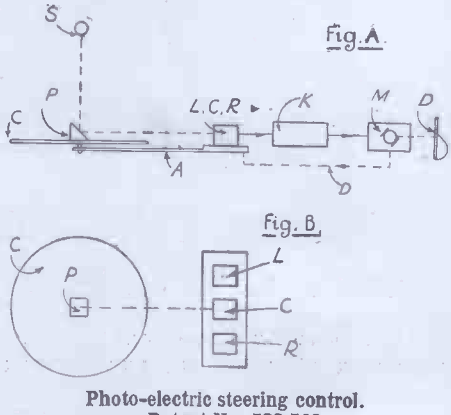


Photo-electric steering control. Patent No. 528,562.

pass card C, and is then reflected on to a row of photo-electric cells L, C, R (shown separately in plan in Fig. B). The cells are carried by an arm A, which is pivoted about the compass so that it swings relatively to it when the craft yaws to one side or other off its course.

Any such movement will cause the reflected ray of light to shift away from the centre cell C and fall either on to the cell R or the cell L. The resulting current is used to operate a relay circuit which, in turn, controls the rotation of a motor M so that it moves the rudder D in the proper direction to bring the craft back on to the correct course. At the same time, a follow-up mechanism (indicated by the dotted line D) causes the motor M to restore the photo-electric cells to the original " balanced " condition.

The invention is concerned with specific improvements in the relay and follow-up mechanism designed to improve the effectiveness of a photo-electric control device of the kind illustrated.—*B. Chance.*

Summary of other Television and Electronic Patents

(Patent No. 524,443.)

Three-colour system of television in which the signals representative of each primary colour are mutually compared so as to ensure an output of correct colour value.—*G. Valensi.*

(Patent No. 524,776.)

Television receiver with push-button tuning of the intermediate-frequency stages to receive alternative programmes. — *Kolster-Brandes, Ltd., and C. N. Smyth.*

(Patent No. 525,425.)

Short-wave oscillation generator for applying therapeutic treatment to patients.—*C. R. Burch and Metropolitan-Vickers Electrical Co., Ltd.*

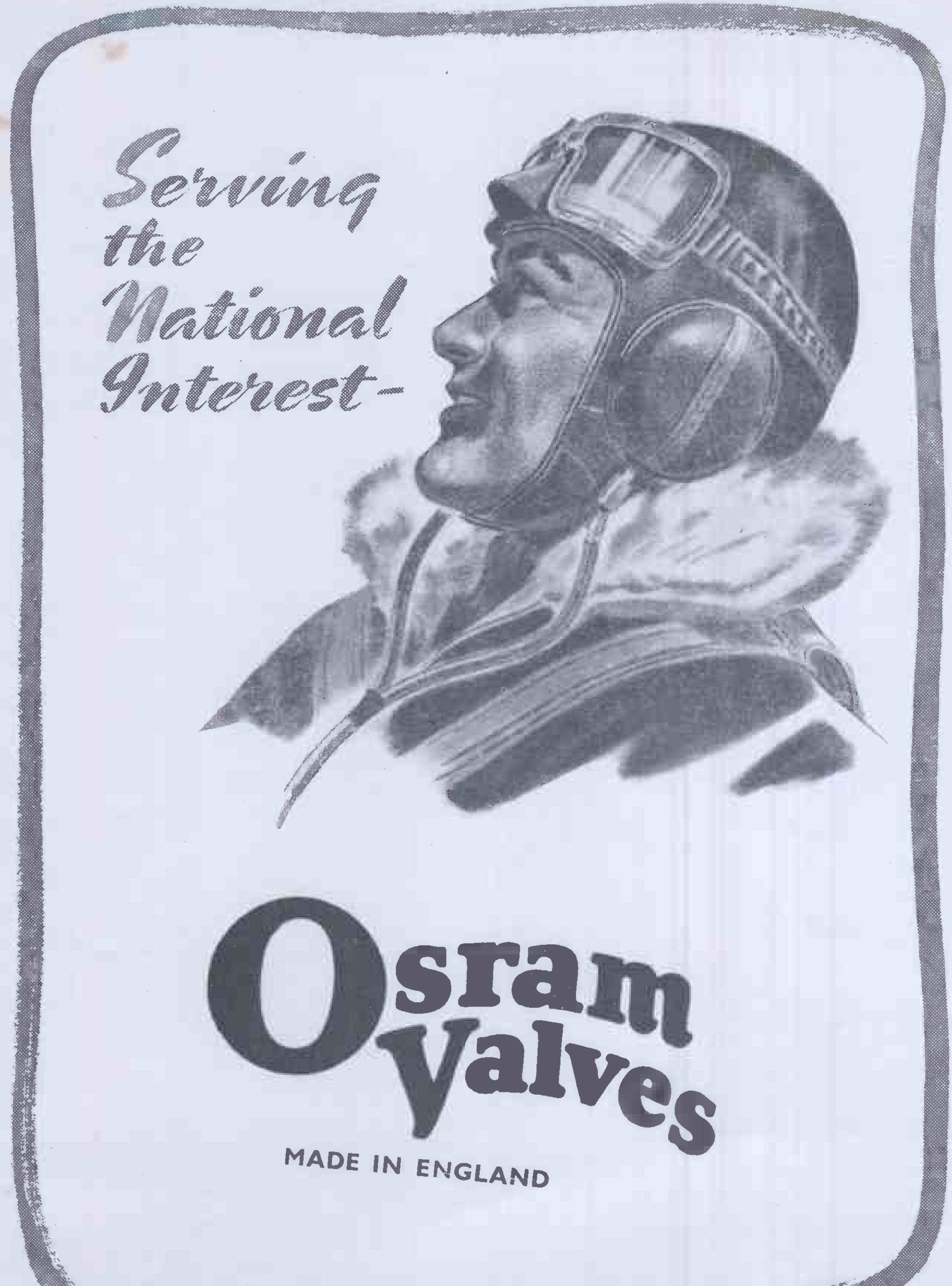
(Patent No. 525,608.)

Electric discharge tube for converting light of one wavelength projected on to a photo-electric surface into light of another wavelength projected on to a fluorescent screen.—*The Mullard Radio Valve Co., Ltd.*

(Patent No. 525,543.)

Television receiver in which a viewing screen of the " storage " type allows a low frame repetition frequency to give a picture free from flicker. — *Radio-Akt. D.S. Loewe.*

(More patents on page iii of cover)


Serving the National Interest-
Osram Valves
MADE IN ENGLAND

Simple Meter Conversions

By W. A. Flint

This article describes simple methods of adapting ordinary meters for a variety of purposes

THE advantages of the moving-coil type of meter for D.C. measurements are well known. It has high sensitivity, good damping, a long scale arc, low power consumption, short time period and high torque-weight ratio.

By the addition of suitable resistances, a D.C. moving-coil milliameter can be adapted to measure D.C. volts; by shunting the meter with a resistance, its current measuring range can be extended; and by adding a suitable metal rectifier it can be easily adapted to read alternating quantities and retain all the advantages of the D.C. instrument itself.

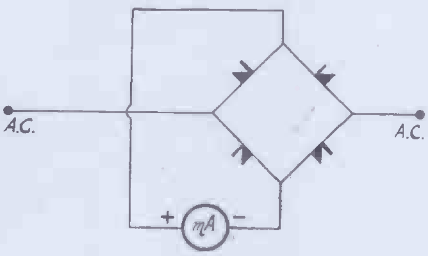


Fig. 1. A.C. milliameter.

D.C. Voltmeter

When designing a multi-range AC/DC test set, it is usual to start with a 1 mA. milliameter in order to get a resistance of 1,000 ohms per volt, i.e., 1 volt through 1,000 ohms gives a current flow of 1 mA. This enables accurate readings to be obtained, and such a meter will not greatly influence the readings to be obtained when reading voltages where there is already a current flow through a resistance, i.e., when reading the screen-grid voltage applied through a voltage dropping resistance.

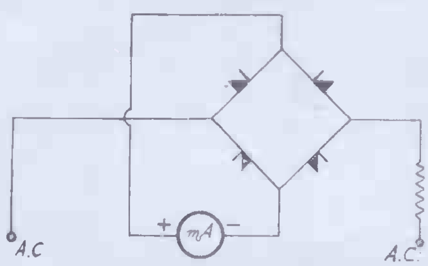


Fig. 2. A.C. voltmeter.

