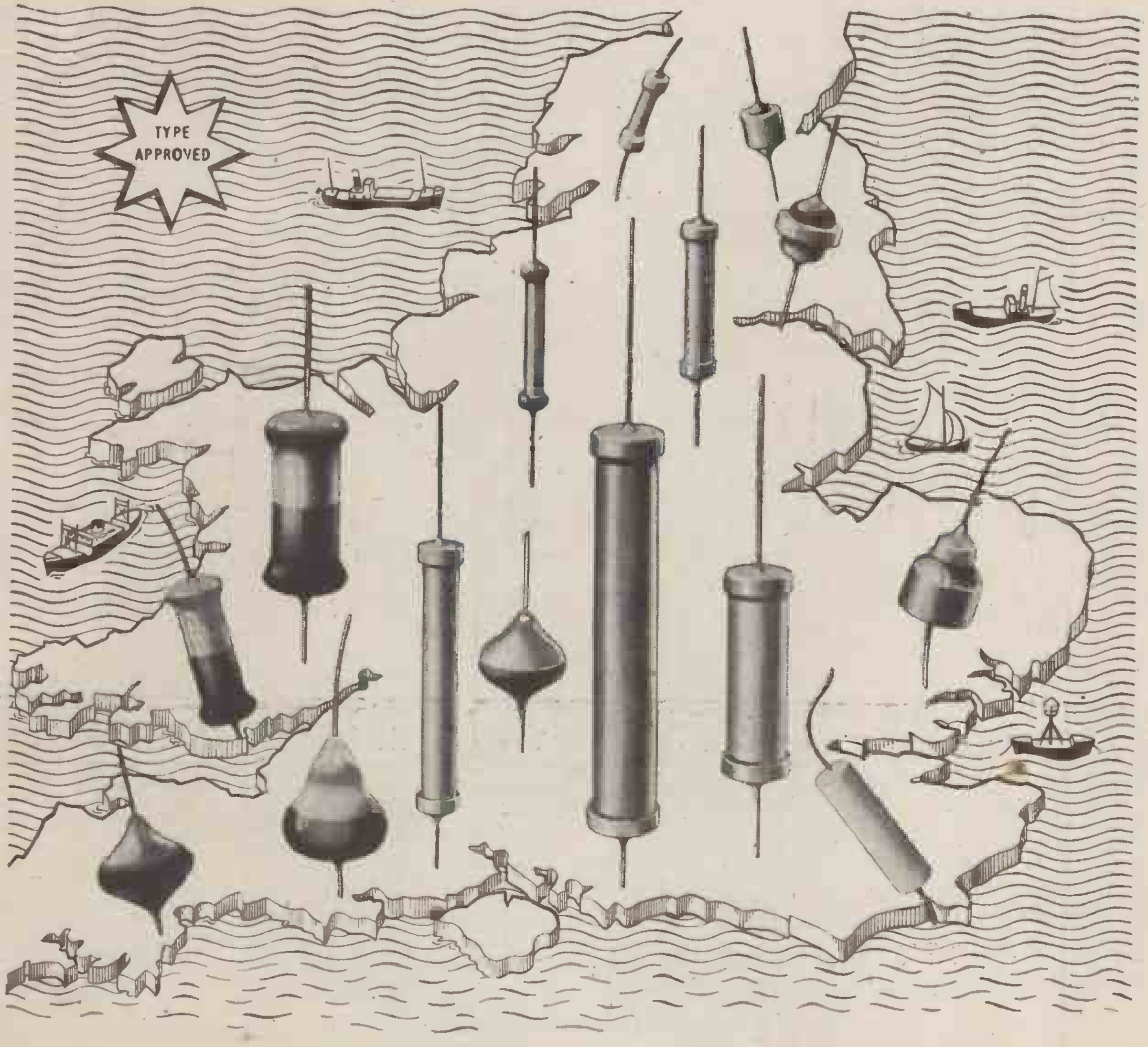
Electronic Engineering

INCORPORATING ELECTRONICS, TELEVISION AND SHORT WAVE WORLD

PRINCIPAL CONTENTS

2/- MAY, 1945



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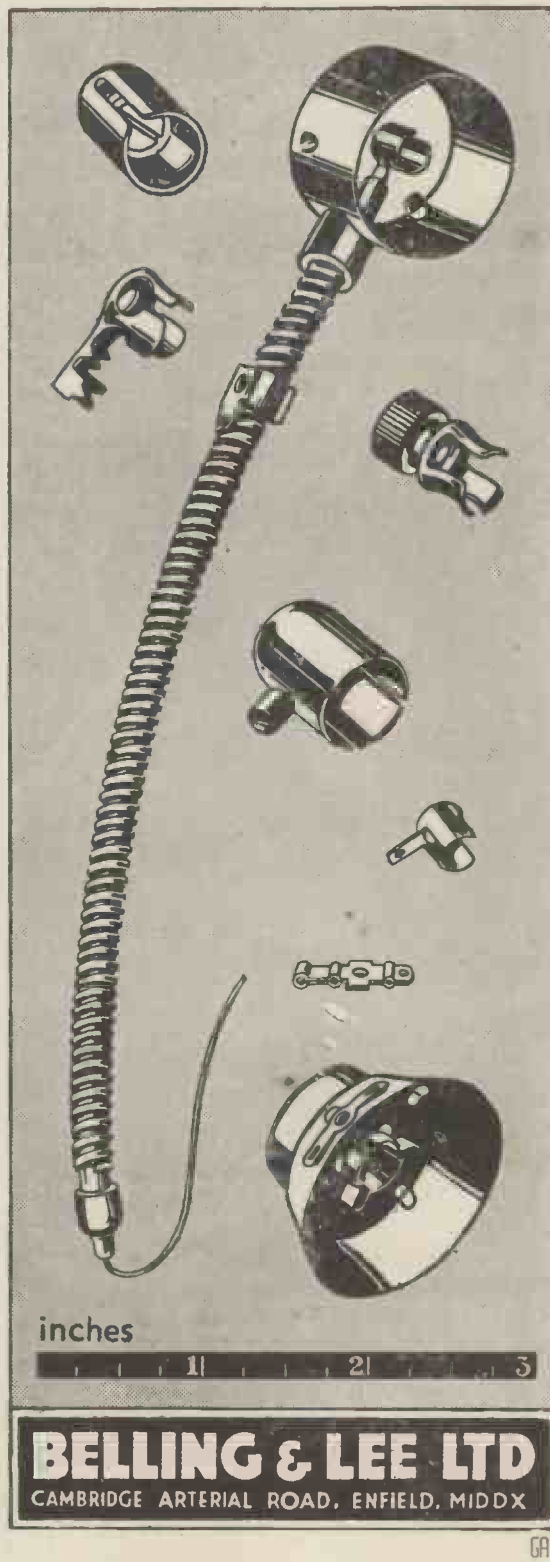
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Sir John Ambrose Fleming
1849-1945

WITH the passing of Sir Ambrose Fleming, D.Sc., F.R.S., at the age of 95, Britain has lost her oldest scientist and one more of the few remaining links with the illustrious past has been broken.

Clerk Maxwell, Rayleigh, Kelvin, Edison, Dewar, Swan—he knew them all. So much has happened in the past forty years, so much of the early work has been forgotten or taken for granted, that we find it difficult to realise that Fleming's life was virtually the life of electricity.

It is to Fleming that we owe the first potentiometer for the direct measurement of potential, the first standard lamp for photometry, the first wave-meter, the word " power-factor " and, his crowning achievement, the thermionic valve.

Educated at University College, London, and St. John's, Cambridge, his first University appointment was at the Cavendish Laboratory under Clerk Maxwell, and in 1882 he became Professor of Electrical Engineering at University College, a position which he held for over 40 years.

At the same time he became technical adviser to the Edison Swan Electric Co., at whose works at Ponders End the first thermionic valves were made. The effect first noted and described by Edison interested Fleming profoundly, and an account of his own investigations was given to the Physical Society in 1896. After he had been connected with the Marconi Company, the possibility of using the Edison Effect for the detection and rectification of high frequency signals occurred to him and in 1904 he brought out a patent for the first " Thermionic Valve " (No. 24,850).

From this has grown the enormous radio industry of the present day, although, as is usually the case, the founder received very little reward for his pioneer work. He was insistent that the word " valve " was the only appropriate name for the device and even in later years stigmatised the word " diode " as " scientific gibberish."

His scientific activities were carried on to the last, and he was for many years the President of the Television Society, presiding at the age of 85 at a meeting at which Sir Noel Ashbridge described the developments of the most recent branch of radio science.

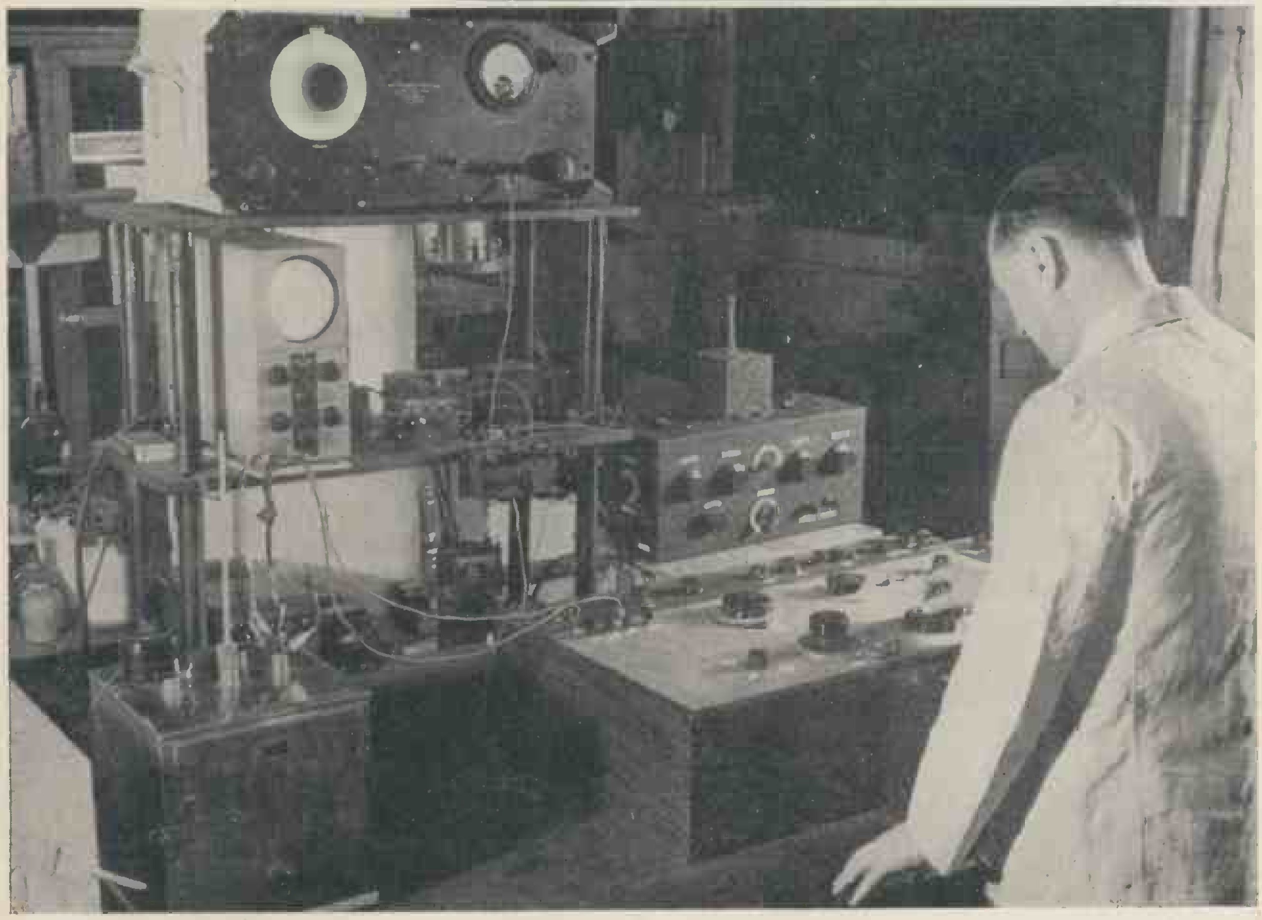
In the words of the Oration of the University of Liverpool: " John Ambrose Fleming has a conspicuous and most honourable place among those men of science whose work has enriched and enlarged the range of human communication."

The Work of The E.R.A.

The British Electrical and Allied Industries Research Association, to give it its full title, was organised in 1920 with the object of undertaking research on problems vital to the industry as a whole.

Dielectric research has always formed an important part of the programme, and the photograph below shows the test equipment for the measurement of losses in mica. The mobile cathode-ray tube equipment shown on the right is used for the recording of circuit-breaking phenomena.

The photographs are from a recently published book "Co-operative Electrical Research," which describes the work and future of the E.R.A.



The Design of Power Amplifiers for Operating Ink Recorders

By DONALD ROBINSON*

PEN recorders have many advantages over other oscillographs for making permanent records of such low frequency phenomena as are associated with electro-encephalography, electro-cardiography and some methods of submarine telegraphy.

They are cheap to operate (using plain paper instead of photo-sensitive material), robust and portable. Their use has been restricted, however, by the fact that appreciable power is required to drive them and hitherto power amplifiers capable of dealing with the frequencies involved—down to one cycle per second—have been inefficient and bulky.

The purpose of this paper is to consider the performance desired of such an amplifier; to analyse the performance of the two types of amplifier at present in use, and then to describe the design by the author, on rather different lines, of an amplifier which is efficient and compact and which should greatly facilitate the use of pen recorders.†

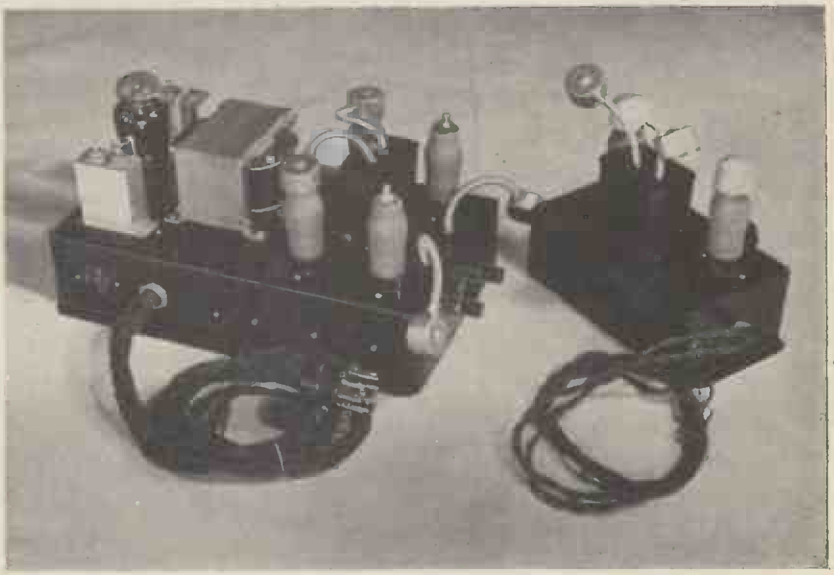


Fig. 6. Experimental single channel amplifier using the circuit described.

Requirements of the Amplifier

1. The frequency response must be linear from about 1 c/s. up to about 2,000 c/s. Although pen recorders do not operate above about 200 c/s., an extended high frequency response is valuable in case it should be necessary to use the amplifier with a cathode ray oscilloscope.

2. A common form of pen recorder is the energised moving coil type such as that designed by A. M. Grass in the U.S.A. An improved type (of which a description is yet to be published) is the four-channel moving iron design of Thorp and Tyrell. As both these types require about two watts for full deflection it is considered reasonable to take this as the minimum power requirement.

3. The D.C. resistance of both the above pens is about 3,000 ohms, and as the reactive component over the most important section of the frequency spectrum (2 to 20 c/s.) is small by comparison with the resistance, it will be taken *for the purposes of calculation* that the amplifier must work into a load of this value.

4. The amplifier should be of small bulk and weight and it should have a good power efficiency in order to have the minimum waste power to dissipate as heat. This is obviously desirable when these amplifiers have to be built into compact multi-channel equipment, or (particularly for the electro-cardiograph and electro-encephalograph applications) into portable equipment.

The First of the Known Methods

Since there is great difficulty in making a matching transformer that can handle the required power at 1 c/s. from a fairly high impedance source, the pen coils must be directly connected to the output stage.

As the current obtainable directly from a valve stage is small, adequate ampere-turns can only be obtained in the pen coils by winding them with as many turns as are practicable from the physical point of view. This means the use of fine-gauge wire and is the reason for the D.C. resistance of about 3,000 ohms.