Most 1 mA milliameters have scales calibrated into 50 divisions and by the suitable choice of full scale voltage ranges, most of the readings can be made direct. For example, on a 10-volt range, a reading of 0.2 mA. would indicate 2 volts; on a 100-volt range 20 volts; and on a 1,000-volt range 200 volts; whereas on a 500 volts full scale range, for example, a reading of 0.2 mA would indicate 100 volts, it being necessary to multiply the reading by 50—or the dial could be calibrated.

Three voltage ranges should be sufficient for all ordinary purposes and they may conveniently be 0-10 volts, 0-100 volts and 0-1,000 volts. The first range would be suitable for measuring cathode voltages generally, the second range screen voltages, and the third for general H.T. measurements.

The ranges may easily be adjusted to suit individual requirements and the values of the resistances may be easily calculated from ohms law. Since the current flow is always 1 mA., the value of the resistance for a certain full scale range becomes 1,000 times the full scale voltage required. Thus for a 10 volts range, a resistance of 10,000 ohms is required.

As the accuracy of the finished instrument depends on the accuracy of the resistances and shunts, it is advisable to have these specially tested for accuracy, unless a meter of known accuracy is available with which to calibrate each and every range individually.

When the meter movement is shunted by a resistance, and current flows through the circuit it has two paths open to it, i.e., one through the meter movement and the other through the shunt resistance. If, therefore, the value of the shunt is arranged so that no matter what the total current through the circuit, only 1 mA. will flow through the meter movement, it will be possible to measure higher ranges of current. In other words, to measure a full scale current of 10 mA., a shunt resistance that will pass 9 mA. must be used. For 100 mA. the shunt must pass 99 mA., and so on.

If the meter resistance is 100 ohms, the value of the shunt to convert it to a 10 mA. meter will be $\frac{100}{9}$ or 11.1 ohms, for 100 mA., 1.01 ohms, and so forth.

Suitable shunts and specially selected resistances are available or the shunts may be wound from resistance wire. For the higher ranges of current, care must be taken to select a wire of sufficient carrying capacity and it is usually preferable to use a single shunt and tap off for the various ranges. If it is desired to have ranges of 10, 100 and 1,000 mA., the approximate length of the wire required for the 10 mA. shunt (11.1 ohms) is calculated and the shunt adjusted against a meter of known accuracy as follows:—Exactly 1 mA. is passed through the meter with no shunt connected. The 10 mA. shunt is then switched in (it is advisable to keep any switch in circuit when adjusting shunts

	Switch A	Switch B	
D.C. volts	1	10v.	1
		100v.	2
		1,000v.	3
D.C. amps.	2	10mA.	1
		100mA.	2
		1,000mA.	3
A.C. volts	3	10v.	1
		100v.	2
		500v.	3

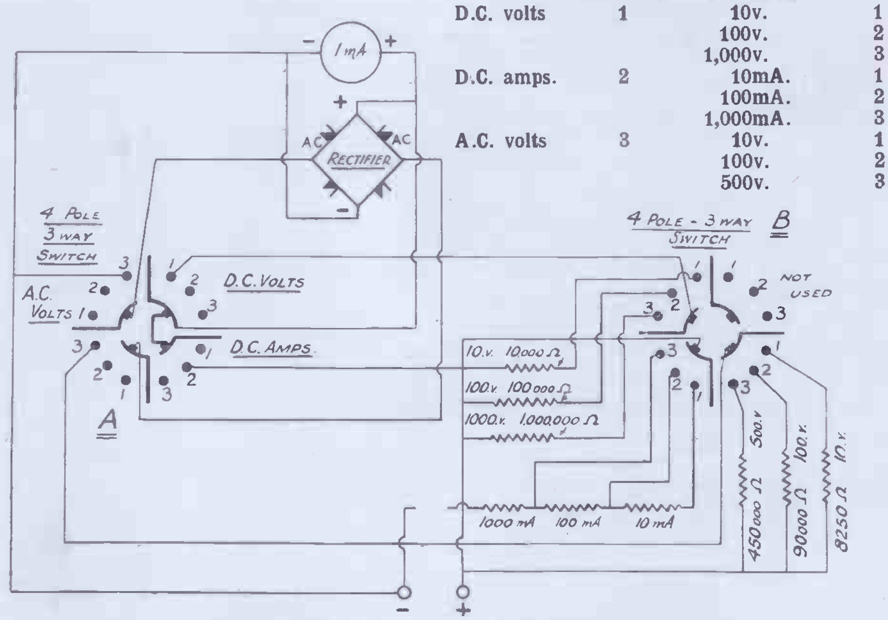


Fig. 3. Circuit diagram of A.C./D.C. test set.

in order to allow for the resistance of the switch contacts) and then the shunt is adjusted so that exactly 0.1 mA. flows through the meter, indicating that 0.9 mA. is flowing through the shunt, so that when reading 10 mA., 1 mA. would flow through the meter and 9 mA. through the shunt.

When the 10 mA. shunt is correct, the circuit is arranged so that exactly 10 mA. flows and then the shunt is tapped off (again keeping any switching in circuit) until the meter again reads 0.1 mA. indicating that the 10 mA. shunt is multiplied by 10 and that the full scale range is now 100 mA. The same procedure may be adopted for further ranges.

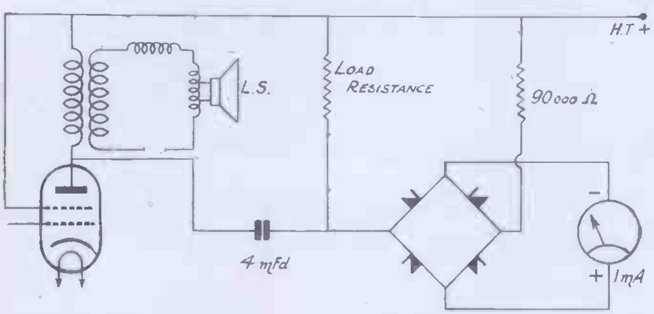


Fig. 4. Use of the 100 volts A.C. range as an output meter.

A.C. Milliameter

The addition of a metal rectifier to the milliameter is shown in Fig. 1 and the effect is to adapt the 1 mA. D.C. instrument to act as an A.C. milliameter with 11 per cent. higher full scale deflection, due to the fact that the instrument movement is still giving a deflection proportional to the mean value of the current passing through it, whereas, in the case of an A.C. sinusoidal quantity, the measurement required is the R.M.S. value which is greater than the mean value and bears a constant ratio to it of 1.11. The 1 mA. D.C. instrument now reads 1.11 mA. R.M.S. A.C. full scale. The scale shape of the D.C. instrument will not be distorted and the frequency error of the rectifier itself can be ignored up to 100,000 cycles per second, so that such a rectifier-type instrument is admirably suited to audio frequency measurements.

For the measurement of higher ranges of alternating current, it will be necessary to use a current transformer, but the design of this is outside the scope of this article.

A.C. Voltmeter

The A.C milliameter may be easily converted into an A.C. voltmeter by the simple addition of suitable resistances, and the circuit arrangement is shown in Fig. 2. These resistances should be of the non-inductive type.

In order to calculate the values of the series resistances to be employed for reading A.C. voltages, it is necessary to know first the voltage drop which occurs across the A.C. terminals of the rectifier, and secondly the current consumption of the complete instrument movement and rectifier at full-scale deflection. The leakage current of the rectifier itself is negligible and hence the current to be passed through the resistance at full scale deflection is 1.11 times the full scale meter current when the waveform is assumed sinusoidal.

The total voltage to be measured at full scale is the sum of the A.C. voltage across the rectifier A.C. terminals and the voltage to be dropped by the series resistance. The value of this resistance is, therefore, easily calculated by subtracting the rectifier voltage drop from the full scale voltage to be measured, and dividing the remainder by 1.11 times the meter full-scale current expressed in amperes. The voltage drop across the rectifier does not change linearly with current, so that if it is an appreciable fraction of the total voltage, it must necessarily distort the scale shape and hence a potential transformer becomes necessary to step up the very low voltages to be measured. For general radio testing, however, a full scale voltage range of less than 10 volts is not likely to be necessary and it is sufficiently accurate to calculate the resistance required for the correct full scale deflection by assuming that the voltage drop across the rectifier and the meter movement at full scale current is 0.9 volt, and across the meter movement 0.1 volt (assuming the use of a 1 mA. D.C. instrument).

For a 10-volt range, therefore, the required resistance will be

$$\frac{10 - 0.8 \times 1{,}000}{1.11} \text{ ohms,}$$

or 8,250 ohms nearest.

For 100 volts full scale, the value of the series resistance is 90,000 ohms and for 500 volts 450,000 ohms.

Very great care must be taken to ensure that the meter movement is connected to the rectifier before any A.C. is applied to the rectifier. Otherwise the rectifier will almost certainly be destroyed due to there being no current path across its D.C. terminals and thus allowing the full circuit voltage to be developed across the rectifier.

Also, it is impossible to incorporate a fuse in a test set in which use is made of a metal rectifier, as blowing of the fuse when it forms part of the D.C circuit will almost certainly result in the rectifier being destroyed, due to the excess voltage.

A suitable design for an A.C./D.C. test set designed on the lines outlined above is illustrated in Fig. 3. It is regretted that an error has occurred in the wiring diagram. The D.C. shunts connected to the negative terminals should have a switch or a push-button control inserted in series where the lead is broken. If a push button is used, it should be arranged to disconnect the shunts when in the " push " position. In this way, a measure of protection is afforded the meter as, no matter what the positions of switches " A " and " B," there will always be some resistance in circuit should either of these switches not be set correctly. Such a meter will be suitable for measuring alternating voltages in mains receivers as well as L.F. voltages in L.F. amplifiers and at the output stage of an amplifier. The meter will have an accuracy of the order of 2 per cent. and be easily read.

The 10-volt range A.C. will be suitable for checking heater circuits, the 100-volt range the heater circuits of universal receivers, and for checking mains and rectifier input voltages, the range of 500 volts is satisfactory.

Output Meters

The frequency error of a rectifier type instrument can be ignored over the whole of the audio-frequency band, and such an instrument can be used for trimming or for measuring the frequency characteristics of a receiver and amplifier.

The most usual requirement is for trimming or ganging with the loudspeaker disconnected and when this is done, it is necessary that the instrument should have incorporated in it a dummy load of suitable impedance to act as a substitute for the loudspeaker. It is also necessary that the instrument shall be choke capacity or else transformer coupled so as to avoid any anode current or even any D.C. voltage drop arising from anode current causing a reading on the meter.

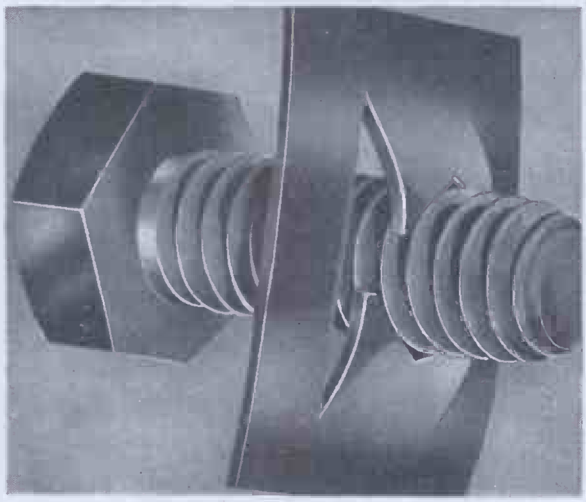
The 100-volt A.C. range should be used connected through a large capacity paper condenser in paralled with the loudspeaker. If the speaker itself is to be disconnected for purposes of quietness, a load of the same impedance must be connected across the voltmeter terminals.

The necessary circuit arrangements are shown in Fig. 4, where the speaker is disconnected and a suitable load resistance used.

This circuit is quite safe as regards avoiding risk of damage to the rectifier due to switching-on surges, so long as the output valve is of the indirectly heated type and the H.T. supply is not switched on by means of a delayed action switch. Where it is possible for

(*Continued on page* 144)

New Technique in Rapid Assembly



This photograph shows clearly the operating principle of the speed nut. Tension is always present on the screw irrespective of shrinkage of material.

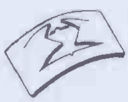
THE advantages of present day mass production methods are apparent to everyone these days, enabling the manufacturer to reduce the selling price of his product, thereby ensuring a larger consumer potential, which in turn permits of the purchaser obtaining the maximum quality for a minimum of outlay.

In any product, in the construction of which a large number of separate parts are used which must be secured by screws and nuts, it is obvious that a large amount of time must be taken in assembly if ordinary nuts are used. Further, with certain materials there is always the possibility of contraction taking place after a period of time and the fastening becoming loose. The ideal, therefore, would be a nut which permits rapid assembly and also remains tight even though contraction or expansion of the assembly takes place owing to local temperature conditions.

An ingenious nut has been devised, known as the Simmonds' Speed Nut, which fulfils these conditions. The speed nut is a strip of metal of any shape or size, having an impression in it which forms two opposed prongs, the ends of which form a diamond shape hole, slightly less in dimension than the root diameter of the screw. These prongs are formed in such a manner so as to engage the thread of a screw, whether it has a machine, wood or metal thread. The base of the speed nut is arched and the final tightening of the screw into the speed nut causes the initial arch to flatten out, with the result that the two prongs tend to close together, thus creating a gripping or clamping action to which the threads merely act as guides. It can be appreciated that should any additional force be applied to the initial stress, then the prongs will clamp even tighter to the root of the thread.

This, of course, is in direct contrast to the behaviour of an ordinary nut, which, when tightened creates a vertical force against the threads. No amount of vibration will cause the speed nut to loosen, and the springy nature of the material gives an ample margin of retaining power fully equal to that of the screw upon which it engages; the speed nut is, of course, readily removable without any resultant damage to the thread.

For heavier duty there are speed nuts having one end of the strip bent back upon itself, so that it engages the threaded portion of the screw at a second point. Speed nuts having the shape of a "U" are also available for blind assembly applications where an ordinary nut would be inaccessible. Actually, there are over 500 different types available, and new variations are being constantly added to meet special

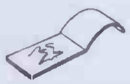


A speed nut is used in conjunction with any threaded member, *e.g.*, bolt, machine screw, wood screw, etc. It can be speedily zipped into position and replaces lock washer and nut.

A speed fix is used in conjunction with any unthreaded member, *e.g.*, rivets, studs, etc. It is pushed on into its final locked position, replacing threaded member, lock washer and nut.

Speed clip for fastening integral studs or rivets where the back of the assembly is inaccessible. It is snapped into a hole and the stud is pushed into the clip.

requirements. To mention only three, the right-angle speed nut suitable for retaining a glass panel on a radio set (this single speed nut dispenses with the need for three separate components), a



Speed nut with integral cable clip.

simple speed nut for torch lamp assembly, and a speed nut with its end formed as a cleat for clamping cables, this latter design being particularly suitable for radio construction.

In addition to the speed nut, there are two other devices—the speed fix and the speed clip—which operate on similar principles, but have their own particular applications.

The speed fix is for use with a non-threaded pin or stud. It is similar in design to the speed nut and operates in much the same way; it may be positioned by the use of a suitable tool. As the arch springs back, the angular side of the prongs clamp against the shank of the pin or stud. The angle at which the prongs are designed causes them to grip still harder should the pins tend to be withdrawn. The resultant fastening is very secure and can be relied upon for many purposes where light riveting would previously have been used. This method of fastening also permits of the use of integrally formed studs, thus dispensing with threaded inserts.

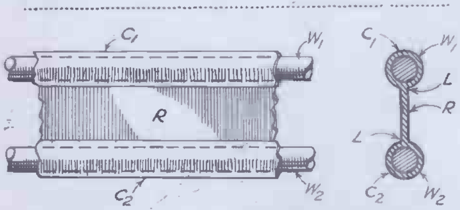
The third member of the system, the speed clip, comprises a variety of fastening arrangements not possible by the speed nut or speed fix. These clips are largely used in the radio industry for knob to shaft, blind location, and cable cleat assemblies, etc., and are particularly efficient for use with plastic materials, obviating the need for pre-set metal components in plastic ware.