The oldest and simplest method is to use a centre-tapped pen coil and to connect each half between H.T. and the anode of one of a pair of valves operating in Class AB_1.

Fig. 1 illustrates a reasonably optimum example of this method, employing two PX4 valves. Suitable operating conditions and an estimate of the performance is found in the following way:—

A number of values of the instantaneous anode voltage (E_a) are divided into the rated maximum anode dissipation of the valves, which for a PX4 is 15 watts. The resultant instantaneous anode current values (I_a) are plotted on the anode characteristics (which are also given in Fig. 1) and the curve "$W_{a \cdot max.} = 15$ watts" is drawn.

A load line AB is then drawn anywhere on the characteristics with a slope of $-\dfrac{2}{R_L}$ where R_L is the load impedance, 3,000 ohms.

If the point A is arbitrarily taken at $I_a = 0$ mA and $E_a = 200$ volts, then the load line must cut the $I_a = 100$ mA co-ordinate at $E_a = 200 - (0.1 \times 3{,}000/2)$, or 50 volts.

A position is then found for the dynamic load line CD which will be parallel with AB, but which

* Telephone Rentals, Ltd.
† A summary of the performance of the three types is given in Fig. 7.

must pass through the operating point P, the latter being so placed that the maximum reduction in E_a, that is, $E_{a.s} - E_{a.min.}$ (terms defined below) is not more than about 10 per cent. greater than the maximum rise in E_a, or $E_b - E_{a.s}$ (also *see* below).

The position of P gives the values of the static anode voltage $E_{a.s}$, the static anode current per valve, $I_{a.s}$ and the bias voltage E_k.

The intersection of the load line with the I_aE_a curve for $E_g = 0$ volt gives the minimum instantaneous anode voltage $E_{a.min.}$ and its intersection with the $I_a = 0$ mA co-ordinate gives the H.T. voltage, E_b.

The operating conditions resulting from the selected positions of points C, D and P in Fig. 1 are:—

Anode voltage	$E_{a.s}$	= 245 v.
H.T. voltage	E_b	= 325 v.
Bias voltage	E_k	= – 30 v.
Anode current per valve	$I_{a.s}$	= 55 mA.
Anode dissipation per valve .... $W_a = E_{a.s}.I_{a.s}$,	W_a	= 13.5 w.
Total H.T. consumption .... $(W_b = (E_b + E_k)2I_{a.s})$	W_b	= 39 w.
Continuous dissipation in the pen coils .... $(W_L = R_L(I_{a.s})^2.)$	W_L	= 9 w.
Power efficiency		= 15%

The maximum output power will approach:—

$$\frac{(E_b - E_{a.min.})^2}{2R_L} \text{ or about 6 watts.}$$

Surveying these results, it will be seen that while the output power is adequate and the power efficiency is reasonable, the continuous dissipation of 9 watts in the pen coil makes this method a very undesirable one.

These results may have suggested the next method, which avoids the need for any continuous dissipation in the pen coil. Before passing on to it, however, it may be useful to cover briefly, for the sake of comparison, the well-known method of estimating the performance of a normal transformer-coupled push-pull stage.

In such a case the load line must have a slope of $-(4/N^2R_L)$ where N is the turns ratio of the output transformer. If the D.C. resistance of the primary of this transformer is ignored the load line will pass through the intersection of the co-ordinates of $I_a = 0$ mA and $E_a = E_{a.s}$. It does not, however, pass through the operating point.

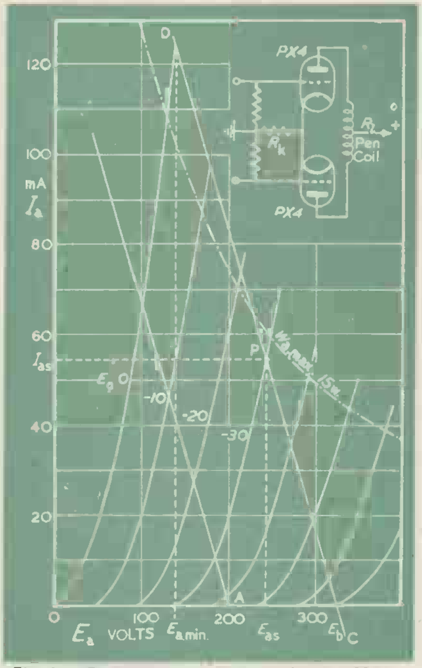


Fig. 1. Operating conditions of two PX4s.

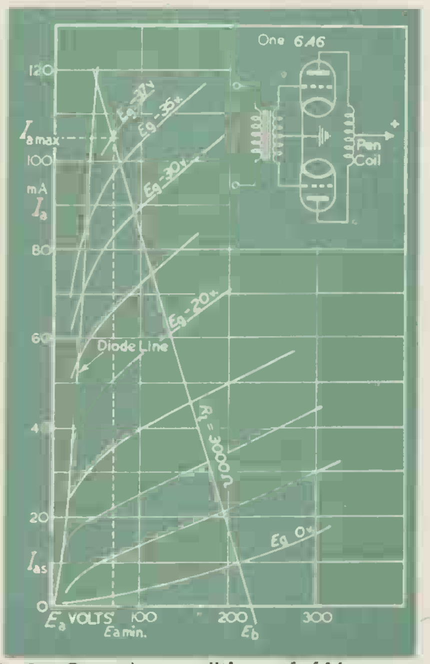


Fig. 3. Operating conditions of 6A6 output stage.

The power output is

$$\frac{2(E_{a.s} - E_{a.min.})^2}{N^2R_L}$$

and the power efficiency lies between 30 per cent. and 50 per cent.

The Second of the Known Methods

A more satisfactory method is to connect the pen coil from anode to anode and to feed the H.T. through two extra resistors and a balancing resistor as in Fig. 2.

This is the more commonly used method. It will be seen that it is a bridge arrangement that can be brought to balance by adjusting the slider of the low value resistor R_3. When this is done there is no D.C. flowing in the pen coil.

About optimum results for this method are obtained by the use of two PX25 valves with the series anode resistors taken as together equal to about three times the resistance of the pen coil, *i.e.*, about 4,500 ohms each.

This value is a compromise between too high an H.T. voltage and the loss of excessive A.C. power in these resistors.

Referring to Fig. 2 the dynamic load line has a slope of:—

$$\frac{2(R_1 + R_2 + R_L)}{R_L(R_1 + R_2)} \text{ or } -\frac{1}{1{,}200 \text{ ohms}}$$

The output power, calculated as for the previous method, should be 4.3 watts, but in practice odd harmonic distortion limits it to just under half this figure—about 2 watts.

It is believed that connecting a 2,000 μF condenser between a centre-tap on the pen coil and earth or H.T., and another one across the bias resistor, would enable the calculated power to be achieved, but the use of condensers of this capacity and of a suitable working voltage is more than a little impracticable and this idea has not therefore been tried.

The H.T. voltage is found in this method by drawing a further load line through the operating point with a slope of $-(1/4{,}500)$ ohms. Its intersection with the co-ordinate $I_a = 0$ mA gives $E_b = 570$ volts.

As Fig. 2 gives $E_k = 20$ volts and $I_{a.s} = 60$ mA, the continuous H.T. consumption $(E_b + E_k)2I_{a.s}$ equals 72 watts.

This represents a power efficiency of $3\frac{1}{2}$ per cent.!

Author's Inverse Feedback Class B Design

Class B operation results in a very low standing anode current and presents a means of avoiding the excessive heating of the pen coil associated with the first method and the heavy power requirements of the second.

The use of Class B brings other difficulties in its train but these have been overcome in the amplifier described below.

Fig. 3 shows the arrangement of the output stage and gives the anode characteristics of one section of the output valve employed, an American high-μ double triode, type 6A6 or 6N7. To avoid confusion each half of this valve will be referred to as if it were a separate valve.

The difficulty of obtaining the fixed bias required for Class B has been overcome by the method normally used for this valve, by operating at $E_k = 0$ volt, *i.e.*, at "zero bias."

The most important consideration in the selection of the operating conditions is that they should be such that it is impossible for even accidental overdriving to cause the anode dissipation to exceed the maximum safe figure.

It is not easy to find a practical value for the latter as anode dissipation is not at maximum at zero output as in Class AB_1, but varies with the drive. It is usually greatest when the output power is about half the maximum power obtainable without distortion.

In addition, in the audio application of Class B it is accepted, in view of the low average power which results from voice or music modulation, that the maximum dissipation can be 50 per cent. above the rated continuous dissipation and it can be taken that this also applies in this instance.

The maximum dissipation (*i.e.*, at half-full output) is usually about 20 per cent. above the dissipation at the maximum undistorted output level, and as it is more convenient to use the latter level as a basis for calculation it can be taken that the permissible dissipation at full output is 25 per cent. above the rated figure.

As the 6A6 is rated at 5 watts per element, the output power has therefore to be limited to a figure at which the anode dissipation does not exceed $6\frac{1}{4}$ watts.

This is done by making use of the "diode line" which provides a limit to the output power analogous to that provided by the flow of grid current in a Class A amplifier when the grid is driven positive.

The diode line passes through each I_aE_a curve at the point where it intersects the E_a co-ordinate equal in voltage to the E_g parameter of the

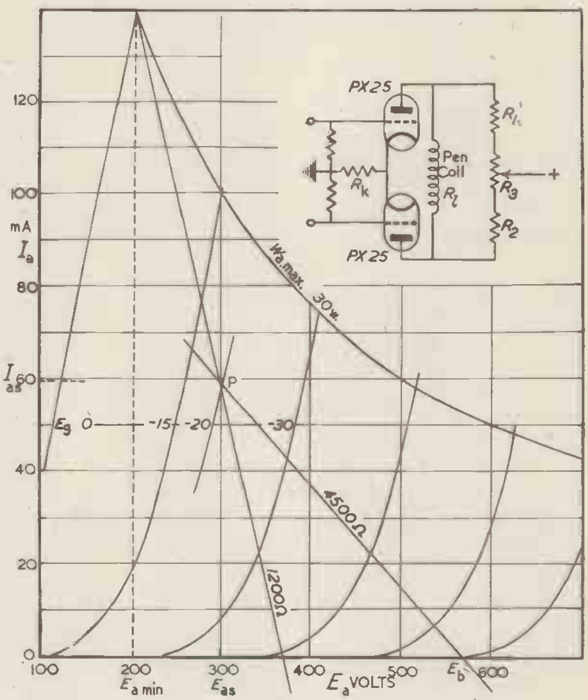


Fig. 2. Method of balancing by means of resistance across pen coil. Operating conditions for two PX25s are shown.

curve. To the left of the diode line $E_g > E_a$ and the grid loses control.

As the I_aE_a curves become rapidly more closely spaced as the diode line is approached, the practical value for $E_{a\cdot min.}$ is about twice the value of E_a at which the load line and the diode line intersect.

The load line must have a slope of $-(2/R_L)$ and the operating point is given by its intersection with the I_aE_a curve for $E_g = 0$ volt.

It will be seen from Fig. 4 that the position chosen for the load line gives :—

H.T. voltage .... E_b = 220 v.

R.M.S. anode current per valve at full output .... $I_{a\cdot d}$ = 35 mA.
$(I_{a\cdot d} = I_{a\cdot max}/2\sqrt{2})$

Static anode current per valve .... $I_{a\cdot s}$ = 10 mA.

Maximum undistorted output power .... W_o = 4 w.
$(W_o = (E_b - E_{a\cdot min.})^2/2R)$

Anode dissipation per valve at full output $W_{a\cdot d}$ = 5.7 w.
$(W_{a\cdot d} = I_{a\cdot d}.E_b - W_o/2)$

Anode dissipation static .... $W_{a\cdot s}$ = 2.2 w.

Continuous dissipation in the pen coil W = 0.3 w.
$(W_L = (I_{a\cdot s})^2R_L)$

Internal impedance of the output stage $2R_a$ = 22,000 Ω

Power efficiency at maximum output ... = 26%

The maximum anode dissipation leaves a small margin of safety, the figures for the output power and the continuous dissipation in the pen coil are very satisfactory and the power efficiency at full output is good. The power efficiency quoted is for the output stage only. If the loss in a valve type H.T. supply regulating device such as is described later is included, the figure will be about 18 per cent.

The reduction in the amount of power that has to be dissipated in the amplifier in the form of heat, compared with the two previous methods, is, however, greater than would appear from this figure, for whereas in the later methods the H.T. consumption is almost the same in the static condition as at full output, in the case of this amplifier the consumption in the static condition, even allowing for the loss in a valve stabiliser, is only about $6\frac{1}{2}$ watts.

Inverse Feedback

The only unsatisfactory features of the design as so far described are first, that the source impedance is very high, resulting in the poor "damping factor " ($F_L/2R_a$) of about 0.14, and secondly, that a certain amount of high order harmonic distortion is produced.

These are the usual defects of Class B operation, but they are also just the defects that can be overcome by the use of Inverse Feedback (I.F.B.).

While the conventional means of applying I.F.B. cannot be used in this application, the method shown in Fig. 4 is successful.

It consists of feeding a portion of the output voltage back to the grids of the driver valves by the connexion of a resistor in series with a blocking condenser from the anode of each output valve to the anode of each pre-driver valve.

The driver stage is dealt with later, but a feature of it that is of interest at the moment is that there is no phase change over it and hence the feedback voltage is in the correct phase.

The percentage of I.F.B. is determined by the relative values of the feedback resistors and the load resistors of the pre-driver valves, for the later resistors constitute the only significant component of the grid-to-earth impedance of the driver stage.

If the feedback resistors are made four times the value of the load resistors referred to, 20 per cent. of the output voltage will be fed back.

As :—

β (the feedback factor) = 0.2. (Resistor R_4 is 25 per cent. of R_3, giving 20 per cent. feedback.)

μ (the amplification factor) = 35.

R_a (the anode impedance) = 11,000 ohms.

E_o (the max. undistorted output voltage) = 112 volts r.m.s.

R_o (the load per valve = R /2) = 1,500 ohms.

M_D (the voltage gain of the driver stage) = 0.7.

The effects of the addition of I.F.B. are :—

(1) The source impedance drops from 22,000 ohms to an effective value of

$$\frac{2R_a}{1 - \beta\mu} \text{ or } 2{,}750 \text{ ohms}$$

This increases the damping factor from 0.14 to about 1.1, a figure comparable with that given by the two amplifiers described earlier or by a " high fidelity " audio amplifier.

(N.B.—The impedance of a typical pen recorder increases by about 50 per cent. at its resonant frequency, but the damping factor rises with it, in this case to about 1.6.)

(2) Distortion generated in the driver and output stage is reduced by a factor $1/(1 - \beta M)$ where M is the gain over the driver and output stages.

$$M = \frac{M_D \cdot \mu R_o}{R_o + R_a} = 2.7$$ and distortion is therefore reduced by about 5/6ths. It is, however, already fairly low as in Class B distortion varies with the load impedance and the resistance of the pen coil is about a third of the load usually used with the 6A6 valve.

The price to be paid for these advantages is small as M is so low. It is that the input voltage required by the driver stage (grid-to-grid, r.m.s.) is increased from $2E_o/M$ or 83 volts to $\dfrac{2E_o(1 - \beta M)}{M}$ or 122 volts—that is by about 60 per cent.

The feedback coupling condensers can be of quite small value as this will result in a useful increase in gain at the lowest frequencies. The values selected reduce the feedback at 1 c/s. by 50 per cent. and as a result the coupling condensers in the earlier stages can be of lower capacity with a corresponding reduction in time that the amplifier remains " blocked " after the application of an excessive input voltage.

The Driver Stage

The requirements of the driver stage are :—

(1) It must provide an output of $2\sqrt{2}(E_{a.min.})$ or 53 volts r.m.s.

(2) It must provide about 0.2 watt of power. The makers of the 6A6 quote a third of a watt when the 6A6 is used with a 10,000 ohm anode to anode load, but the input power required by a Class B stage falls as the anode load impedance is reduced.

(3) It must have excellent regulation as the power required to drive the 6A6 varies in a non-linear manner with the instantaneous input voltage.