As these fixing devices are produced by mass production methods from strip material they are comparatively inexpensive and are obtainable in four materials, i.e., cold rolled steel, stainless steel, spring steel and phosphor bronze, and in addition various plated finishes are available.

These products are manufactured by Simmonds Aerocessories, Limited, at Great West Road, Brentford, London.

A Flexible Low-attenuation Transmission Line

THE properties required of a high-frequency transmission line are usually both of a mechanical and an electrical nature. Thus the line must usually possess flexibility and resistance to heat and moisture as well as a low power factor and dielectric constant. Difficulties have, however, always been



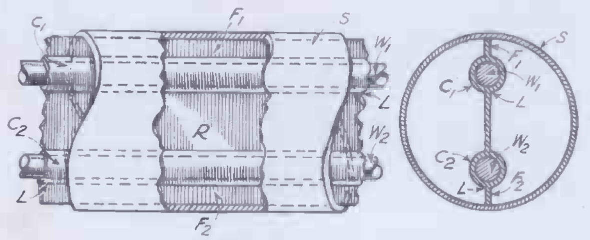
Figs. 1. and 2. Part section and cross section of simple unshielded line

experienced in satisfying all these requirements. Embedment of the conductors in rubber or rubber-like materials has invariably given rise to high power factors, and high capacities, while replacement of the embedding material by thermoplastic materials of better electrical properties has resulted in little success because of inflexibility. This has led to the almost universal use of various forms of discrete supports, such as insulating discs, mechanical spiders, etc., spaced along the conductors for maintaining them in a desired relation, while permitting the requisite degree of flexibility in the line. Such spaced supports are, however, expensive, particularly in their installation, where difficulties may arise not only in attaching them to the conductors, but also in fitting them within the outer conductor maintaining them in upright alignment.

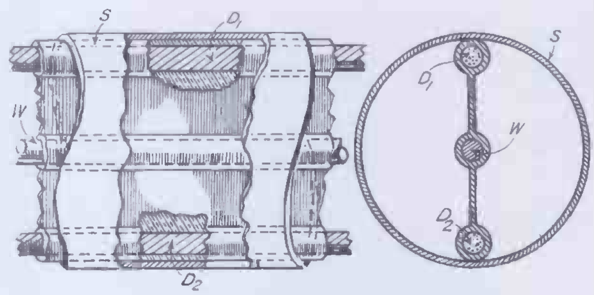
A new proposal, emanating from the R.C.A. Laboratories, is to use a polystyrene embedment, or an embedment of an equivalent material, which may be moulded in a continuous process without the use of plasticizing agents, this embedment being formed as a ribbon-like support having a thickness no greater than the embedded wires. The ribbon form makes for flexibility, and this property is enhanced by providing that the surfaces of the ribbon proper intersect the coated surfaces of the wires at a sharp angle.

An arrangement as described is illustrated in its simplest form in

Figs. 1 and 2, where W1 and W2 are the parallel wires of an unshielded transmission line and R is the embedding ribbon. C1 and C2 indicate the regions where the insulation coats the wires, and at regions such as L there is shown



Figs. 3. and 4. Sheathed line of similar construction.



Figs. 5. and 6. Sheathed concentric line assembly.

the manner in which the ribbon and coating surfaces intersect at a sharp angle.

The arrangement of Figs. 3 and 4 illustrates a case where the transmission line is sheathed. Here the ribbon extends cross-sectionally beyond the wires, as shown at F1 and F2. Such an arrangement serves to centre the line within the shield S and also to prevent short-circuiting should there be any break in the continuity of the coatings C1, C2.

Figs. 5 and 6 show a concentric line arrangement. The single central wire W is, of course, embedded and cords D1, D2, of cotton or similar material, are also embedded along the edges of the ribbon as illustrated, so as to support the central conductor in correct relation to the outer conductor S.

In moulding transmission lines of the type described above, either liquid styrol or granulated polystyrene may be used. When styrol is used it may be polymerized, *e.g.*, by means of a high-powered source of ultra-violet light mounted close to the point where the ribbon leaves the mould. When granulated polystyrene is used, it may be made plastic by heating the injector or the mould or both.

When the wires or cables are of large diameter, the thickness of the ribbon may be much less than the diameter of the wire. Thus in one practical example, using 14-gauge (B. and S.) cable, the polystyrene ribbon is .020 in. to .030 in. thick. The same ratio of ribbon thickness to wire gauge cannot usually be maintained, however, with very fine wires. Thus when 36-gauge wire is used, the ribbon thickness should be about the same as the wire diameter.

Where even higher flexibility is required, certain vinyl resin compounds (*e.g.*, vinyl chloride) may be substituted for polystyrene; but this can only be done at the expense of the electrical characteristics of the line, such resins having a higher dielectric constant than polystyrene.

A Roof-spotter Alarm System

A ROOF-SPOTTER alarm system has been developed by Londex Limited, of Anerley Works, London, which, by reason of its unit assembly, is suitable for either small or large premises—that is it can be arranged to provide a warning at a single point or at a considerable number situated over a large area.

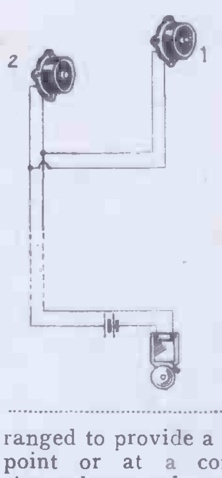


Fig. 1. Simple arrangement of roof-spotter alarm system, using a special delayed action push-button 1, and ordinary push-button 2.

The system is shown in its simplest form in Fig. 1, and it will be seen to consist of a delayed-action push-button mounted within easy access of the look-out man on the roof. When this push-button is pressed, an alarm circuit is closed for a pre-determined time, say 10 seconds to 2 minutes, and therefore it is not necessary for the look-out man to remain exposed to danger. Push-button 2 is without delayed action and as soon as the raiders have passed the spotter presses this and gives, for instance, three short signals indicating that immediate danger is over.

Fig. 2 shows the fundamentals of an extension of the system which provides for a number of alarm points at remote parts. There are two push-buttons in this case also, but the current does not actuate the alarm devices direct. The two bush-buttons are used to control small Londex relays which are installed in all the places where the alarm has to be given. Though each relay coil needs only a few milliamps, they are capable of controlling sirens and hooters with up to 50 amps consumption.

Both the push-button circuit and the alarm devices can be fed from either the mains or an auxiliary battery, the latter, of course, rendering the whole system independent of mains failure.

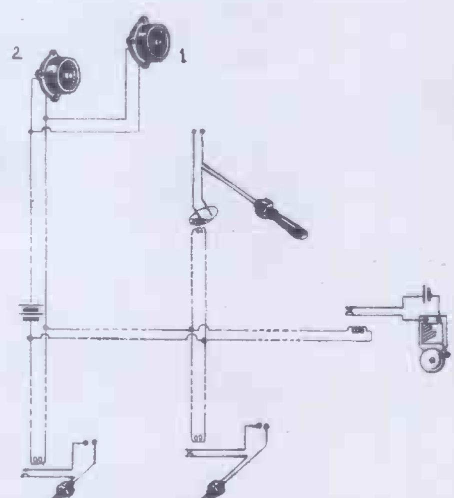


Fig. 2. Roof-spotter alarm system using a special delayed action push-button and ordinary push-button together with relays for signalling to several remote parts.

Notes on the Design of Vibrator-operated Power Supplies

By S. L. Robinson

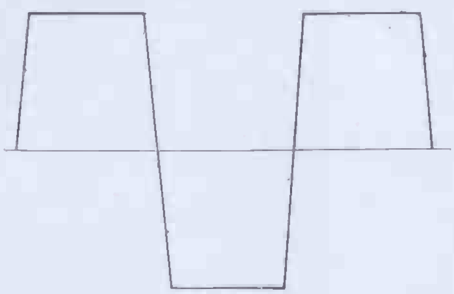
THE use of vibrator-operated power supply units is becoming increasingly popular, and it is obviously an advantage in war time to be independant of dry batteries and the mains. This article describes briefly some of the uses and applications of vibrator packs.