(4) A low resistance D.C. path must be provided from the grids of the 6A6 to earth so that grid current (which

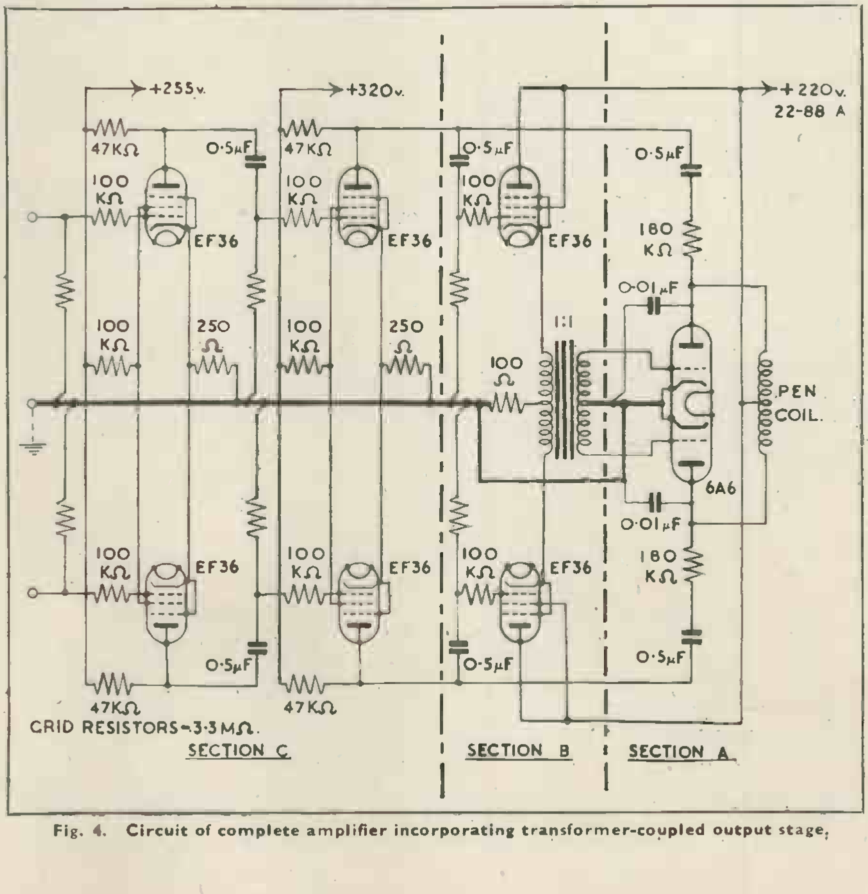


Fig. 4. Circuit of complete amplifier incorporating transformer-coupled output stage.

at full output is about 8 mA per valve) will not produce bias.

The transformer-coupled cathode-follower arrangement shown in Section B of Fig. 4 meets these requirements. The regulation in particular is excellent as the source impedance of a cathode-follower stage is twice the reciprocal of the slope of the valves used, giving a figure in this case of 360 ohms.

The key component of the whole amplifier is the driver transformer which must have a secondary of low D.C. resistance (*see* requirement 4 above), must not saturate at 53 volts 1 c/s. and must have adequate inductance for the stage to be linear down to 1 c/s.

It was felt that the low source impedance might make it just possible to design a transformer for this purpose and the problem was referred to Dr. G. A. V. Sowter, who is well known for his work on high permeability core materials.

He produced a transformer which can be said to have made the design of this amplifier possible. It is satisfactory from 10,000 c/s. down to about 0.5 c/s., even without inverse feedback.

H.T. Supply

The simplest method of obtaining an H.T. supply with adequate regulation is to use a mains transformer with low copper loss, a 5Z4, U.50 or other small full-wave rectifier and a swinging choke. The latter component, however, tends to radiate a lot of interference and in applications where very high voltage gain is needed it is preferable to use a simple valve stabilising arrangement such as that shown in Fig. 5.

The latter circuit also provides a "decoupled" source of H.T. for the voltage amplifying stages.

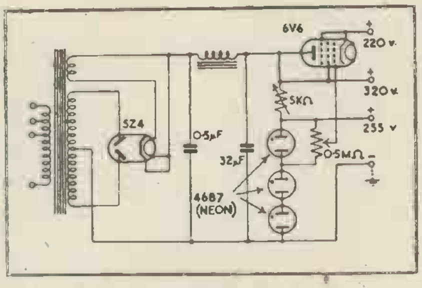


Fig. 5. Stabilised H.T. supply.

The Complete Amplifier

The foregoing survey of the methods of driving pen recorders was part of the design work associated with the production of a four-channel electro-encephalograph on which the writer is working in collaboration with Mr. Thorp of the Pharmacological Department of the Wellcome Physiological Research Laboratories.

Fig. 6 illustrates the experimental single-channel unit built there. Its performance conforms fairly closely with the calculations given above and it is very stable and free from hum.

It is linear from 0.5 c/s. to 6 kc/s. and it is only 3 db down at 10 kc/s.

There is no sign of "cross-over" distortion and harmonic distortion is considered to be well under .5 per cent. total.

The circuit of the first two stages (one of which, being experimentally an afterthought, is on a separate chassis) is given in Section C of Fig. 4. The common bias and screen resistors provide phase inversion and the amplifier can be driven with an input applied to the grid of either input valve, or if the phase is correct, to both of them.

The amplifier is differential to the extent of about 75 : 1, *i.e.*, the output voltage obtained with an input applied in the same phase to both input valves is about one seventy-fifth of the voltage obtained when the same input is applied to one input valve only.

Amplifiers of this general type are often required to exhibit this feature in order that they shall not amplify hum picked up by an object to which their input leads are attached.

The performance of the three amplifiers discussed is summarised in Fig. 7. The figures for the Class B amplifier include the voltage drop, etc., in the H.T. regulating valve of Fig. 5.

FIG. 7

Amplifier	As Method 1	As Method 2	Author's
Power waste in pen coil	9 watts	0 watt	0.3 watt
H.T. voltage	325 volts	570 volts	320 volts
H.T. consumption, static	39 watts	72 watts	6.5 watts
H.T. consumption at full output	39 ,,	72 ,,	22.5 ,,
Maximum output power	6 ,,	2 ,,	4 ,,
Power efficiency	15%	3½%	18%
Damping factor	1.25	1.1	1.1

Balanced Amplifiers

A Letter to the "Proc. I.R.E."

SEVERAL articles have recently appeared on amplifiers employing cancellation of in-phase signals. These include differential input amplifiers; direct-current phase inverters; and resistance-capacitance-coupled push-pull amplifiers.

In any amplifier having three input terminals and/or three output terminals, the input voltages, e_1 and e_2, may be written

$$e_1 = (e_1 - e_2)/2 + (e_1 + e_2)/2$$
$$e_2 = -(e_1 - e_2)/2 + (e_1 + e_2)/2$$

The first term in each case is the differential signal; the second, the "in-phase" signal. If the amplifier does not transmit the in-phase signal, while the amplification of the differential signals is μ_1, μ_2 the output voltages will be

$$E_1 = \mu_1(e_1 - e_2)/2$$
$$E_2 = -\mu_2(e_1 - e_2)/2$$

In a push-pull amplifier, $\mu_1 = \mu_2 = \mu$ and the differential output voltage is μ times the differential input. In a phase inverter, $e_2 = 0$, but $E_2 = -E_1 = -\mu e/2$. In a differential input amplifier, which is followed by single-ended stages, $\mu_2 = 0$, and $E_1 = \mu_1(e_1 - e_2)/2$.

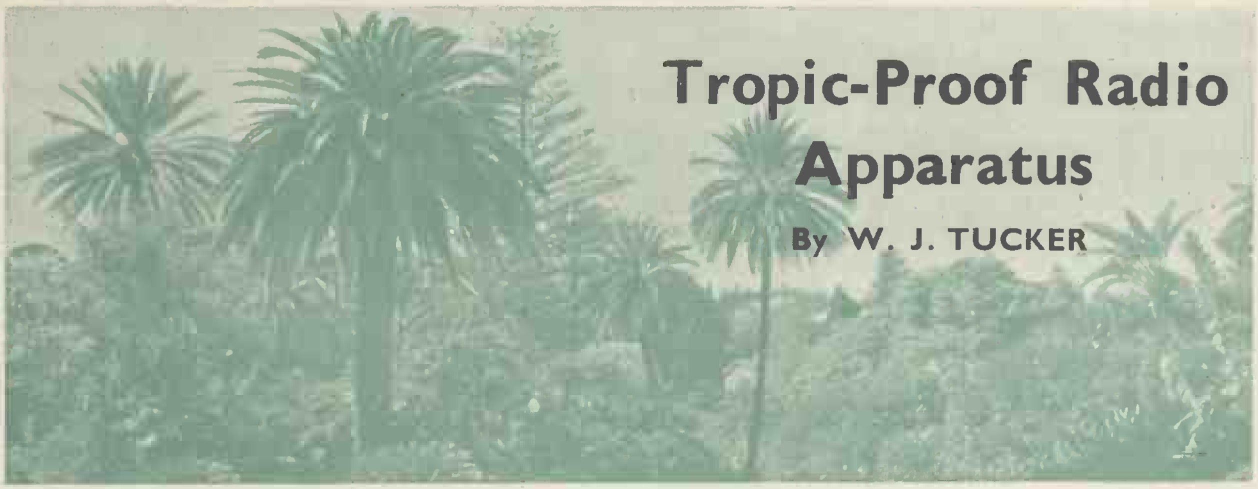
Thus it is seen that all three types of amplifiers are merely special uses of in-phase signal degeneration. Perhaps the failure to recognise this fact may in part account for the frequent republishing, as original, of several of the circuits the writer developed for this purpose. These have been used in most of the equipment we have built since 1936, and because of their apparent wide usefulness were made available to many workers in biophysics in private communications, in advance of publication.

Thus the so-called "Toennies" differential input amplifier is merely an application of in-phase degeneration by use of a large cathode resistor, where μ_2 is made zero; *i.e.*, one output terminal is disregarded.

However, some of the circuits recently published appear to be identical both in structure and function to those the writer has published, although no reference has been made thereto.

FRANKLIN OFFNER,
Offner Electronics, Inc.,
Chicago, Ill.

—*Proc. I.R.E. Mar.*, 1945.



Tropic-Proof Radio Apparatus

By W. J. TUCKER

THE problems associated with rendering radio apparatus proof against the deleterious effects of exposure to tropical and sub-tropical conditions are engaging the attention of designers and production engineers throughout the industry.

If we are to develop our export industry, the degree of success which will attend our efforts will depend on our ability to produce reliable apparatus suitable for use all over the world.

To tackle the problems of correct design for tropical conditions, it is necessary to fully appreciate what these conditions are, and to be aware of the faults and most prevalent causes of breakdown.

It has been well established that the defects which arise in radio apparatus exposed to tropical conditions are usually caused by the high moisture content of the air, and temperature variations which cause heavy condensation.

In tropical areas high relative humidity is maintained throughout the year with only a slight drop in the winter season. 95 per cent. relative humidity is quite common in many areas and the average figure for all areas which can be designated as "tropical" is never lower than 80 per cent. A typical day and night temperature cycle during the summer months would be 40° C.-60° C. during the day falling very slowly to 20° C.-25° C. in the night. The temperature in any given area is affected by the proximity of the sea and the height above sea-level. The relative humidity, however, is sufficiently high in all parts to constitute an ever-present source of danger to all equipment and materials which are in any way susceptible to attack by moisture.

In humid atmospheres moisture will penetrate through the smallest of pin-holes and crevices and, generally speaking, very little opportunity arises for this moisture to be dried off.

Many materials absorb moisture from the air and the effect of temperature rise is to drive air out from semi-enclosed places, such an unsealed coil and transformer cans. A decrease in temperature causes the air to contract and moisture-laden air is drawn into the semi-enclosed places. In addition, wide temperature variations cause condensation on non-absorbent surfaces, and the water which collects sets up corrosion and encourages the growth of fungus. Fungus has a destructive influence on many materials, and during the process of disintegration, chemical reactions occur which will accelerate corrosion and decomposition.

In such atmospheric conditions, despite all precautions which may be taken to ensure adequate protective packing and storage, the useful life of radio apparatus is usually determined by the ability of the insulating materials employed to withstand moisture penetration.

Effect of Moisture

Electrical insulation is attacked by moisture in one or other of the following ways.

The absorption which takes place lowers the insulation resistance of the material, causes surface leakage, leads to distortion, swelling, and general mechanical deterioration. In time, the electrical constants of the material are seriously impaired.

The surface insulation is also influenced by the condensation of water on the surface and by fungus growth.

The maintenance of electrical insulation at 100 per cent. efficiency throughout the service life of a radio equipment is of major importance, but it is also necessary to give adequate attention to other materials used in such apparatus.

Metals, unless protected by a suitable treatment, will corrode very quickly. It has been found that this corrosion takes place much quicker in humid atmospheres than in straightforward water immersion tests. Electrolytic action between metals of widely different electrical potential is a well-understood phenomenon, and the process of corrosion which occurs when two such metals are in contact with water is considerably accelerated when these metals are also exposed continually to a moist atmosphere.

Wood is very badly affected by humidity and also by fungus growth. The rotting process set up by moisture can be retarded by the use of rot-proofing compounds, which serve also as protection against ants and fungus growth.

Leather and glass are both attacked by fungus growth, although the attack on glass is indirect, insofar as fungus spores thrive on organic dust which settles on glass surfaces and cause an etching effect on the glass, thus impairing visibility through the windows of meters, tuning dials, etc.

Among the insulating materials affected are linen, cotton, and all cellulose products, such as paper, varnished fabric, empire cloth and insulating tape. These materials may be protected to a limited degree by varnishing or dipping, but the degree of protection afforded has been proved to be inadequate.

Most of the plastic materials are satisfactory, the exceptions being cellulose acetate and fabric laminated phenolic sheets. Some of the paper laminated classes of material are unsatisfactory, but most of the trouble arising with these is attributable to bad design of the components which fail. Laminated paper material has its limitations, but used with discretion, it can be proved to be satisfactory for many applications.

Protective Finishes

So far as protective coatings are concerned, recent evidence has shown that varnishes are to be preferred to waxes, since the latter, unless of mineral origin, grow fungus; and both mineral and vegetable waxes show tendencies to soften and deform at temperatures too low to leave an ample safety margin. The soft waxes also collect dirt and dust which becomes embedded in the surface layers of the material. When moisture penetrates these surface layers the collected impurities form a leakage path across the surface of the component.

Varnishes (on materials other than those quoted above) afford satisfactory protection provided that they are applied properly. The protective surface of the varnish coating must be complete and free from pinholes and cracks. Varnish cannot be used, therefore, on components which are subject to flexing.

Rubber is subject to deterioration in sunlight and does not resist moisture penetration as well as some of the synthetic rubber-like materials.

Brass in particular is susceptible to a form of corrosion set up when in contact with ammonia, and free ammonia is released from some phenolic moulding compounds.

The application of protective finishes to metals and other materials has been the subject of special tests and study. It is proposed to deal in detail in a later section with all the protective measures which have been developed.

Layout of Components

The layout design of radio apparatus affects its operational reliability to a considerable degree. The most common fault is, strangely enough, compactness. In all the various types of apparatus which have been the subject of special review and test, the compact and tightly packed units have been the most prolific source of trouble. This is due to the fact that moisture penetrates into all but totally sealed places and has no opportunity to dry out. The more compact the apparatus the greater the risk of this trapped moisture settling on metal parts and being absorbed by insulating materials generally. It is not suggested that the only remedy is to make all apparatus large enough to clean out with a bucket and broom. The layout, however, should be such that moisture traps are not created, and if space limitations do not permit of an expansive layout then components should be arranged to permit drainage through ventilating holes in the bottom of the cabinet or cover.

Moisture traps are created by placing sheets of laminated phenolic material together, sandwich fashion. If these laminated sheets or strips are carrying contacts then the moisture which is trapped between them will create leakage paths across from contact to contact.

Stacking up moulded case condensers sandwich fashion, and securing them with two common fixing screws to the chassis is another prevalent assembly practice which should be discontinued. The moulded cases absorb a certain amount of moisture which causes surface leakage across from one condenser to another, thus completely upsetting circuit constants.

Parts of the apparatus which run hot should not be permitted to be adjacent to waxed components or in contact with thermo-plastic insulation such as polystyrene and polythene. Precautions should be taken to keep dirt and dust off the vanes of air-spaced condensers. Fungus growing on this dust will cause short circuits. Windows and escutcheons should not be made of cellulose acetate or methyl methacrylate. These materials will warp after extensive exposure to humid atmospheres.

An example of the influence of tropical requirements on component design: A totally enclosed moisture-proof potentiometer (Type H) made by The Morgan Crucible Company, Ltd.

The most important point to watch in apparatus layout is wiring.

In compact layouts it is extremely difficult to avoid bunching leads together or allowing criss-cross patterns in the wiring to develop.

Moisture absorbed into the insulating sleeving turns the latter into high resistance conductor, with the result that all sorts of peculiar and unexpected circuit conditions are set up. The electrical properties of the sleeving, even if it remains an insulator, are so changed by moisture absorption that the normal and intended circuit arrangement will be completely upset.

The absorption of moisture into fabric sleeving takes place in two different ways. The absorption can take place through cracks in the varnish coating and penetrate through the wall to the conductor, or it can be drawn along the sleeving from any open end by capillary action. The latter form of absorption is the most serious since the moisture can be conducted right into the contact surfaces of the component, and there is every chance of it accumulating until breakdown takes place across adjacent contacts.

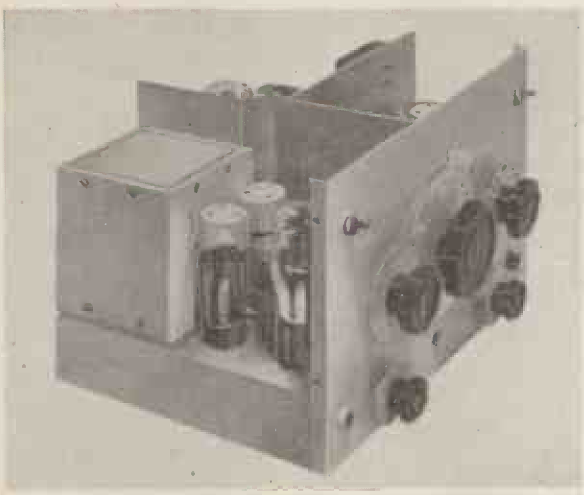
Wires which are bunched together and then get saturated with moisture are a source of considerable trouble due to the lowered insulation resistance from one conductor to another. Also fungus will thrive on cellulose surfaces and will cause disintegration.

Plastic sleeving of the polyvinylchloride type is the real answer to all these problems. The plastic sleevings are stable over a temperature range quite wide enough for normal tropical use. They are completely impervious to moisture and will stand repeated flexing during assembly.

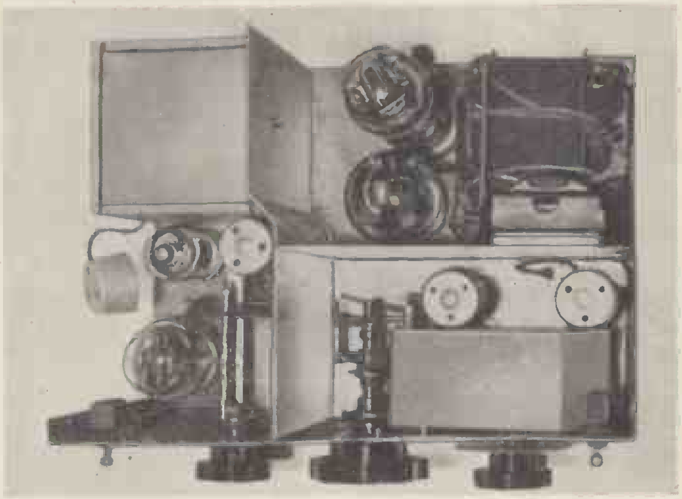
In a later article it is proposed to deal in more detail with protective finishes and general component design. Tables and curves showing the results of specific tests will be provided.

A Frequency Meter with Self-Contained Standard Oscillator

By G. P. ANDERSON



General view of Frequency Meter.



Internal view.

BY means of the unit to be described it is possible for the experimenter of moderate means to measure radio frequencies up to 30 Mc/s. or higher, with an accuracy of approximately 1 part in 10^4, using ordinary components. This figure will be largely dependent upon the accuracy of the 100 kc/s. standard oscillator, which should preferably be controlled by a 100 kc/s. quartz bar; details are, however, included for the construction of a simple stable electron-coupled oscillator, and suggestions are made for the calibration of such an oscillator. Provision is also made to enable a sub-standard set of harmonic frequencies spaced every 100 kc/s. and 10 kc/s. to be available for the calibration of receivers and test oscillators; a signal may also be provided at 2 Mc/s. and harmonics thereof. By means of these frequencies, a receiver may be completely and accurately calibrated throughout the short wave range.

The unit consists of:—

(1) A 100 kc/s. standard frequency oscillator (SFO); a 10 kc/s. multivibrator (MV); and associated harmonic amplifier.

(2) A variable frequency oscillator (VFO) covering a range of approximately 1.4 to 2.8 Mc/s., and including its own harmonic amplifier.

(3) A valve for detecting the heterodyne or " beat " of the VFO against the standard frequencies, or against the carrier of another oscillator, such as a transmitter.

(4) A self-contained power supply.

Sections

1.1. 100 kc/s. Oscillator

A small RF pentode is used in this stage, and follows normal oscillator design. Care should be taken to provide thermal insulation for the crystal, or the coil and condenser in the case of the electron-coupled model. This may be done by constructing a small box of balsa wood, such as is used in model aircraft making. Reduction by a few cycles of the fundamental crystal frequency may be made by fitting a small trimming condenser across the crystal.

1.2. Multivibrator

The principle of operation of the multivibrator circuit is well known, and in this unit a twin-triode valve is used, and locked on 10 kc/s. by feeding the output of the 100 kc/s. oscillator into the cathode circuit. A control is provided on the front panel for adjusting the MV to produce harmonics 10 kc/s. apart, although the circuit used is sufficiently stable for this control to be ignored once it has been set.

1.3. Harmonic Amplifier

Another small RF pentode is used as an amplifier, and it also serves to isolate the oscillators from the effects of variation in the external loading. No provision is made for tuning the output of this stage, as it was found to produce signals of satisfactory strength over the range required.

A Yaxley-type switch is used to connect either the 100 kc/s. or the 10 kc/s. output to the grid of the amplifier, and also to stop the MV oscillating when using the 100 kc/s. oscillator alone. A third position is provided to switch off the standard frequencies by disconnecting the H.T. supply to the 100 kc/s. oscillator, as well as suppressing the 10 kc/s. oscillations.

2.1. Variable Frequency Oscillator

This consists of an electron-coupled oscillator using a RF pentode, and while short-period stability is very necessary, the method employed for frequency measurement does not call for the retention of highly accurate calibration over long periods.

The oscillator is designed to cover a total range of approximately 2.8 to 1.4 Mc/s.—a frequency ratio of 2 : 1 —the harmonics of which will provide complete coverage over the short-wave bands. This range is obtained by increasing the tuning capacity in the grid circuit by 12 steps of 25 $\mu\mu F$ each, and using a variable condenser with a swing of about 30 $\mu\mu F$ to cover each step. This tuning condenser, C_{14}, is made up of a larger variable condenser modified by the removal of plates until the capacity swing is just sufficient to cover each step on C_{13}. The frequencies covered by each range on the original model are set out in Table I, and it will be seen that each range slightly overlaps those adjacent to it.

Care should be taken to insulate the coil and condensers associated with the grid circuit from the heat of the valves; in the original model the coil was arranged below the chassis, and balsa wood sections were fitted between the valve and the condensers. An on-off switch is provided in the H.T. supply to switch off the VFO section.

2.2. Harmonic Amplifier

This is of the same design as that following the SFO, but is permanently connected to the output of the VFO.

3.1. Detector

The outputs from the two amplifiers are fed through small capacities to the grid of a triode valve, where they are mixed and detection takes place. A resistance-capacity output circuit enables a pair of headphones to be used to listen to the resulting heterodyne frequencies. A further lead from the grid through a small capacity to a terminal on the back of the unit allows a signal to be fed from the oscillators to a receiver, a few inches of insulated wire wrapped round the aerial lead being sufficient to provide strong signals in a short-wave receiver. It may also be used to enable an outside signal to be fed to the detector, to be mixed with either a standard frequency or the VFO.

4.1. Power Supply

The H.T. consumption of the unit is about 20 mA, and in the interests of economy, both of space and purse, resistance smoothing is utilised, and a perfectly smooth signal is obtained from both oscillators. A transformer supplying 250-0-250 volts was used, and approximately 120 volts is available across C_{28} when the unit is in operation. The heater windings required depend upon the valves used. A switch S4 is provided in order that the heaters may be left on during periods when no oscillation is required.

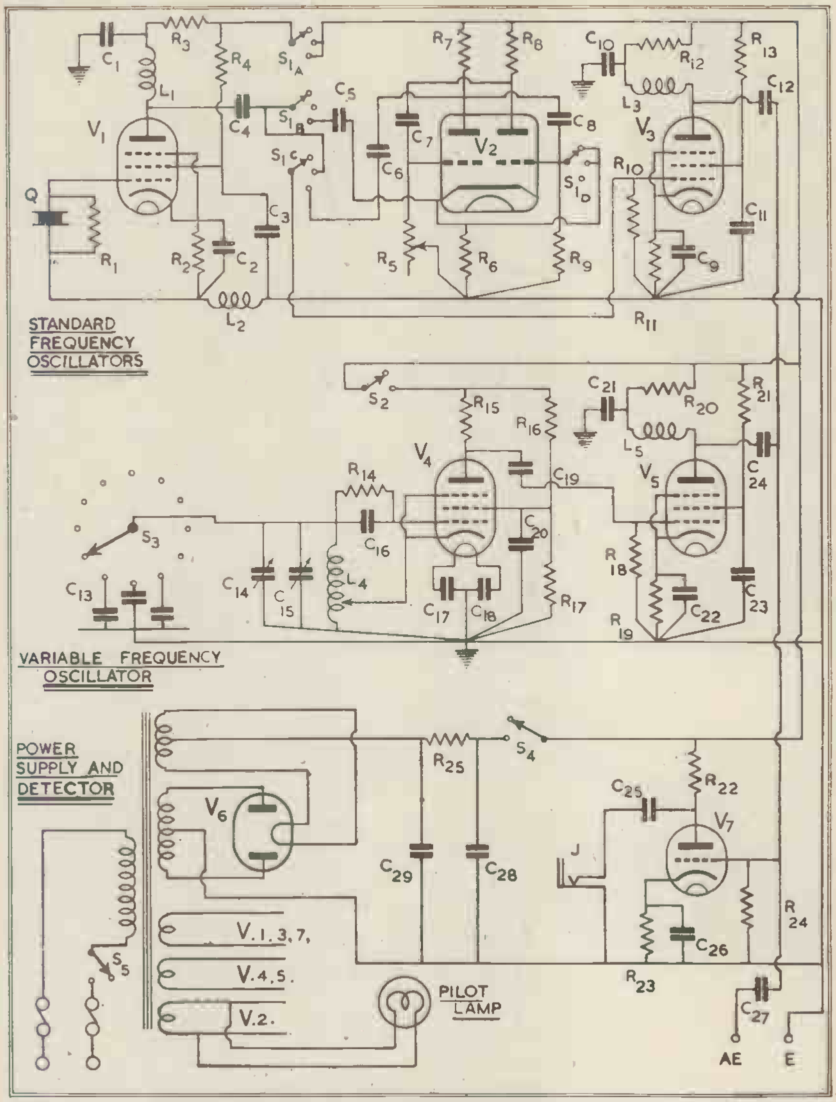
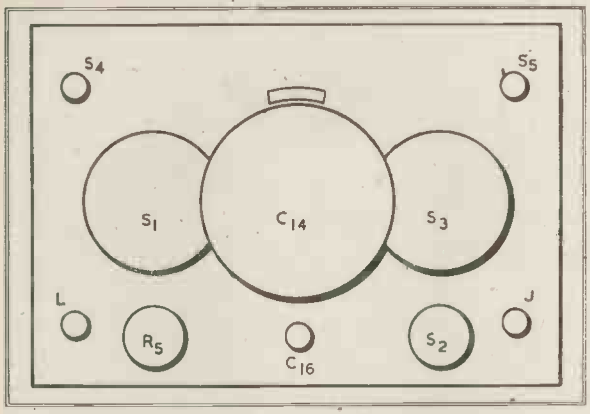


Fig. 1. Circuit diagram of the frequency meter and self-contained standard oscillators.

Component Values

C 1, 2, 3, 9, 10, 11, 17, 18, 20, 21, 22, 23, 33	0.01 μF.	V 1, 3, 4, 5	6K7, KTW 61, SF 41, etc.
C 7, 8, 12, 24	0.002 μF.	V 2	53, 6A6, 6N7, etc.
C 6	0.001 μF.	V 6	5Z4, MU 12/14, etc.
C 31	0.0004 μF.	V 7	6C5, ML 4, etc.
C 4, 16, 19, 27, 32, 34	0.0001 μF.		
C 5	0.00005 μF.		
C 14 30 $\mu\mu$F variable.	C 15 150 $\mu\mu$F trimmer.		
(C 15 may be made up of a smaller trimmer, and fixed capacities.)			
C 30	160 $\mu\mu$F variable.		
C 25	0.5 μF.	S 1	4-pole 3-way Yaxley type switch.
C 26	50 μF.	S 2	Rotary On-Off switch.
C 28	16 μF.	S 3	1-pole 12-way Yaxley type switch.
C 29	8 μF.	S 4, 5	On-Off toggle switch.
C 13	0—275 $\mu\mu$F, in steps of 25 $\mu\mu$F.		
(Step "1" = 275 $\mu\mu$F ; Step "2" = 250 $\mu\mu$F . . . Step "11" = 25 $\mu\mu$F ; Step "12" = 0 $\mu\mu$F.)			
R 1	1 MΩ ½ watt.	R 9, 16, 17, 22	33 KΩ 1 watt.
R 2, 11, 19	220 Ω 1 watt.	R 10, 18, 24	470 KΩ ¼ watt.
R 3, 12, 20	4,700 Ω 1 watt.	R 14	39 KΩ ¼ watt.
R 4, 13, 15, 21, 27	100 KΩ 1 watt.	R 23	1,000 Ω 1 watt.
R 5	30 KΩ potentiometer.	R 25	6,800 Ω 3 watts.
R 6	330 Ω 1 watt.	R 26	100 KΩ ¼ watt.
R 7, 8	2,700 Ω 1 watt.	R 28, 29	10 KΩ 1 watt.
L 1	RF Choke (Eddystone 1066).		
L 2, 3, 5	RF Choke (Eddystone 1010).		
L 4	42 turns, tapped 20 turns from earth, 1 in. diameter, 24 DSC.		
L 6	350 turns, tapped 90 turns from earth, 1 in. diameter, 40 Enam.		



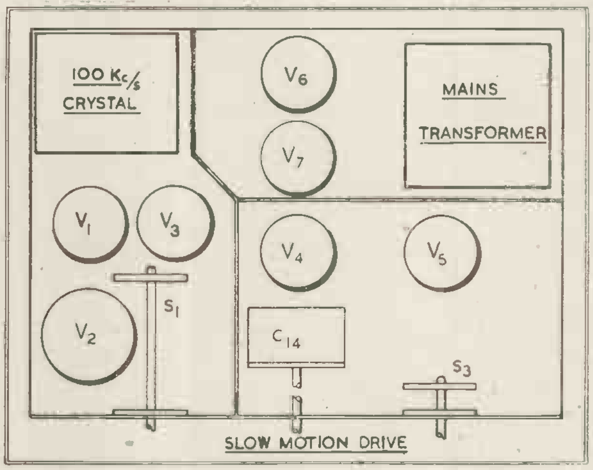


Fig. 3 (above). Front elevation of unit, showing position of controls.
Fig. 4 (right). Plan view of unit, showing layout.

Construction

A chassis was available measuring 11 × 8 × 2 in., and this was found to be just large enough to take the apparatus. The front panel is 7½ in. high, and the screens between the sections are soldered to the chassis and to the panel, pieces of similar shape being fitted above and below the chassis. In order to complete the screening, the finished unit should be fitted into a metal cabinet, with care taken to ensure adequate ventilation. It is suggested that small holes be drilled through the sides and back just below the level of the chassis, and similar holes be made around the valve holders; heating of the valves will then set up convection currents in the air, and an outlet should be provided towards the top of the cabinet.

A good quality dial, of the precision type and fitted with a vernier, should be used for the VFO, since the accuracy of the measurement of frequency will depend upon the accuracy with which the VFO dial can be read. For convenience the dial should be arranged to read "0" when the condenser is at maximum capacity, so that an increase in dial reading will indicate an increase in frequency. A reliable slow motion drive should also be fitted, and should be free from backlash.

Care should be exercised in wiring the unit, and 16 S.W.G. wire should be used wherever possible. All wires connected to earth in each stage should be joined at one point, and these in turn interconnected by one earth "main," which should be connected to the chassis at one point only, at the junction of the earth "main", and the earthy leads in the VFO stage (Fig. 1).