A vibrator pack is a unit used for converting low voltage D.C. supply, generally either 6 or 12 volts, to high tension suitable for anode supply. It consists of a vibrator, a transformer and a rectifying system, the output being rectified unsmoothed D.C. The vibrator, the heart of the vibrator pack, is a mechanical device which breaks up and reverses the D.C. supply, making it suitable for feeding into the primary of a step-up transformer. This vibrator is the only part of the unit which is subject to wear, but replacement can be readily carried out as a vibrator employs a similar method of connection as a valve. The life of a vibrator depends wholly on the associated circuit with which it is used—with a correctly designed circuit its "life" can be expected to be well in excess of 1,000 hours.

The output from the transformer, which can be any desired voltage, is rectified either by a valve or by the secondary contacts on a "self-rectifying" type of vibrator.

In the design of vibrator-operated power supplies engineers are faced with a larger number of factors and problems than are encountered in the design of an equivalent A.C. power supply. No one component in the supply can be divorced from the rest, since its function and design depends upon the design and operation of the other components. Therefore, the vibrator, transformer, tuning capacity (commonly called buffer condenser), battery voltage, battery supply lead resistance, H.T. output required, all must be considered when the power supply is to be designed. Knowing the nominal battery voltage, i.e., 4, 6, 12 or 32 volts, the approximate lead, switch, fuse, and L.T. choke resistances, the required H.T. output, and the variation in battery voltage encountered in service, the problem resolves itself into correlating the three important items of the supply, namely, the vibrator, transformer and buffer condenser.

In the design of A.C. power supplies the designer considers mainly, losses,



Correct waveform of vibrator output.

regulation, and output. All these factors must be considered in the design of a vibrator-operated power supply, and in addition, size, and primary current drain are of importance. Size because this type of supply is often used with portable apparatus, and primary current drain because of the limitation of battery drain. Furthermore, it must be appreciated that vibrator life is largely determined by the loading applied to the contacts.

Since it is necessary to operate this type of power supply in a multitude of apparatus designs, having varying values of L.T. lead resistance, it is customary to rate the vibrator pack as furnishing the required output at an input voltage of 4, 6, 12 or 32 volts as measured from the centre tap of the transformer primary to the reed terminal on the vibrator holder. When this is not done it is necessary to specify that the "centre tap voltage" be stated at which the required output is to be secured. Since it is also necessary to operate the vibrator pack from a battery the state of charge of which is variable, it is necessary to design the pack components so that they will perform safely within a fairly wide range of input voltages, say from 5.75 volts to 9 volts in the case of a 6-volt accumulator, which is being charged at the same time as it is being used, as, for example, in the case of a car radio receiver.

Voltages at the centre tap will vary considerably, depending upon whether a "starting" or "running" condition is being considered. In the case of a 6-volt vibrator pack this may range from 5.25 volts to 8.5 volts or a 62 per cent. variation. This compares with the much lower percentage variation encountered in A.C. mains. In addition, since indirectly-heated valves are now used for practically all applications involving vibrator-operated power supplies, "no load" condition for the supply is present every time the apparatus is switched on, and this must be fully considered in the design because a vibrator is not only a mechanical device, but is limited by transients.

Reducing Interference

A certain amount of R.F. interference or "hash" is produced by the Vibrator contacts and precautions have to be taken to prevent this interference from causing trouble in the associated equipment. The amount of suppression required depends on the equipment to be used, and it can be readily seen that the design of a vibrator pack as H.T. supply for a communication receiver is very different to that for a transmitter, or an L.F. amplifier.

The standard vibrator pack includes suppression which is ample for normal purposes, such as broadcast receivers, L.F. amplifiers, and transmitters, but for a communication receiver with a sensitivity of the order of one microvolt, special precautions have to be taken, especially if the receiver is to cover a wide band of frequencies; this method is dealt with more fully later.

First, consider the case of a vibrator pack being used with an L.F. amplifier. As only low-frequency amplification is concerned, special precautions are not required as regards "hash" suppression, and the amount of L.F. smoothing required is less than that required with a 50-cycle mains supply because of the much higher frequency of the vibrator-generated A.C. Great care should be taken, however, when designing the filter circuit for use with an amplifier employing either class "AB" or "B" to allow for the swinging load of the output valves.

PREMIER RADIO

PREMIER SMOOTHING CHOKES

Type	Current	Henrys	Res	Prices
C 40/500	40 MA	20-34 H	500 ohms	5/3
C 60/180	60 MA	8 H	180 ohms	5/3
C 60/400	60 MA	25-34 H	400 ohms	7/6
C 60/500	60 MA	18-30 H	500 ohms	5/3
C 100/400	100 MA	20-34 H	400 ohms	9/6
C 150/185	150 MA	20-34 H	185 ohms	13/6
C 200/145	200 MA	20-34 H	145 ohms	15/9
C 250/120	250 MA	25 H	120 ohms	17/6
C 60/2500	60 Speaker Field Replacement			7/6

Premier Pick-ups.

Heads (Will fit any tone-arm) 7/6
Premier Pick-up with Volume Control 2,000 ohms 12/6
Premier de Luxe, 6,000 ohms 17/6

ANOTHER SPECIAL OFFER.
Rothermel Brush Piezo Crystal Pick-ups. Standard, S.8 Model with arm, 34/9. P.U. head only, De Luxe Model, 24/6. Heads for Garrard or Calloro Autochangers, 34/6.

PREMIER 1941 HIGH FIDELITY AMPLIFIER KITS

Each Kit is complete with ready drilled chassis, selected components, specially matched valves and full diagrams and instructions.

	Kit of Parts with Valves	Completely Wired and Tested
4-watt A.C. Amplifier	£2 14 0	£3 11 6
4-watt A.C./D.C. ,,	£3 0 0	£3 17 6
6-watt A.C. ,,	£6 16 6	£7 13 6
8-10-watt A.C./D.C.	£6 11 6	£7 9 0
15-watt A.C. ,,	£7 18 9	£9 8 0

Black Crackle Steel Cabinet 17/6 extra.

MAINS TRANSFORMERS

Wire-ends. All L.T. Windings Centre-Tapped

SP. 250	250-0-250 v. 60 m.a. 4 v. 1-2 a, 4 v., 2-3 a., 4 v. 2-3 a. ...	11/9
SP. 300	300-0-300 v. 60 m.a., 4 v. 2-3 a, 4 v., 2-3 a., 4 v. 2-3 a. ...	11/9
SP. 301	300-300 v. 150 m.a. 4 v. 2-3a, 4 v. 2-3 a, 4 v. 1 a, 4 v. 1a ...	15/-
SP. 350A	350-350 v. 100 m.a. 5 v. 2a (not C.T.), 6.3 v. 2-3 a	14/-
SP. 350B	350-350 v. 100 m.a. 4 v. 2-3a, 4 v. 2-3 a, 4 v. 2-3 a...	14/-
SP. 351	350-350 v. 150 m.a. 4 v. 1-2a, 4 v. 2-3 a, 4 v. 3-4 a...	15/-
SP. 352	350-350 v. 150 m.a. 5 v. 2a, 6.3 v. 2 a, 6.3 v. 2 a...	15/9

Auto Transformers. Step up or down. 100-125 v. to 200, 230 or 250 v. A.C., 60 watts, 9/11; 125 watts 13/6; 250 watts, 18/6.

L.T. Transformers, all C.T.

4 v. 2-3 a.	... 9/11	6.3 v. 2-3 a. ...	9/11
2.5 v. 5 a.	... 9/11	7.5 v. 3 a. ...	9/11
5 v. 2-3 a.	... 9/11	12 v. 3-4 a. ...	15/-

Push-Pull Driver Transformers, 3 : 1 6/6
Universal Output Transformers. 11 Ratios. Single or Push-Pull ... 6/6
Bell Transformers, 3-5 or 8 volts ... 5/6

MATCHMAKER UNIVERSAL OUTPUT TRANSFORMERS

Will match any output valves to any speaker impedance.
11 ratios from 13 : 1 to 80 : 1, 5-7 watts, 15/9. 10-15 watts, 20/6. 20-30 watts, 35/-.
Potentiometers, all resistances, 2[4 each; switch 3/3.
Ceramic Valve Holders, chassis mounting, 4 or 5 pin, 7d. each. 7 pin, 11d. each.
Valve Screens, for American type valves. Small size, 1/2. Medium size, 1/2.