All supply leads (H.T. and heater) should be run in screened wire, with the screens bonded to earth at intervals. It is also advisable to run the output leads from each stage to the detector in screened lead. With these precautions and the decoupling provided, no trouble whatever was experienced with instability or hand capacity.

It is suggested that if the ECO circuit has to be employed in place of a 100 kc/s. crystal, the tuning condenser C_{30} could be controlled from the front panel by means of an extension handle. This control could take the place of the MV locking control R_5, which, as pointed out earlier in the article, very rarely requires adjustment once it has been set, and could therefore be fitted inside the chassis.

Setting up the Oscillator

First check the 100 kc/s. oscillator with the switch in the "100 kc/s." position; if crystal control is used a check on a short-wave receiver with the BFO switched on, or with the detector oscillating, will soon show if this stage is working satisfactorily;

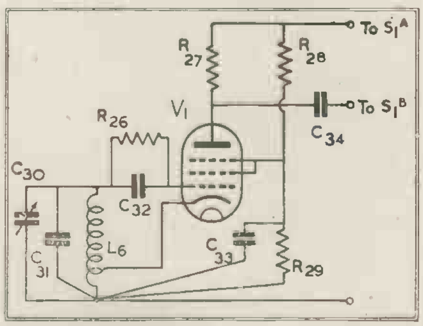


Fig. 2. Electron-coupled oscillator, for use where a 100 kc/s. crystal is not available.

as good strong signals are produced around 7 Mc/s., this is a good frequency at which to make this test. If the ECO circuit is used, first let the apparatus run for a half-hour before attempting to calibrate the oscillator. Then, by listening to the Overseas Programme on 1,500 metres (200 kc/s.), adjust C_{30} until a loud heterodyne is obtained. C_{30} should then be set to the position of "zero beat." Further adjustment may then be made by checking the beat with suitable short-wave stations, such as the American Standard Frequency station WWV, on 5, 10, or 15 Mc/s.

Having obtained satisfactory operation of the 100 kc/s. section turn the switch to "10 kc/s.," and adjust R_5 to lock the MV on the 10 kc/s. harmonic. This is done by setting a receiver (a set covering the medium-wave range and provided with reaction or a BFO is very suitable) to one of the 100 kc/s. points while running the 100 kc/s. oscillator only, and then switching on the MV. If the receiver is now tuned up or down in frequency, a series of evenly spaced whistles will be heard. If they have a rough sound, slight adjustment of R_5 should be made until they assume the pure note indicating stable control by the 100 kc/s. oscillator. The number of whistles audible between 100 kc/s. points should be counted, and adjustments made to R_5 until there are nine such points *between* each 100 kc/s. point, *i.e.*, they are spaced 10 kc/s. apart. This will be found to be when R_5 has approximately three-quarters of full resistance in circuit. This adjustment will not be found to be very critical, and in making it use may be made of the detector circuit incorporated in the unit; the procedure will be apparent from a study of the notes relating to this section which appear below.

As has been stated, the VFO should cover a frequency range the limits of which are in the ratio of 2:1, and in order to obtain this condition adjustment of C_{15} is made as follows.

First set C_{15} to a capacity of approximately 75 $\mu\mu$F, and set the range selector switch S3 to the lowest frequency range, "1," and the tuning dial to "0." Switch on the VFO only, and search on the receiver for the second harmonic of the VFO signal; this should lie between 2.5 and 3.5 Mc/s., and the approximate frequency or receiver dial reading should be noted. Let us call it "$2F_{min}$" since it is the second harmonic of the lowest frequency of the VFO.

Now turn S3 to range "12" and the tuning control to minimum capacity; again search for the VFO signal in the receiver between approximately 2.5 and 3.5 Mc/s. This time the signal heard will be the fundamental frequency and the position should be noted. Let this be "F_{max}."

It is apparent that if we are to obtain the frequency ratio of 2:1 referred to above, then "F_{max}" must be a higher frequency than "$2F_{min}$," the exact values of these frequencies being unimportant, and this condition may be achieved by adjustment of C_{15}, following this simple rule:—

If "F_{max}" is less than "$2F_{min}$," decrease C_{15} slightly, and *vice versa*.

After this adjustment a check should be made on "$2F_{min}$" and on "F_{max}," and if necessary further adjustments made to C_{15} until the desired coverage is attained. As will be seen from an examination of Table 1, the total overlap of "F_{max}" over "$2F_{min}$" in the original model is approximately 30 kc/s.

Now check the operation of the detector by inserting headphones in the jack on the panel, and switch on both the VFO and the 100 kc/s. oscillator. Upon rotating the tuning condenser a series of whistles should be heard at different dial readings. (In addition to the main 100 kc/s. heterodynes a number of weaker ones will be heard due to harmonics of the VFO beating with higher order harmonics of the 100 kc/s. oscillator, but there should be no difficulty in identifying the correct points.) If now the 10 kc/s. MV is turned on, these heterodynes should be heard spaced 10 kc/s. apart; also if the response of the headphones is good enough, a 10 kc/s. note will be heard.

The next step is to identify one of the 100 kc/s. points on the dial, and in order to do this use is made of a roughly calibrated receiver. Having identified one such point and noted the range and dial reading, it is a simple matter to check the remaining harmonics in the detector stage, and to identify them. To complete the calibration the dial readings of the 10 kc/s. harmonics on each range should be noted. An example of the method suggested is shown in Table 2 and its use will next be explained. It should be mentioned that in the method adopted there is no necessity for these readings to be taken to nearer than one division; it is only required to be able to identify each 10 kc/s. harmonic.

Use of the Apparatus

1. To measure the frequency of a signal heard on a receiver.

Switch on the VFO and tune it through the ranges until a signal from it is heard on the receiver. Tune this until it is "zero beating" with the signal of which the frequency is required, and note the exact dial reading on C_{14}. Let this be x. Switch on the 10 kc/s. MV, and by listening to the output of the detector obtain the exact dial readings of the 10 kc/s. harmonics immediately above and below x, and identify them from the calibration chart. Let these be y and z respectively. Since only a small part of the swing of the condenser is used in these measurements the frequency change between y and z may be assumed to be on a straight line, and hence we can interpolate for x thus:—

$y - z$ divisions = 10 kc/s.

so $x - z$ divisions $= \dfrac{10\,(x - z)}{y - z}$ kc/s.

If this be evaluated and added to the frequency of z, the fundamental frequency is obtained. A knowledge of the approximate frequency of the signal being measured will enable the operator to determine the harmonic being used, and so find the correct frequency. Should the approximate frequency not be known, if the above procedure is repeated and the harmonic of another fundamental frequency found to beat with the unknown signal, an inspection of the frequencies of the harmonics will show the correct frequency. For example, supposing the unknown signal is heterodyned by a harmonic of a frequency determined as 1,902 kc/s., and also by one found to be 2,536 kc/s. The harmonics of these two frequencies are set out thus:—

	kc/s.	kc/s.
Fundamental	1,902	2,536
× 2	3,804	5,072
× 3	5,706	7,608
× 4	7,608	10,144
× 5	9,510	12,680

from which it is obvious that 7,608 kc/s. is the frequency of the signal being measured.

2. To measure the frequency of a transmitter.

This is done in exactly the same way as above except that the first measurement is made by listening on the unit with only the VFO switched on, and adjusting it until it is heterodyned with the transmitter carrier.

3. A series of 100 kc/s. and 10 kc/s. points are always available for the calibration of receivers and other apparatus, by using the SFO only. A further series of standard signals at 2 Mc/s. and harmonics thereof is, of course, obtainable by setting the VFO to 2 Mc/s.

4. To set a receiver or transmitter to a given frequency.

The foregoing procedure is reversed. Suppose, for example, it is required to set a receiver to 7,431 kc/s. This is a harmonic of 3,715.5 kc/s. and of 2,477 kc/s. The latter frequency falls within the range of the VFO, and is found by noting the exact dial readings of 2,470 and 2,480 kc/s., using the MV. Suppose that these are a and b respectively.

Then 10 kc/s. $= b - a$ divisions.

and 7 kc/s. $\doteq \dfrac{7(b - a)}{10}$ divisions.

This figure, added to a, will give the dial reading of 2,477 kc/s. This procedure may then be repeated using the fourth harmonic of 1,857.75 kc/s., to confirm and to give greater accuracy.

TABLE 1.—Frequencies covered by each range on S3 on the original model

Range	Frequencies (kc/s.)	Range	Frequencies (kc/s.)
1	1,460—1,490	7	1,830—1,940
2	1,490—1,545	8	1,935—2,065
3	1,545—1,600	9	2,050—2,205
4	1,590—1,675	10	2,200—2,400
5	1,675—1,760	11	2,390—2,650
6	1,750—1,840	12	2,600—2,950

TABLE 2.—Example of calibration chart, for Range 6

Dial	Frequency	Dial	Frequency
5	1,750	56	1,800
15	60	66	10
26	70	76	20
36	80	86	30
46	90	96	40

The Influence of Illumination on the Fatigue of Photo-Electric Cells

By A. SOMMER, Dr.Phil.*

THE users of photo-electric cells can be divided into pessimists and optimists. The pessimists are horrified if a photo-cell is not stored in a light-proof box, while the optimists are disappointed if they find a cell rapidly deteriorating when they expose it to such a strong illumination that a photo-electric current of several milliamps is passing through the cell. In the following a brief survey will be given of the possible causes for the fatigue effect and some figures will be suggested indicating the maximum illumination to which cells of various types can be safely exposed under various conditions.

It should be mentioned that a change of sensitivity under the influence of illumination is not in all cases towards lower values. It is often found that the photo-electric sensitivity increases during an initial period, but this rise is almost in every case followed by a decrease to values below the initial sensitivity. The word "fatigue" is used in this article for any change of sensitivity.

(a) Causes of Fatigue

Every fatigue effect indicates a change of either the chemical constitution or the physical structure of the photo-sensitive surface. Such changes can be due to one or more of the following three causes:

(1) The actual emission of the photo-electrons can change the properties of the sensitive surface layer. This effect may be caused by rearrangement of the molecules forming the surface layer or by the release of minute quantities of adsorbed gases. It is to be expected that such an effect becomes more noticeable with increasing illumination and consequently increased electron emission.

(2) If the photo-cell contains gas, positive ions are formed if a potential greater than the ionisation energy of the gas is applied to the anode. The impact of these ions on the (negative) photo-cathode tends to reduce the sensitivity. This effect is probably again due to changes of the surface structure, but it is much more evident than (1).

(3) If the photo-cell contains a gas other than the rare gases, an effect on the cathode will be produced even at anode potentials lower than the ionisation energy of the gas, because chemical reaction may take place with the very reactive alkali metals which are contained in all modern photo-cathodes. This effect becomes worse at high anode potential because the effect of ion bombardment, described under (2), will be added.

Of the three effects mentioned above (3) is by far the most dangerous and (1) is of least importance. In modern high quality commercial cells, effect (3) has been reduced to a negligible minimum. We can therefore confine our discussion to effects (1) and (2).

(b) Practical Considerations

It is possible to give some quantitative data about the fatigue effect for the following three cases:

(a) *Photo-cells exposed to light in open circuit.*

It can be stated quite definitely that illumination of the photo-electric cathode has no effect on the cathode if no photo-current is passing. On exposure to light a small number of photo-electrons will leave the cathode but the space charge set up by these electrons will prevent any further emission so that effect (1) cannot take place. If a cell should be found to deteriorate under these conditions it can be assumed that effect (3) is produced owing to a poor vacuum in the cell during the manufacturing process. Such a cell would probably have lost sensitivity to an equal extent if it had been kept in the dark for the same period.

There is, of course, one exception to this statement: the thermal radiation of the light source must not be so great as to heat the cell appreciably above room temperature. All cells are liable to deterioration at temperatures above 40° to 50° C.—a temperature which may be reached in strong sunlight.

(b) *Vacuum photo-cells in closed circuit.*

In vacuum cells we have only to reckon with effect (1). It appears that cells of the "Silver-Oxygen-Caesium" type are more constant under continuous illumination than "Antimony-Caesium" cells. To give an idea of the order of photo-current that can be taken for any length of time (*i.e.*, for many hours) it may be stated that the first type will stand 0.5 μA per sq. cm. of cathode surface (*i.e.*, 5 μA of an average cell of 10 sq. cm. cathode area is used), while with the "Antimony-Caesium" type .1 μA per sq. cm. should not be exceeded. The values are correspondingly higher for shorter periods, and for quick measurements currents of several hundred micro-amps can be taken from any vacuum cell of average dimensions.

(c) *Gasfilled photo-cells in closed circuit.*

In gasfilled cells effect (1) is the same as in vacuum cells. But in addition effect (2) has to be reckoned with, which means that gasfilled cells are subject to more severe restrictions. Again, the "Silver-Oxygen-Caesium" type is more stable than the "Antimony-Caesium" type, but even with the first type currents above .2 μA per sq. cm. may not be constant over long periods. It is difficult to make a more exact statement because strong illumination on a cell of low "gas amplification" will usually produce a more constant photo-current than a weaker illumination on a cell of higher "gas amplification," the actual photo-current being the same in both cases. For very short periods the safe maximum current is approximately 2 μA per sq. cm. This limitation of gasfilled cells is not too serious because in practice a vacuum cell can be used if such comparatively large photo-currents are expected.

In conclusion, two remarks should be added to the above statements. Firstly, all the figures given are only very approximate. Even among individual samples of what was previously in this article called a "high quality commercial cell," the constancy will vary a lot and many cells may be used with much larger photo-currents than those indicated above. Generally speaking, the given maximum figures are on the conservative side.

* Cinema Television, Ltd.

Aerial-Coupling Circuits

A Series of Data Sheets

Part IV.—R.F. Transformer Coupling

By S. W. AMOS, B.Sc.(Hons.), Grad.I.E.E.

THIS method of aerial coupling is perhaps the most common of all, probably because, by suitable design, it can be made to give a very good performance with respect to voltage gain, selectivity and constancy of reflected capacitance. This may be attributed to the fact that the performance of the circuit (given in Fig. 1(a)) is a function of two variables, namely, the inductance of the aerial transformer primary L_1 and the mutual inductance M between primary and secondary.

It is customary in practical circuits to make L_1 either very small (less than 50 μH generally) or else very large (greater than 1,000 μH), and so the value of L_1 has been made 30 μH, a value regarded as typical of small primary windings, and 2,000 μH, an average value for large primary windings. The curves which follow have been expressed in terms of the parameters k and M, which are defined by the usual relationship :—

$$k = \frac{M}{\sqrt{L_1 L_2}}$$

The electrical equivalent of Fig. 1(a) is given in Fig. 1(b) and the analysis of the circuit is carried out in three appendices at the end of the article. The three fundamental formulæ are as follows :—

Reflected capacitance

$$= \Delta C_2 = \frac{1}{\omega^2 L_2} \cdot \frac{k^2\omega^2 c L_1}{1 - \omega^2 c L_1(1 - k^2)}$$

(derived in Appendix I)

Voltage gain

$$= \frac{M}{C_2\left[r\left(j\omega L_2 + \frac{1}{j\omega C_2}\right) + R_2\left(j\omega L_1 + \frac{1}{j\omega c}\right)\right]}$$

(derived in Appendix II)

in which $C_2 = \frac{1}{\omega^2 L_2} - \Delta C_2$

Selectivity factor (Appendix III)

$$= \frac{1}{1 + \frac{k^2 L_1 \omega Q r}{\left(\omega L_1 - \frac{1}{\omega c}\right)^2}}$$

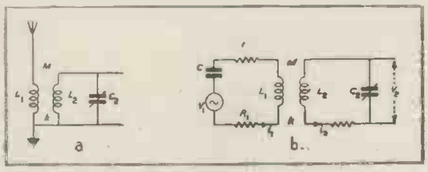


Fig. 1. The circuit and its equivalent.

As given above, these three expressions apply when the primary inductance L_1 is small so that R_1, its effective R.F. resistance (which is likely to be only a few ohms), may be neglected in comparison with r, the aerial resistance which, as in earlier data sheets, is assumed constant at 40 ohms. For large primary coils R_1 will not be negligible compared with r; in fact, it is likely at some frequencies to be greater than r and so, in the three expressions above, one should substitute $(r + R_1)$ for r in order to get expressions appropriate to the high inductance primary. As R_1 will vary with frequency, the Q of L_1, as for L_2, has been assumed constant at 100. We therefore get the values of R_1 given in Table I at various medium-wave frequencies. These values will be used in subsequent calculations.

TABLE I

Value of ω ($\times 10^6$ rads/sec.)	3	4	5	6	7	8	9	10	11	12
Value of frequency (kc/s.)	477	636	785	954	1,113	1,272	1,431	1,590	1,749	1,908
Value of R_1 (ohms)	60	80	100	120	140	160	180	200	220	240

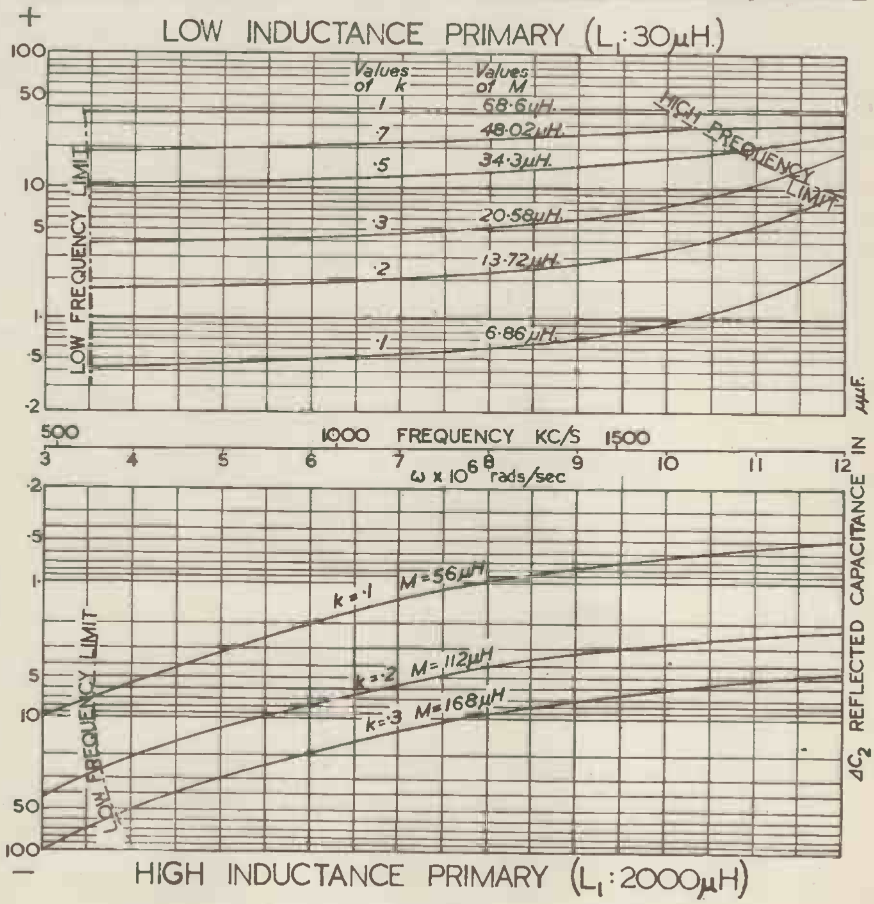
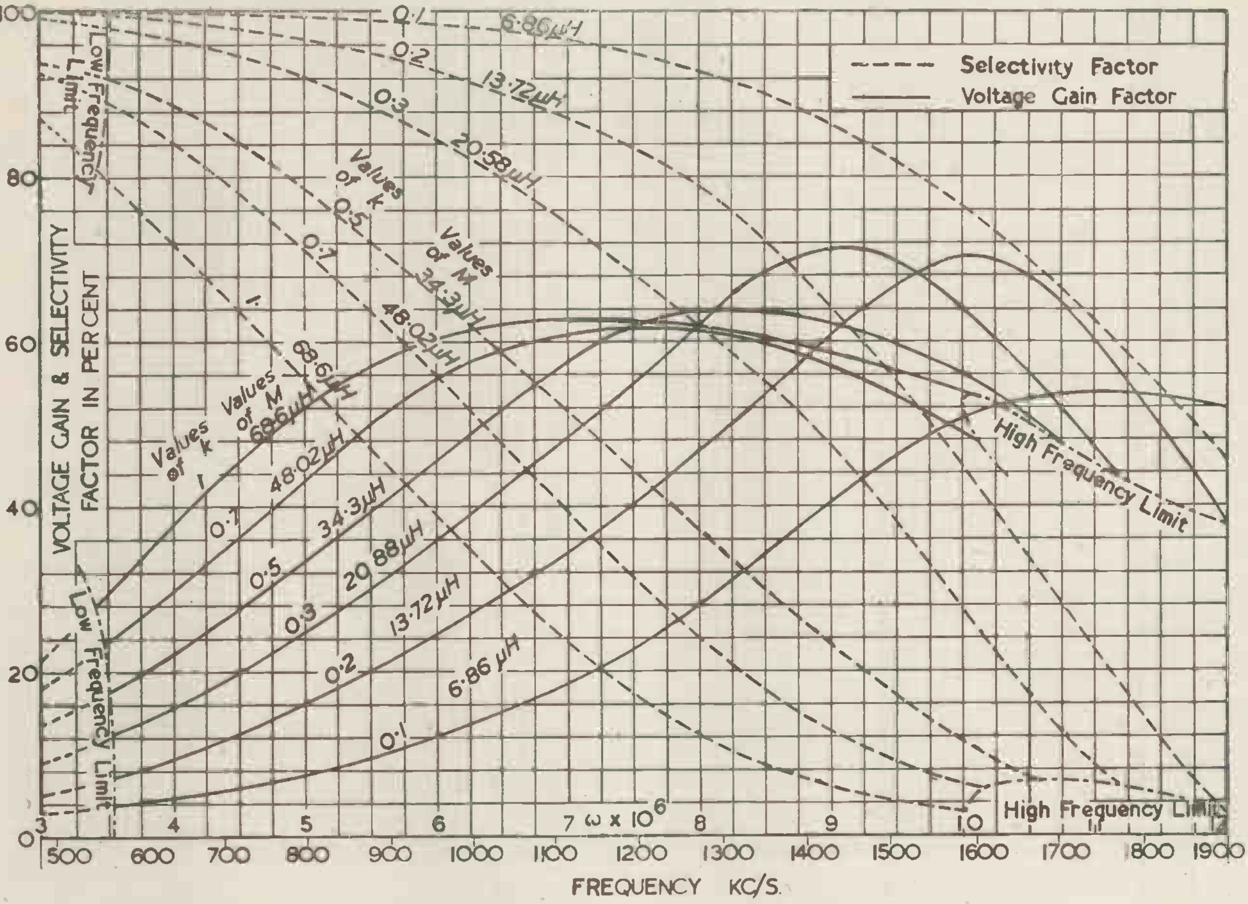


Fig. 2. Reflected capacitance against frequency for various values of k.

Voltage Gain and Selectivity for Low Inductance Primary (*Fig. 3*)



Reflected Capacitance

Examination of the expression for ΔC_2 given above shows the following points. If L_1 is so small that $\omega^2 cL_1(1 - k^2)$ is very much less than 1, then ΔC_2 is positive so that the capacitance in the secondary tuned circuit is increased by the connexion of the aerial. This occurs when $L_1 = 30\ \mu$H. On the other hand, if L_2 is large (2,000 μH), then $\omega^2 cL_1(1 - k^2)$ usually greatly exceeds unity so that ΔC_2 is negative. We can, in fact, neglect unity in comparison with $\omega^2 cL_1(1 - k^2)$ for this case, and ΔC_2 is given by

$$\Delta C_2 \approx -\frac{1}{\omega^2 L_2} \cdot \frac{k^2}{1 - k^2}$$

As an example of the use of the expression for the reflected capacitance, we will calculated the value of ΔC_2 for $L_1 = 30\ \mu$H, k = 0.3 and for $\omega = 8 \times 10^6$ rads/sec.:

In the same way, the value of ΔC_2 at other frequencies and for other values of k may be calculated. The results are exhibited in graphical form in Fig. 2 for k = 0.1, 0.2 0.3, 0.5, 0.7 and 1 for the small primary windings, and for k = 0.1, 0.2 and 0.3 for the high inductance winding. The frequency extremes of the waveband likely to be received with an average tuning condenser of 500 $\mu\mu$F maximum capacitance are also indicated here. In the determination of these boundary frequencies the total stray capacitance, together with the minimum capacitance of the condenser, is assumed to be 30 $\mu\mu$F. The maximum capacitance is assumed to be 510 $\mu\mu$F, plus reflected capacitance.

Voltage Gain

We will calculate the value of $\frac{V_2}{V_1}$ for $L_1 = 2{,}000\ \mu$H, k = 0.3 and $\omega = 8 \times 10^6$ rads/sec. From Fig. 2 we can see that ΔC_2 for this particular case is $-10.5\ \mu\mu$F. The calculation is set out in steps below.

$R_1 = 160$ ohms (from table on p. 505).

$R_2 = 12.56$ ohms (from table in Part I)

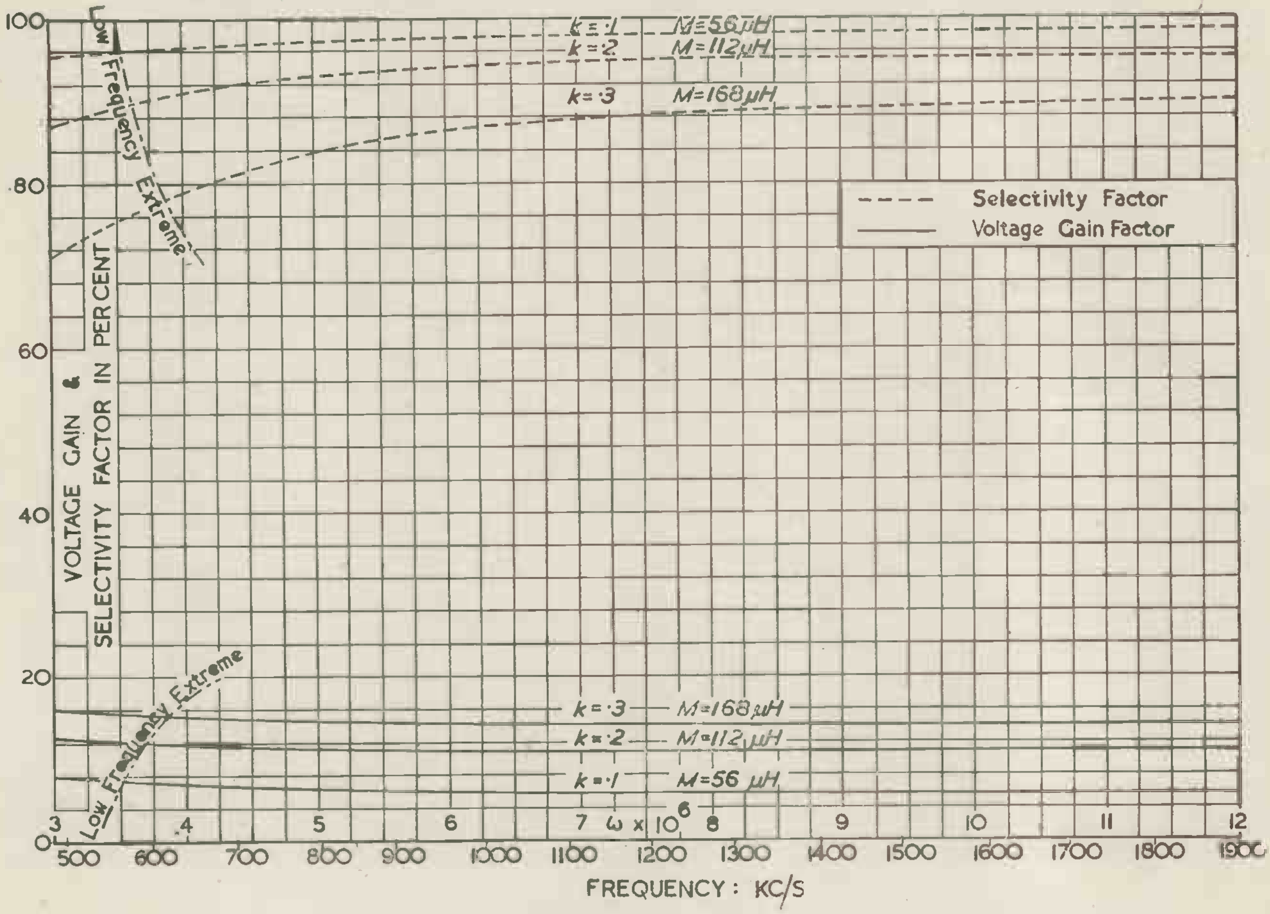
$$\frac{1}{\omega^2 L_2} = \frac{1}{8^2 \times 10^{12} \times 157 \times 10^{-6}} = 99.5\ \mu\mu\text{F}$$

$$C_2 = \frac{1}{\omega^2 L_2} - \Delta C_2 = 99.5 - (-10.5) = 109.5\ \mu\mu\text{F}$$

$$\omega L_2 = 8 \times 10^6 \times 157 \times 10^{-6} = 1{,}256 \text{ ohms}$$

$$\Delta C_2 = \frac{1}{8^2 \times 10^{12} \times 157 \times 10^{-6}} \cdot \frac{.3^2 \times 8^2 \times 10^{12} \times 200 \times 10^{-12} \times 30 \times 10^{-6}}{1 - 8^2 \times 10^{12} \times 200 \times 10^{-12} \times (1 - 0.3^2)} = 99.5 \times \frac{0.09 \times 0.384}{1 - 0.384 \times 0.91} = 5.28\ \mu\mu\text{F}$$

Voltage Gain and Selectivity for High Inductance Primary (Fig. 4)



$$\frac{1}{\omega C_2} = \frac{1}{8 \times 10^6 \times 110 \times 10^{-12}} = 1{,}136 \text{ ohms}$$

$$\omega L_1 = 8 \times 10^6 \times 2{,}000 \times 10^{-6} = 16{,}000 \text{ ohms}$$

$$\frac{1}{\omega c} = \frac{1}{8 \times 10^6 \times 200 \times 10^{-6}} = 625 \text{ ohms}$$

$$\frac{V_2}{V_1} = \frac{0.3 \times \sqrt{157 \times 2{,}000 \times 10^6}}{110 \times 217{,}000} = 7.04$$

Similarly, the values of $\frac{V_2}{V_1}$ can be calculated for other values of k and ω and also for the small value of L_1 (30 μH).

In Figs. 3 and 4, the solid curves illustrate the variation of voltage gain factor (*i.e.*, twice voltage gain—as explained in Part I). The frequency limits are also given here. The extreme constancy of the voltage gain for the high inductance primary is noteworthy.

Selectivity Factor

We shall calculate the selectivity factor for the small primary coil when k = 0.3 and ω = 8 × 10⁶ rads/sec. Using the expression given in the Introduction, we have

$$\omega L_1 = 8 \times 10^6 \times 30 \times 10^{-6} = 240 \text{ ohms}$$

$$\frac{1}{\omega c} = \frac{1}{8 \times 10^6 \times 200 \times 10^{-12}} = 625 \text{ ohms}$$

Selectivity factor

$$= \frac{1}{1 + \dfrac{0.09 \times 30 \times 8 \times 100 \times 40}{385^2}}$$

The curves showing the variation of the selectivity factor with frequency are shown by the dotted lines for the low inductance primary in Fig. 3 and for the high inductance case in Fig. 4.