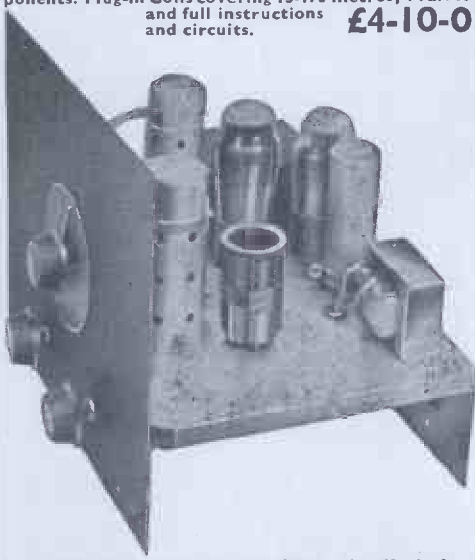
"LEARNING MORSE"

Premier Morse Practice Key on Bakelite Base and Brass Movement	3/3
General Purpose Morse Key ...	5/10
Heavy Duty TX Key on Cast Base ...	10/-
Bakelite Buzzers	2/-
3 Henry Chokes	7/6
Complete Kit of Parts for Valve Oscillator as described in W.W. "Learning Morse"	25/-

NEW PREMIER S.W. A.C. RECEIVER

In response to many requests we have now produced an A.C. version of the popular Premier Short Wave SG3 Kit. Circuit: Pentode H.F. Stage, Pentode Detector, Beam Power Output, and F.W. Rectifier. 200-250 v. A.C. Operation. Built-in Power Pack. Hum-free operation. For use with Phones or P.M. Speaker.

Complete Kit of Parts with drilled chassis, all components. Plug-in Coils covering 13-170 metres, 4 valves and full instructions and circuits. £4-10-0



Battery Version also available Kit £3 8 0
Extra Coils 9—15, 200—2000m. also supplied

★ *"The Wireless World" said they were* **"very much impressed . . . . ."** ★

See full Test Report pp. 492-3 December issue
Send for full details

PREMIER BATTERY CHARGERS for A.C. MAINS

Westinghouse Rectification complete and ready for use

To Charge :	6 volts at 1 amp. 22/6
2 volts at ½ amp. 11/9	12 volts at 1 amp. 24/6
6 volts at ½ amp. 19/-	6 volts at 2 amps. 37/6

MOVING COIL SPEAKERS

All complete with transformer. Rola 6½ in., 15/-; 8 in. P.M.s, 17/6; 10 in. P.M.s, 22/6.

ENERGISED MODELS

Plessey 8 in., 175 ohm field, 7/6; G.12 energised, or 2,500 field, 63/-. 10 in. B.T.H. 1,600 ohm field, less transformer, 11/6. Magnavox 154, 2,000 ohms Universal transformer, 15/-.

PREMIER SHORT-WAVE KITS for OVERSEAS NEWS

Incorporating the Premier 3-Band S.W. Coil. 11-86 Metres without coil changing. Each Kit is complete with all components, diagrams and 2-volt valves. 3-Band S.W. 1-Valve Kit, 14/9. 3-Band S.W. 2-Valve Kit, 22/6.

DE LUXE S.W. KITS

Complete Kit, including all Valves, coils, wiring diagrams and lucid instructions for building and working. Each Kit supplied with a steel Chassis, Panel and plug-in coils to tune from 13 to 170 metres.

1 Valve Short-Wave Receiver or Adaptor Kit	20/-
1 Valve Short-Wave Superhet Converter Kit	23/-
1 Valve Short-Wave A.C. Superhet Converter Kit	26/3
2 Valve Short-Wave Receiver Kit ...	29/-

SHORT-WAVE CONDENSERS

Trolitul Insulation. Certified superior to ceramic. All-brass construction. Easily ganged

15 m.mfd. ...	1/9	100 m.mfd. ...	2/3
25 m.mfd. ...	2/-	160 m.mfd. ...	2/6
40 m.mfd. ...	2/-	250 m.mfd. ...	2/11

SHORT-WAVE GEAR

Short-Wave Coils, 4- and 6-pin types; 13-26, 22-47, 41-94, 78-170 metres, 2/- each, with circuit. Premier 3-Band S.W. Coil, 11-25, 19-43, 38-86 metres. Suitable any type circuit, 2/11.

4-pin or 6-pin Coil Formers. Plain or Threaded, 1/2 each.

Utility Micro Cursor Dials, Direct and 100 : 1 Ratios, 4/3.

Bakelite Dielectric Variable Condensers. .0005 mf. Suitable Tuning or Reaction, 1/6 each.

Short-Wave H.F. Chokes. 10-100 m., 10½d. each. High grade Pie-Wound U.S.A. type, 1/9 each.

Lissen Dual Range Screened Coils. Medium and Long Waves, 2/9 each.

REPLACEMENT VALVES

FOR ALL SETS

Europa Mains Valves. 4 v. A.C. Types A.C./H.L., A.C./L., A.C./S.G., A.C./V.M.S.G., A.C./H.P., A.C./V.H.P., A.C.P., all 5/6 each. A.C./H.P., 8/6. A.C./Pens, I.H., 10/6; A.C./P.X.4, 9/-; Oct. Freq. Changers, 9/6; Double Diode Triodes, 9/9; 3½-watt D.H. Triode, 12/-. 350 v. F.W. Rect., 10/8; 13 v. .2 amp. Gen. Purpose Triodes, 5/6; H.F. Pens. and Var.-Mu. H.F. Pen., Double Diode Triodes, Oct. Freq. Changers, 8/6 each. Full and Half-wave Rectifiers, 9/8 each. D.D. Pens. 10/3

FULL RANGE OF BATTERY VALVES IN STOCK. WRITE FOR LATEST LISTS.

Cardboard Electrolytic 500 v. Conds. 4 MFD, 8 MFD, 1/9; 8×8, 3/6; 4×4, 2/8; 8×4, 3/3; 4×4×1, 3/9; 4×4×4, 4/3; 16×8, 4/-; 16×16, 5/-.

PREMIER MICROPHONES

Transverse Current Mike. High grade large output unit. Response 45-7,500 cycles. Low hiss level, 23/-.

Moving Coil Mike. Permanent magnet model requiring no energising. Response 90-5,200 cycles. Output .25 volt average. Excellent reproduction of speech and music. 49/-.

Microphone Transformers. Suitable for all mikes. Tapped secondaries. A, 20 and 40 : 1; B, 30 and 60 : 1; C, 50 and 100 : 1. 6/6 each.

Microphone Stands. Bakelite table stand, 9 in. ring, 7/6 each.

ALL ENQUIRIES MUST BE ACCOMPANIED BY 2½d. STAMP.

PREMIER RADIO CO.

ALL POST ORDERS to : JUBILEE WORKS, 167, LOWER CLAPTON ROAD, LONDON, E.5 (*Amherst* 4723).

CALLERS to : Jubilee Works, or 169, Fleet Street, E.C.4 (*Central* 2833), or 50, High Street, Clapham, S.W.4 (*Macaulay* 2381).

It is worth mentioning at this stage that vibrator packs can be obtained giving quite large outputs. Masteradio, Ltd., manufacture units which operate from 6 or 12 volts; in the case of the 6-volt types, outputs up to 30 watts can be obtained. The normal ratings of this type are 300 volts at 100 mA., but higher voltages can be obtained with an appropriate reduction in milliamps. In the 12-volt type, outputs up to 60 watts can be obtained at any desired voltage. The overall efficiency of a well-designed 30-watt vibrator pack is more than 70 per cent., and in the case of the 60 watt type is in excess of 80 per cent., the consumption being 6 amps at 12 volts, while the output could be 400 volts at 150 mA.

Next consider the use of vibrator packs for transmitters. There are many cases where portable power is required for transmitters such as field equipment and for vehicles. For low power a single vibrator pack would be sufficient to supply all the stages necessary, but where maximum output is required, a number of vibrator packs can be used, one for each stage. Again, as in the case of amplifiers, no precautions need be taken about R.F. "hash" and in some cases where direct radiation does not matter, vibrator packs can be used completely without any R.F. filter. No difficulty will be experienced with keying in any usual manner.

Vibrator Packs for Communication Receivers

The design of vibrator packs for supplying H.T. to a communication receiver with high sensitivity, and covering a multiplicity of wave bands has been specialised in by Masteradio, Ltd., for use by the Services, and much development has been carried out which has proved that any type of receiver can be operated from a vibrator pack H.T. unit, without the slightest trace of interference from the power pack.