Appendix I

Reflected Capacitance

Applying Kirchhoff's laws to Fig. 1(b), we have

$$V_1 = i_1 Z_p - j\omega M i_2 \ldots\ldots\ldots \quad (1)$$

in which $Z_p = R_p + jX_p = (r + R_1) + j\omega L_1 + \dfrac{1}{j\omega c}$

and $0 = i_2 Z_s - j\omega M i_1 \ldots\ldots\ldots \quad (2)$

in which $Z_s = R_s + jX_s = R_2 + j\omega L_2 + \dfrac{1}{j\omega C_2}$

From (2) $i_1 = \dfrac{Z_s}{j\omega M} \cdot i_2$

Substituting for i_1 in (1)

$$V_1 = \frac{Z_p Z_s}{j\omega M} i_2 - j\omega M i_2$$

Since $V_2 = \dfrac{i_2}{j\omega C_2}$

$$\frac{V_2}{V_1} = \frac{M}{C_2(Z_p Z_s + M^2\omega^2)}$$

$$= \frac{M}{C_2 Z_p\left(Z_s + \dfrac{M^2\omega^2}{Z_p}\right)} = \frac{M}{C_2 Z_p Z} \qquad \ldots\ldots (3)$$

in which Z represents the impedance of the secondary circuit in the presence of the primary. Rationalising Z we have

$$Z = Z_s + \frac{M^2\omega^2}{Z_p}$$

$$= R_s + jX_s + \frac{M^2\omega^2}{R_p + jX_p}$$

$$= R_s + jX_s + \frac{M^2\omega^2(R_p - jX_p)}{R_p^2 + X_p^2}$$

$$= R_s + \frac{M^2\omega^2 R_p}{R_p^2 + X_p^2} + jX_s - j\frac{M^2\omega^2 X_p}{R_p^2 + X_p^2}$$

By neglecting resistive terms in comparison with reactive ones

$$Z = R_s + \frac{M^2\omega^2 R_p}{X_p^2} + jX_s - j\,.\,\frac{M^2\omega^2}{X_p}$$

At resonance, the reactive terms vanish, giving

$$jX_s - j\,.\,\frac{M^2\omega^2}{X_p} = 0$$

$$i.e.,\ M^2\omega^2 = X_p X_s \qquad \ldots\ldots\ldots\ldots\ldots (4)$$

The value of Z at resonance is thus given by

$$Z = R_s + \frac{M^2\omega^2 R_p}{X_p^2} \qquad \ldots\ldots\ldots\ldots (5)$$

Substituting for X_p and X_s in (4)

$$M^2\omega^2 = \left(\omega L_1 - \frac{1}{\omega c}\right)\left(\omega L_2 - \frac{1}{\omega C_2}\right)$$

$$\therefore M^2\omega^2 = \omega^2 L_1 L_2 + \frac{1}{\omega^2 c C_2} - \frac{L_2}{c} - \frac{L_1}{C_2}$$

Remembering that $M = k\sqrt{L_1 L_2}$ and rearranging,

$$\omega^2 L_1 L_2(1 - k^2) + \frac{1}{\omega^2 c C_2} - \frac{L_2}{c} - \frac{L_1}{C_2} = 0$$

Multiplying throughout by $\omega^2 c L_2$ and rearranging,

$$C_2 = \frac{1}{\omega^2 L_2}\,.\,\frac{1 - \omega^2 c L_1}{1 - \omega^2 c L_1(1 - k^2)}$$

In the absence of a primary coupling, $C_2 = \dfrac{1}{\omega^2 L_2}$, so that the reflected capacitance $\triangle C_2$ is given by

$$\triangle C_2 = \frac{1}{\omega^2 L_2} - \frac{1}{\omega^2 L_2}\,.\,\frac{1 - \omega^2 c L_1}{1 - \omega^2 c L_1(1 - k^2)}$$

$$= \frac{1}{\omega^2 L_2}\,.\,\frac{k^2\omega^2 c L_1}{1 - \omega^2 c L_1(1 - k^2)} \ldots (6)$$

Appendix III

Selectivity Factor

Expression (5) has already shown that, at resonance, the resistance of the secondary circuit in the presence of the primary is greater than the normal amount R_s by the amount $\dfrac{M^2\omega^2 R_p}{X_p^2}$. Substituting for R_p and X_p and remembering the general formula for the selectivity factor given in Part I, we have

Selectivity factor

$$= \frac{1}{1 + \dfrac{1}{R_2}\,.\,\dfrac{M^2\omega^2(r + R_1)}{\left(\omega L_1 - \dfrac{1}{\omega c}\right)^2}}$$

$$= \frac{1}{1 + \dfrac{k^2 L_1 Q\omega(r + R_1)}{\left(\omega L_1 - \dfrac{1}{\omega c}\right)^2}} \qquad \ldots\ldots (9)$$

For a small secondary coil, we are justified in neglecting R_1 in comparison with r, as explained earlier, whereas, for the large primary coil, it is better to use Expression (9).

Appendix II

Voltage Gain at Resonance

Substituting in (3) the value of Z given in (5)

$$\frac{V_2}{V_1} = \frac{M}{C_2 Z_p\left(R_s + \dfrac{M^2\omega^2 R_p}{X_p^2}\right)} \approx \frac{M}{C_2\left(Z_p R_s + \dfrac{M^2\omega^2 R_p}{X_p}\right)}$$

But, from (4), $\dfrac{M^2\omega^2}{X_p} = X_s \qquad \therefore \dfrac{V_2}{V_1} = \dfrac{M}{C_2(Z_p R_s + X_s R_p)}$

Substituting for Z_p, R_s, X_s and R_p, $\dfrac{V_2}{V_1} =$

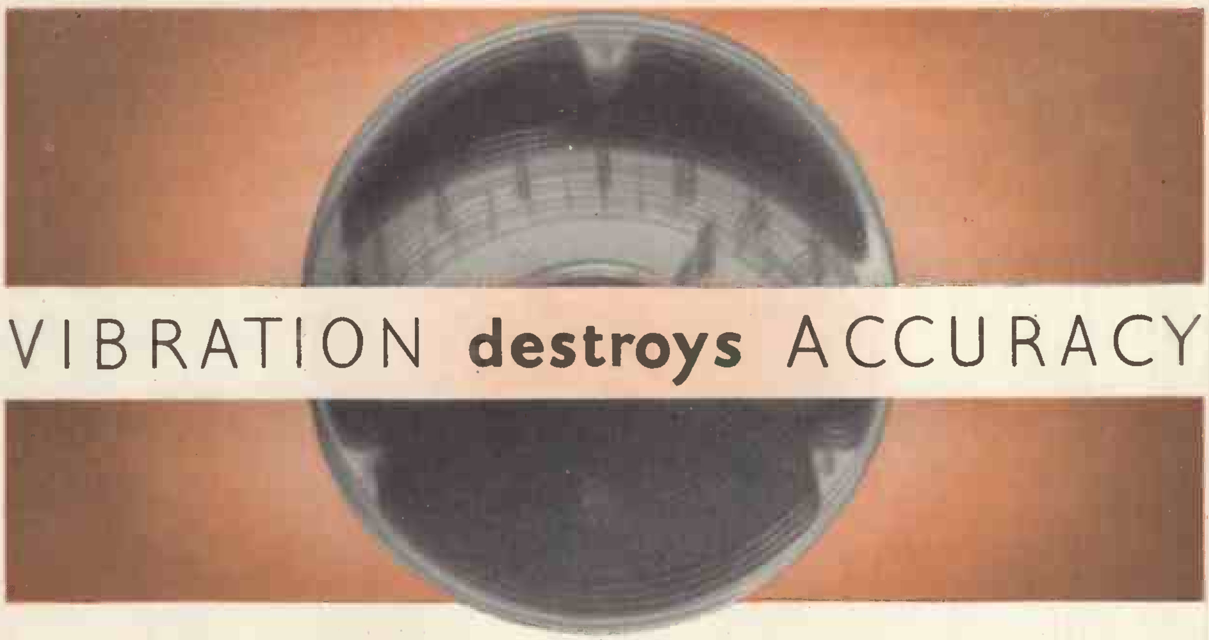
$$\frac{M}{C_2\left[R_2\left(r + R_1 + j\omega L_1 + \dfrac{1}{j\omega c}\right) + (r + R_1)\left(j\omega L_2 + \dfrac{1}{j\omega C_2}\right)\right]}$$

In evaluations of $\dfrac{V_2}{V_1}$ for the small primary coil it will be possible to neglect R_1 in comparison with r and $(r + R_1)$ in comparison with $j\omega L_1 + \dfrac{1}{j\omega c}$ and so we have

$$\frac{V_2}{V_1} \approx \frac{M}{C_2\left[R_2\left(j\omega L_1 + \dfrac{1}{j\omega c}\right) + r\left(j\omega L_2 + \dfrac{1}{j\omega C_2}\right)\right]} \qquad \ldots\ldots (7)$$

For the large primary coil R_1 will probably exceed r and the following version is then sufficiently accurate :—

$$\frac{V_2}{V_1} \approx \frac{M}{C_2\left[R_2\left(j\omega L_1 + \dfrac{1}{j\omega c}\right) + (r + R_1)\left(j\omega L_2 + \dfrac{1}{j\omega C_2}\right)\right]} \qquad \ldots (8)$$



Low Frequency Amplification

Part VII. Increasing Low Frequency Response

By K. R. STURLEY, Ph.D., M.I.E.E.

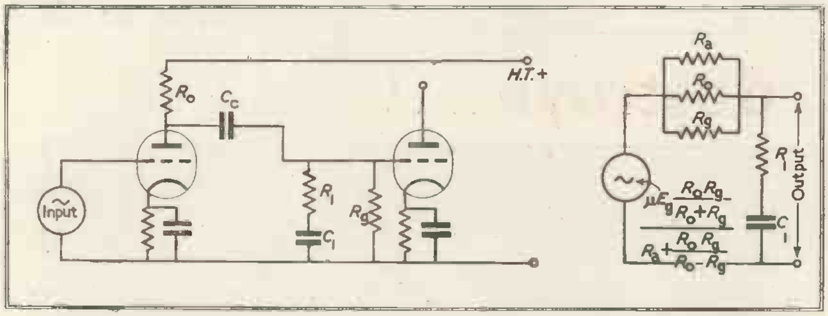


Fig. 41. (a) Circuit for increasing L. F. Response. (b) Equivalent circuit.

LOW frequency response can be increased in comparison with the response at medium frequencies in other ways than by the use of an anode decoupling circuit. An alternative form is shown in Fig. 41a; the grid leak resistance of the succeeding valve is paralleled by a series combination of resistance and capacitance, so proportioned that the reactance of the capacitance at medium frequencies is negligible in comparison with the resistance, which is effectively in parallel with the grid leak resistance R_g and the anode load resistance R_o. The Thévenin equivalent (*see* the Appendix to Part VI) is shown in Fig. 41b; R_1 and C_1 are placed on the anode side of C_c in order that C_c may not affect the performance of the R_1C_1 circuit. C_c is assumed to have a reactance much less than R_g and is neglected.

Amplification at medium frequencies is

$$A_m = \frac{\mu R_o' R_1}{(R_a + R_o')\left[\frac{R_a R_o'}{R_a + R_o'} + R_1\right]}$$

where $$R_o' = \frac{R_o R_g}{R_o + R_g} \quad \ldots\ldots\ldots\ldots 44$$

Amplification at low frequencies is

$$A_1 = \frac{\mu R_o'}{R_a + R_o'} \cdot \frac{R_1 + \frac{1}{j\omega C_1}}{\frac{R_a R_o'}{R_a + R_o'} + R_1 + \frac{1}{j\omega C_1}}$$

The amplification ratio is

$$\frac{A}{A_m} = \frac{\left(R_1 + \frac{1}{j\omega C_1}\right)\left[\frac{R_a R_o'}{R_a + R_o'} + R_1\right]}{\left[\frac{R_a R_o'}{R_a + R_o'} + R_1 + \frac{1}{j\omega C_1}\right]R_1}$$

$$= \frac{(1 + j\omega C_1 R_1)\left[\frac{R_a R_o'}{R_a + R_o'} + R_1\right]}{\left[1 + j\omega C_1\left(\frac{R_a R_o'}{R_a + R_o'} + R_1\right)\right]R_1}$$

$$= \frac{\frac{\frac{R_a R_o'}{R_a + R_o'} + R_1}{R_1} + j\omega C_1\left[\frac{R_a R_o'}{R_a + R_o'} + R_1\right]}{1 + j\omega C_1\left[\frac{R_a R_o'}{R_a + R_o'} + R_1\right]} \quad \ldots 46a$$

$$= \frac{B + j\frac{R}{X}}{1 + j\frac{R}{X}} \quad \ldots\ldots\ldots\ldots 46b$$

where $$B = \frac{\frac{R_a R_o'}{R_a + R_o'} + R_1}{R_1} = \frac{R}{R_1}$$

$$R = \frac{R_a R_o'}{R_a + R_o'} + R_1 \text{ and } X = \frac{1}{\omega C_1}$$

Low frequency increase in amplification is therefore 47

$$+ 20 \log_{10}\left|\frac{A_1}{A_m}\right| = + 10 \log_{10}\frac{B^2 + \left(\frac{R}{X}\right)^2}{1 + \left(\frac{R}{X}\right)^2}$$

Expressions 46b and 47 are identical in form to Expressions 27 and 26, given in Part V for the anode decoupling circuit. Hence the frequency and phase responses of this circuit are identical to those of the anode decoupling circuit, and therefore may be used to cancel attenuation and phase distortion due to cathode self-bias or to the screen decoupling circuit. The curves for frequency response, which are identical to those of Fig. 35 (Part V), are reproduced in Fig. 42, and to illustrate their use we shall assume the following valve and circuit constants:

$\mu = 100$, $R_a = 50{,}000\ \Omega$, $R_o = 200{,}000\ \Omega$, $R_g = 1\text{M}\ \Omega$, $R_1 = 40{,}000\Omega$, $C_1 = 0.1\ \mu\text{F}$, and C_c has a reactance negligible in comparison with R_g.

$$R_o' = \frac{R_o R_g}{R_o + R_g} = 166{,}666\ \Omega$$

$$\frac{R_a R_o'}{R_a + R_o'} = 38{,}400\ \Omega$$

$$R = 78{,}400\ \Omega \qquad B = \frac{R}{R_1} = 1.96$$

$$f_o = \frac{1}{2\pi R C_1} = 39.8\ \text{c/s.}$$

A close approximation to true frequency response is obtained by reading from the $B = 2$ curve after locating 39.8 c/s. on the frequency scale against $R/X = 1$, so that gains of + 5.3 db and +3.3 db are registered at 20 and 50 c/s. respectively.

A disadvantage possessed by this method of low frequency increase, in common with the anode decoupling circuit, is that it is obtained at the expense of medium frequency amplification and of reducing the A.C. load resistance. This leads to a low A.C./D.C. load resistance ratio, approximately $\frac{R_1}{R_o + R_1}$, with its possibility of harmonic distortion if attempts are made to get large output voltages from the stage (*see* the last section of Part III).

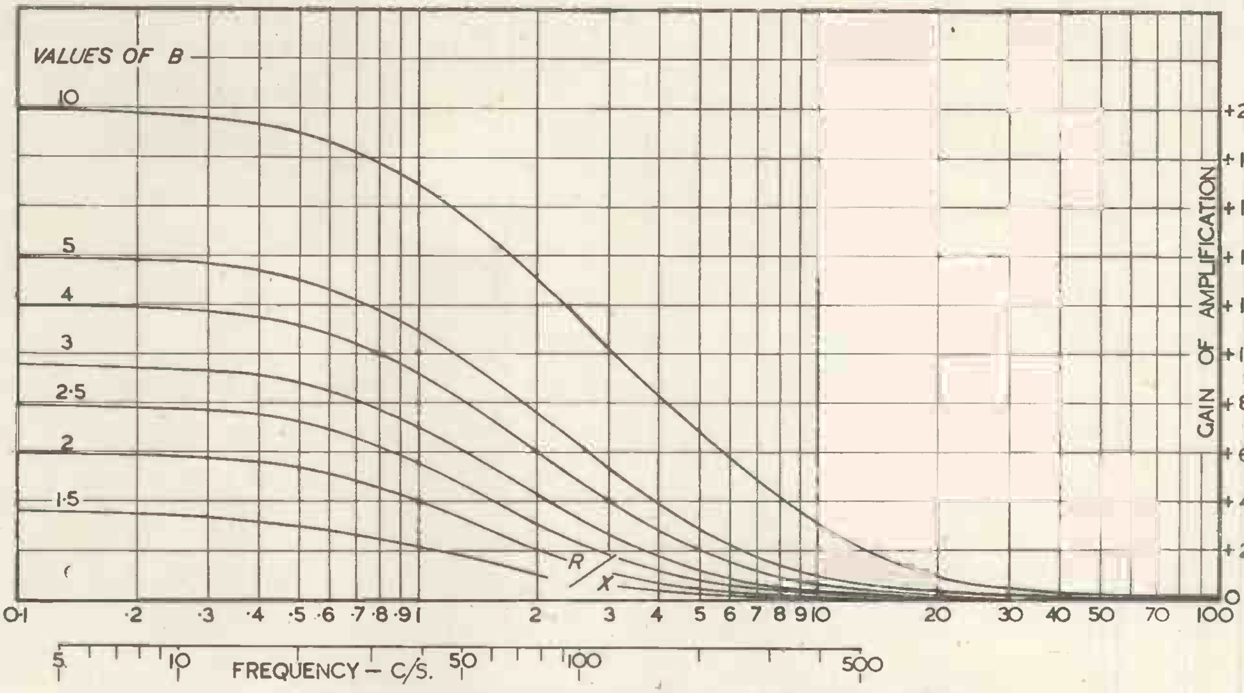


Fig. 42. Frequency response curve for circuit giving increased L.F. amplification.