The suppression necessary on a vibrator pack for use with a communication receiver is, however, more complicated than with an amplifier, owing to the R.F. "hash" generated by the vibrator being picked up in the aerial, R.F., and detector stages, and being amplified along with the signal through the various stages.

The suppression of "hash" on any given frequency is a comparatively simple matter, by using tuned filter circuits, but this method becomes impracticable when a wide band of frequencies has to be covered by the receiver, and other methods must be employed. This calls for the use of special components, such as filter chokes of extremely low self capacity, and by-pass condensers of very low-loss dielectric. Quite often as many as three filters, all completely different, are required in series with each other, while condensers across the supply can be as large a capacity as 2,000 μF.

Also extremely important is the design of the vibrator transformer, and taking into account the 115 cycle square type of wave form (*see* diagram, the transformer, and associated components must reflect a pure resistive load on the vibrator, and provide the necessary power factor correction to ensure efficient conversion. Transformers designed for normal power use satisfy none of these requirements. It must also be remembered that layout of components, screening and the earthing play just as important a part as in the design of a receiver.

Finally, to sum up the position, it is obvious that the design of vibrator power supply units is a specialised subject, and few hard and fast rules can be laid down, as every case merits individual consideration.

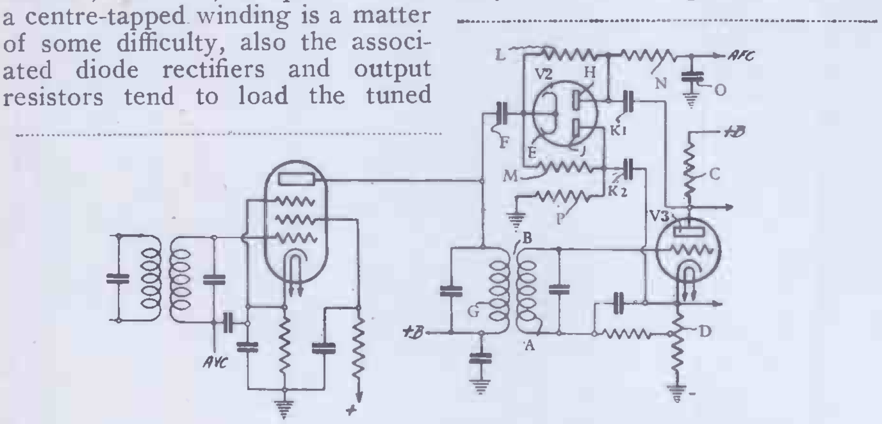
Major Edwin H. Armstrong has been awarded the Holley Medal of the American Society of Mechanical Engineers for his work on frequency modulation. Major Armstrong is professor of Electrical Engineering at Columbia University.

Improved Discriminator Circuit for Automatic Frequency Control

IN a very common form of discriminator circuit tuned transformer coupling is employed, one side of the primary winding being connected to a centre tap on the secondary and the secondary winding feeding a pair of rectifying diodes in phase opposition. In production, however, the provision of a centre-tapped winding is a matter of some difficulty, also the associated diode rectifiers and output resistors tend to load the tuned circuit of a discriminator network, thus reducing the gain and selectivity of the signal transmission circuit.

In a circuit arrangement recently devised by the Radio Corporation of America, these difficulties are overcome by means of a further valve arranged to provide an intermediate frequency output in push-pull.

Referring to Fig. 1, it will be seen that this tuned secondary winding A of the I.F. transformer B is connected between the grid and cathode of the amplifier valve V3, which has equal impedances C and D arranged in its anode and grid circuits respectively. In operation, therefore, intermediate frequency signals in push-pull are set up across these impedances and are used in place of the signals provided in the conventional arrangements by the centre-tapped secondary. It will be appreciated that in the new arrangement, the tuned secondary winding is not damped by the diodes of the discriminator and no centre tap is required.



The two discriminator diodes have a common cathode E connected via the blocking condenser F to the high-voltage side of the tuned primary winding G of the I.F. transformer and have anodes H, J which are connected via blocking condensers K1, K2 respectively to the anode and cathode of the valve. Rectified voltages are set up across the loads L and M associated with the two rectifiers, and the sum of these voltages, which is the required A.F.C. voltage, is developed at the anode and is led off through the filter N, O. The anode is earthed through the leak resistance P.

As shown in Fig. 2, the intermediate frequency signals developed across the impedances C, D may be rectified by the diodes O, R to provide a rectified signal output in the load S for application to the low-frequency amplifier of the receiver.

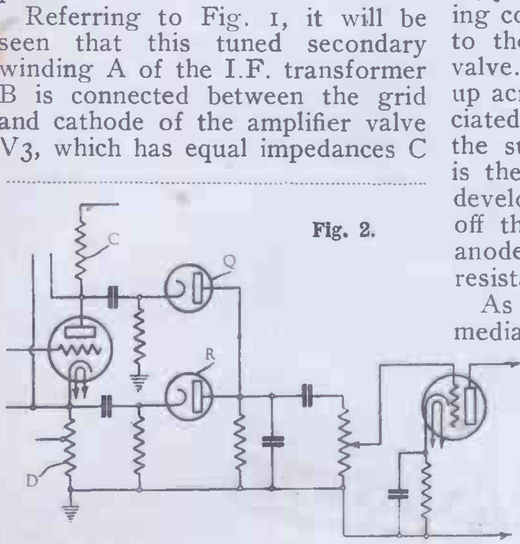


Fig. 2.

The Short-wave Radio World

A Review of the Most Important Features of the World's Short-wave Developments

THE Wien bridge, though old, has remained in the laboratory until recently. Lately, however, there have been several commercial applications.

The fundamental circuit, shown in Fig. 1, is quite simple, and with some modifications can be applied to such uses as heterodyne reduction, oscillator frequency selection, sound analyzing, frequency measurement, capacity measurement, and condenser power-factor measurement. The following notes are concerned with its application in simple audio-oscillator circuits, of the type suitable for speech amplifier testing and similar uses. Without going into the operation of the bridge, it may be said that it is a selective network with a sharp selectivity characteristic, Fig. 2 being typical.

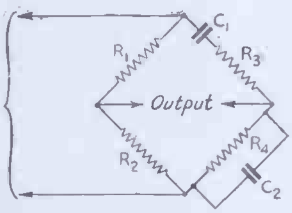


Fig. 1. The fundamental Wien bridge circuit. Frequencies for which the bridge is balanced do not appear in the output circuit.

In many applications of the bridge circuit it is necessary to use a transformer to get balance to earth. The circuit in Fig. 3, the equivalent parallel-T network, has a common earth connection for input and output, and is therefore frequently more convenient to use. The bridge shown in Fig. 1 can be tuned by varying the two resistances R_3 and R_4 simultaneously, or by varying C_1 and C_2 simultaneously. The equivalent parallel-T network can be tuned with three ganged resistances, R_5, R_6 and R_7, or three ganged condensers, C_3, C_4 and C_5. If, in the circuit of Fig. 1, $R_1=2R_2$, $C_1=C_2$, and the ganged resistances R_3 and R_4 have the same value, the simple equation defining the frequency of the null point is:

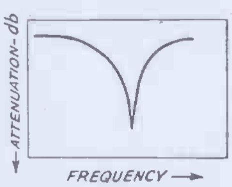


Fig. 2. A typical selectivity characteristic of the Wien bridge. This characteristic is basic in audio oscillators using the circuit.

$$f = \frac{1}{2\pi R_3 C_1}$$

In the parallel-T network of Fig. 3, if $C_3=C_4=\frac{1}{2}C_5$ and $R_5=R_6=2R_7$, the equation is:

$$f = \frac{1}{2\pi R_5 C_3}$$

Audio oscillators using the Wien bridge or its equivalent parallel-T network as a selective feed-back network have recently been introduced commercially. Essentially such an oscillator is an amplifier with both positive and negative feed-back. The positive feed-back occurs at all frequencies while the degenerative feed-back just cancels the positive feed-back at all frequencies except the frequency for which the network is tuned. By varying this null point, the frequency of oscillation is varied. An audio oscillator using the Wien bridge or its equivalent-T for the denegerative feed-back has many advantages over the conventional heterodyne and LC audio oscillators. The heterodyne oscillator is a complicated affair with many valves and circuits; while the LC oscillator requires an iron-core coil and large condensers to tune to audio frequencies, with the result that the frequency is usually varied in steps rather than continuously. The Wien bridge oscillator requires only one valve, and can be tuned continuously by resistances; in addition, it gives practically harmonic-free output because of the sharp characteristic curve of the degenerative network.

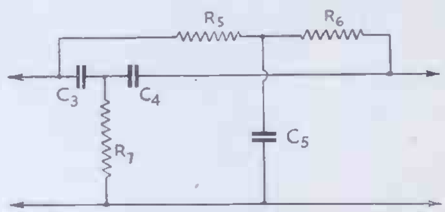


Fig. 3. The equivalent parallel-T network, for use with earthed input and output circuits.

Fig. 4-A is a simplified oscillator of this type, using the parallel-T network. The Wien bridge and a transformer are used as the selective network in the oscillator shown in Fig. 4-B. In both circuits the positive feed-back is obtained through the 1:3 transformer, T_1, with the low-impedance side in the grid circuit. The amount of regeneration is controlled by the cathode resistance R_9. A by-pass condenser across R_9 is apt to cause the oscillator to produce all sorts of gurgling sounds, and therefore should

not be used. The oscillator can be tuned by either of the methods previously described. If the feed-back is too great a resistance, R_8, of the order of 500 ohms, will have to be put in series with the grid lead.

The circuit of Fig. 4-A has the advantage of requiring only one cheap transformer, but has the disadvantage that a three-gang resistance that stays ganged is needed. Condenser tuning might be used instead, or a number of feed-back circuits could be switched in or out for different fixed frequencies. The circuit of Fig. 4-B has the advantage of being easily tuned by a two-gang resistance, but has the disadvantage of requiring two transformers.

It would be possible to obtain the 180-degree phase shift necessary for degeneration by using a voltage from the anode circuit of the second section of a double-triode valve. This would make possible the elimination of a transformer, thereby cutting cost and reducing size. However, in this connection it must be pointed out that the two feedback voltages applied to the grid must be exactly 180 degrees out of phase. This condition is hard to fulfil when more than one valve is used.

In Fig. 4-A the resistance of the potentiometer across the output should be high so that the degenerative voltage is unaffected. The resistances in the degenerative network should be fairly high

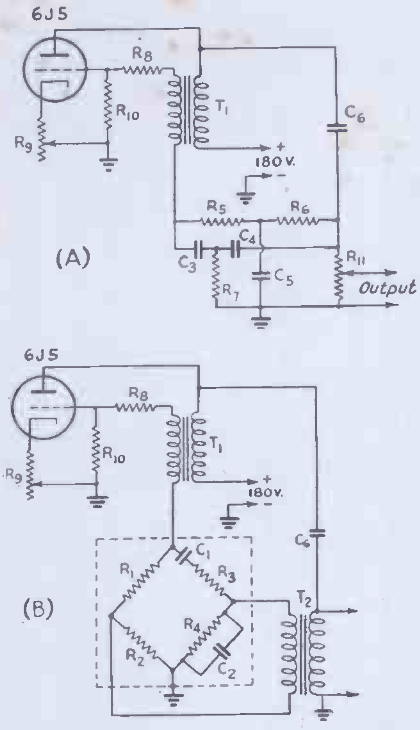


Fig. 4. Simple bridge audio oscillator circuits. A—using the parallel-T network B—using the bridge.

C_1, C_2—0.15 μfd.
C_3 C_4—0.15 μfd.
C_5—0.3 μfd.
C_6—0.1 μfd. or larger.

R_1—4000 ohms.
R_2—2000 ohms.
R_3, R_4—10,000-ohm potentiometers.
R_5, R_6—10,000 ohms.
R_7—500 ohms.
R_8—500 ohms.
R_9—5000-ohm variable.
R_{10}—0.25 megohm.
R_{11}—0.6-megohm potentiometer.
T_1—3 : 1 audio transformer.
T_2—500-2000-ohm transformer.

in value so that a substantial degenerative voltage can be developed at low current.

The cathode-resistance control should be advanced only far enough to set up reliable oscillation. The anode-supply voltage can be varied from about 150 volts to about 450 volts without causing a noticeable shift in oscillator frequency.

At medium and low output, the harmonic distortion is so small as to be negligible. However, at high output a small second harmonic is discernible. The output volume is too great for comfort when using a headset, but if greater volume is needed for loudspeaker operation it is advisable to add an amplifier rather than to try to get more output directly from the oscillator. This will keep the harmonic content to a minimum.—**Q.S.T.** (January).

Acknowledgment to " Q.S.T." for the article " A Stabilised 2½-metre Oscillator " which appeared on page 94 of the February issue was accidentally omitted.

"Simple Meter Conversions"

(*Continued from page* 132)

the H.T. circuit to be made or broken in this way, or where a directly heated valve is in use, it is necessary that the instrument rectifier circuit is not connected across the transformer until after the anode current is flowing and disconnected before the H.T. circuit is broken.

With a load resistance of 10,000 ohms (a common value for a pentode output valve) the full scale deflection of 100 volts will correspond to an L.F. output of 1 watt. Using the 500-volt range, the corresponding full scale reading is 25 watts.

For valves with a matching impedance other than 10,000 ohms, the load resistance may readily be changed to

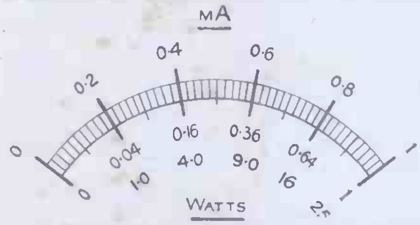


Fig. 5. 100v. and 500v. A.C. ranges in terms of watts.

suit, the value used being obtained from the valve maker's data sheets. The watts output of a valve may be calculated from the formula $\frac{V^2}{R}$ where V is the voltmeter reading and R the value of the load resistance in ohms.

With a load resistance of 5,000 ohms for example, a reading of 50 volts corresponds to ½-watt. Such an output meter may be made up with a series of resistances to suit the valves most commonly in use and the scale calibrated to read direct in watts from the formula $\frac{V^2}{R}$.

Meter reading mA.	Load resistance values 10,000	7,500	6,000	5,000	4,000	3,000	2,000
	Watts output						
0.2	0.04	0.053	0.07	0.08	0.1	0.14	0.2
0.4	0.16	0.213	0.27	0.32	0.4	0.54	0.8
0.6	0.36	0.48	0.6	0.72	0.9	1.2	1.8
0.8	0.64	0.85	1.07	1.28	1.6	2.14	3.2
1.0	1.0	1.33	1.7	2.0	2.5	3.4	5.0

The scale is of square law shape and a typical scale for 100-volt and 500-volt ranges with 10,000 ohms load resistance is shown in Fig. 5, while the following table gives the calibration of the cardinal points of the scale for a range of load resistances when using the meter for 100 volts full-scale deflection.

The resistances must be capable of dissipating the L.F. or output of the valve under test, while the 4-mfd. series condenser should be increased inversely in proportion as the load resistance is decreased.

Decibels

Sometimes the scale instead of being calibrated in watts is marked in terms of plus or minus so many dB above or below a standard output of say 1 watt. A typical scale shape is shown in Fig. 6 and can be calculated from

$$20 \log_{10} \frac{V1}{V2}$$

where V1 is the power output being measured (in volts) and V2 the standard output in volts.

The 1-watt and 25-watt scales are obtained by using the 100-volt and 500-volt ranges of the instrument with a load resistance of 10,000 ohms. If the

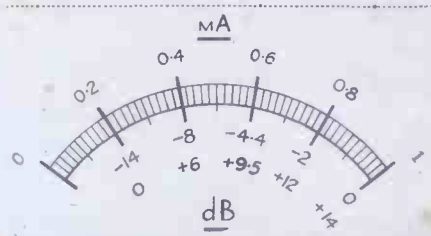


Fig. 6. 100v. and 500v. A.C. ranges in terms of decibels.

same marking is employed for other values of load resistance, the direct gain or loss in dB will still be correct, but the zero level instead of being 1-watt will become the full scale value given by the bottom figure of the appropriate column of the table above.