Neutralisation of Low Frequency Phase Distortion due to the Coupling Capacitance C_c

Phase distortion, producing a variation in the time delay of different frequency components as they pass through an amplifier, is as a rule much more serious than attenuation distortion at low frequencies, and particularly is this true of television reception. As stated in Part I, low frequency phase distortion tends to produce a variable background to a received picture. If we consider the transmission of a half-white, half-grey picture, the voltage wave shape corresponding to this is the square wave shown in Fig. 43a, having a fundamental frequency equal to that of the frame, *viz.*, 50 c/s. The grid coupling capacitance C_c and leak resistance R_g to a vision frequency amplifier stage may cause phase distortion, tilting the horizontal part of the wave shape in Fig. 43a as shown in Fig. 43b. From A to B (Fig. 43b) the background shades gradually from white to less white, from C to D it is black and from D to F it shades gradually from black to grey, *i.e.*, there is not only a variable background but also a wide black line at what should be the sharp transition from white to grey. The wave can also be given a tilt in the opposite direction as in Fig. 43c, but this condition is encountered to a less extent in practice. The downward tilt from A to B in Fig. 43b can be explained by the fact that the coupling capacitance C_c is building up a charge reducing the voltage which drives current through R_g at the flat positive top of the input voltage applied to C_c and R_g. The discharge current I through R_g decays exponentially from an initial current I_o according to the law $I = I_o\epsilon^{-t/R_gC_c}$ and the line AB in Fig. 43b is actually part of an exponential curve $E = E_o\epsilon^{-t/R_gC_c}$ where E_o is the initial voltage amplitude OA and equals I_oR_g. Similarly during the "negative" period of the square wave, the capacitance C_c discharges through R_g, and the voltage decays according to the law $E = E_o\epsilon^{-t/R_gC_c}$ where E_o = GC. Thus, if the square wave input has a fundamental frequency of 50 c/s., and C_c and R_g are 0.1 μF and 0.5 MΩ, the percentage fall in voltage from A to B (Fig. 43b) is $(1 - \epsilon^{-20/100}) \times 100$ per cent. = 18.2 per cent., because the time "t" from A to B is equal to one-half the fundamental period, *i.e.*, to 0.01 second. Attenuation distortion due to these component values is given by the ratio $\dfrac{R_g}{\sqrt{R_g^2 + \left(\dfrac{1}{\omega C_c}\right)^2}}$, so that the attenuation of the 50 c/s. fundamental component frequency is only 0.18 per cent., and is negligible. Phase angle displacement at 50 c/s. is $\tan^{-1}\dfrac{1}{\omega C_cR_g} = 3°38'$, which is equivalent to a time advance of $\dfrac{3°38'}{360° \times 50} = 201.5$ microsecs., and it is this time advance of the low frequency components of the square

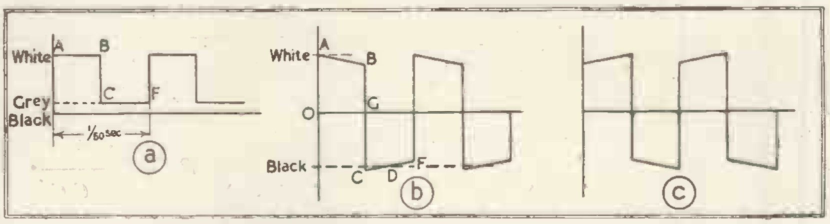


Fig. 43. Examples of phase distortion of square wave input voltage.

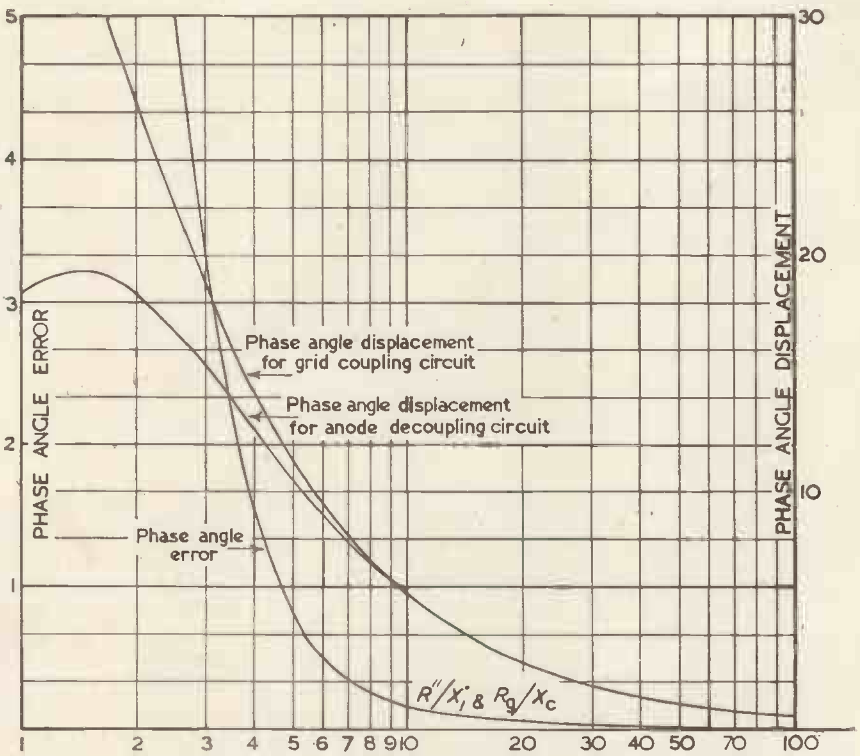


Fig. 45. Curves illustrating cancellation of phase distortion in grid circuit by reversed phase distortion in anode circuit.

wave which causes the downward tilt of the section AB and the upward tilt of the section CDF in Fig. 43b. The voltage tilt can be reduced by increasing C_c or R_g, and can, under certain conditions, be corrected by introducing a compensating tilt in the opposite direction (*see* Fig. 43c) from a circuit in some other part of the amplifier. The maximum value of R_g is limited to about 0.5 MΩ owing to considerations of valve life (*see* the last section of Part III on grid emission), so that only C_c can be increased. Making $C_c = 0.5$ μF reduces the fall in voltage to 3.92 per cent. at 50 c/s. and this would generally be regarded as satisfactory. If square wave input voltages having fundamental frequencies lower than 50 c/s. are to be amplified, still higher values of C_c would be required. When an extended high frequency response is needed the increased bulk of the larger value of C_c may be undesirable because it increases the input-earth capacitance of the amplifier to cause increased loss of high frequencies.

Under certain conditions the anode decoupling circuit, analysed in Part V, and the circuit for increasing low frequency response described at the beginning of this article can be used to provide the compensating tilt to cancel the fall AB in Fig. 43b. A square wave grid input voltage is caused by C_1 in association with R_o (*see* Fig. 44) to produce an output voltage wave shape similar to that shown in Fig. 43c, and, provided that R_1 is much greater (by about 10 times) than the reactance of C_1 at the fundamental frequency of the square wave, almost exact compensation of grid circuit tilt can be achieved. Analysis of Fig. 44, assuming $R_1 \gg \frac{1}{\omega C_1}$, shows that

$$\frac{E_o}{E_1} = g_m\left(R_o + \frac{1}{j\omega C_1}\right)\frac{R_g}{\left(R_g + \frac{1}{j\omega C_c}\right)}$$

$$= \frac{g_m R_g\left(R_o + \frac{1}{j\omega C_1}\right)\left(R_g - \frac{1}{j\omega C_c}\right)}{R_g^2 + \left(\frac{1}{\omega C_1}\right)^2}$$

$$= \frac{g_m R_g}{R_g^2 + \left(\frac{1}{\omega C_1}\right)^2}\left[R_o R_g + \frac{1}{\omega^2 C_1 C_c} - \frac{j}{\omega}\left(\frac{R_g}{C_1} - \frac{R_o}{C_c}\right)\right] \quad \ldots 48$$

The reactance (j) term becomes zero when $R_g C_c = R_o C_1$, and there is then no phase shift between the input and output voltage components. Phase distortion due to the grid circuit coupling capacitance and resistance can be completely cancelled by making the grid and anode circuit time constants equal when $R_1 \gg \frac{1}{\omega C_1}$.

If R_1 is comparable to $\frac{1}{\omega C_1}$, phase distortion can be reduced but not cancelled, except over a limited range of frequencies.

Another way of regarding the problem is to note that the phase angle displacement of C_c and R_g is positive whereas that due to the anode decoupling circuit or special circuit of Fig. 41a is negative, so that the phase angle displacement of the one may be used to cancel or reduce that from the other. For example, the phase angle displacement between the grid voltage E_g and input voltage E_1 of Fig. 44 is

$$\phi_g = \tan^{-1}\frac{\frac{1}{\omega C_c}}{R_g} = \tan^{-1}\frac{X_c}{R_g} \quad \ldots\ldots 49$$

Phase angle displacement for the anode decoupling circuit of Fig. 44 is (*see* Expression 28, Part V)

$$\phi_1 = \tan^{-1}\frac{-(B-1)\frac{R''}{X_1}}{B + \left(\frac{R''}{X_1}\right)^2} \quad 50a$$

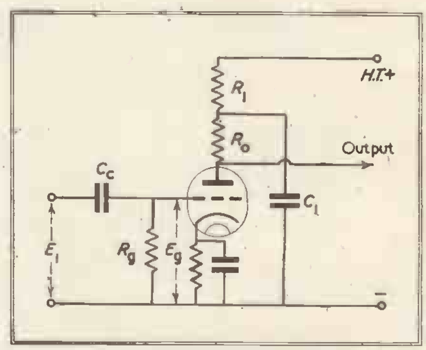


Fig. 44. Cancellation of phase distortion in grid circuit by opposing form of distortion in anode circuit.

$$\text{where } B = \frac{(R_1 + R_o)\left(\dfrac{R_a R_{g2}}{R_a + R_{g2}} + R_o\right)}{R_o\left(R_1 + R_o + \dfrac{R_a R_{g2}}{R_a + R_{g2}}\right)}$$

$$= \frac{(R_1 + R_o)(R_a + R_o)}{R_o(R_a + R_o + R_1)}$$

$$R'' = \frac{(R_a + R_o)R_1}{R_a + R_o + R_1}$$

(note that $R_{g2} = \infty$ in Fig. 44)

and $X_1 = \dfrac{1}{\omega C_1}$

When $\left(\dfrac{R''}{X_1}\right)^2 \gg B$, phase angle displacement becomes

$$\phi_1 = \tan^{-1} - (B - 1)\frac{X_1}{R''} \quad \text{.......... 50b}$$

so that by making

$$(B - 1)\frac{X_1}{R''} = \frac{X_c}{R_g} \quad \text{.................... 51a}$$

the phase angle displacement due to the grid circuit can be cancelled by that of the anode circuit for all frequencies which make $\left(\dfrac{R''}{X_1}\right)^2 \gg B$.

Expression 51a may be rewritten as

$$R''C_1 = R_g C_c(B - 1) \quad \text{............ 51b}$$

and arbitrarily selecting the following relationship for the lowest frequency for negligible phase distortion:

$$\left(\frac{R''}{X^1}\right)^2 = (2\pi f R'' C_1)^2 = 100B$$

$$f = \frac{10\sqrt{B}}{2\pi R'' C_1} \quad \text{........................ 52}$$

Replacing $R''C_1$ by $R_g C_c(B - 1)$ we find that the lowest frequency which can be amplified with negligible phase distortion is

$$f_o = \frac{10\sqrt{B}}{2\pi(B - 1)R_g C_c} \quad \text{............. 53}$$

Thus if $R_g = 0.5$ MΩ and $C_c = 0.5\,\mu$F, and $B = 2$

$$f_o = \frac{1.59 \times 1.414}{0.5 \times 0.5} = 9 \text{ c/s.}$$

In Fig. 45 are plotted curves of ϕ_g (Expression 49) and ϕ_1 (Expression 50a) against $\dfrac{R_g}{X_c}$ and $\dfrac{R''}{X_1}$ for $B = 2$. The horizonal logarithmic scales for $\dfrac{R_g}{X_c}$ and $\dfrac{R''}{X_1}$ are the same because Expression 51a gives $\dfrac{X_1}{R''} = \dfrac{X_c}{R_g}$ for $B = 2$.

It will be seen that the two phase angle displacement curves are almost identical down to $\dfrac{R_g}{X_c} = \dfrac{R''}{X_1} = 10$, thus justifying the arbitrary assumption that $\left(\dfrac{R''}{X_1}\right)$ should be $10\sqrt{B}$ for negligible phase distortion. The phase angle error, the difference between ϕ_g and ϕ_1, is plotted as the dotted curve $(\phi_g - \phi_a)$, and at $\dfrac{R''}{X_1} = 10\sqrt{B} = 14.4$ the error is approximately 7 minutes, which at a frequency of 9 c/s. represents a time delay of $\dfrac{7 \times 10^6}{60 \times 360} = 324$ microseconds. Such a value of time delay is insignificant at 9 c/s., and a square wave input of this fundamental frequency would be reproduced substantially unchanged at the output.*

Curves for other values of B may be plotted as in Fig. 45, but the scales for $\dfrac{R_g}{X_c}$ and $\dfrac{R''}{X_1}$ will only be identical for $B = 2$; for example when $B = 4$, $\dfrac{R_g}{X_c} = 1$ is located with $\dfrac{R''}{X_1} = 3$. Expression 53 shows that for a fixed value of $R_g C_c$, f_o is decreased by increasing B, which means making R_1 greater in comparison with R_o. This confirms the simpler mathematical analysis culminating in Expression 48, which suggests zero phase distortion conditions independent of f when R_1 is much greater than X_1.

This method of compensation requires a comparatively large anode load resistance and is not suitable for correcting phase distortion in television V.F. amplifiers, which must have a low load resistance in order to secure extended high frequency response. For example, probable values of R_o and R_1 in a V.F. amplifier are 2,000Ω each, so that if $C_c = 0.5\,\mu$F, $R_g = 0.5$ MΩ and $B = 2$.

$$R''C_1 \approx R_1 C_1 = R_g C_c(B - 1) = 0.25$$

$$\text{Hence } C_1 = \frac{0.25 \times 10^6}{2{,}000} = 125\ \mu\text{F},$$

a value difficult to obtain in practice. If, however, R_o and R_1 can be made large, C_1 is reduced, *e.g.*, for $R_o = 100{,}000\Omega$ and $R_1 = 50{,}00\Omega$, $C_1 = 5\ \mu$F, a value which can be realised.

* This method of treating low frequency phase compensation has, so far as the author is aware, not been published before.

Metrosil

Metrosil is the trade name of a silicon carbide composition developed by the Metropolitan-Vickers Co., Manchester, which shows a marked departure from Ohm's law in its resistance properties.

With D.C. the current varies as the fourth or even the fifth power of the voltage, or, conversely, the voltage characteristic is given by $V = KI^b$, where V is the voltage across the material, K a constant depending on the thickness of the material, and b an index varying between 0.2 and 0.25.

Metrosil is supplied in the form of disks of varying diameter and thickness, each face of the disk being coated with metal to provide good contact with the external circuit.

For thin disks K = 60–70, and for thick disks K = 3,000–4,000. The above law holds between 1 milliampere and 10 amperes, and provided that the disks are not overheated the characteristics will not vary more than 2 per cent. for an indefinite period.

Metrosil is not appreciably affected by humidity except in extreme cases, and can be impregnated with wax as a preventative.

At very low current densities the material behaves as an ohmic resistance with a resistivity of about 10^9 ohms per in. per sq. in.

Its characteristics make it of particular importance as a surge suppressor, but other applications such as condenser protection and temperature compensation suggest themselves.

It can also be used to render a relay extremely responsive to a small voltage change.

—Metropolitan-Vickers Co., Manchester.

Plastics in High Frequency Insulation

By PAUL I. SMITH

PLASTICS have proved of considerable interest to electronic engineers for insulation at radio frequencies, and it is, therefore, of interest to discuss at some length the various materials suitable and their several characteristics.

The most important rigid dielectric is, of course, polystyrene, which possesses outstandingly low electrical power factor, high dielectric strength, great arcing resistance and low moisture absorption. Other materials of growing interest are aniline formaldehyde, certain specialised phenolic and melamine resins, all of which are rigid materials, and non-rigid plastics, such as polythene.

Polystyrene

In considering polystyrene for insulation at radio frequencies it is important to remember that this thermo-plastic solid formed by the polymerisation of monomeric styrene is capable of appreciable modification, *i.e.*, various polymeric substances with molecular weights 80,000 to 200,000 can be produced commercially, each with specific and characteristic electrical and mechanical properties, and these in turn may be suitably modified by plasticisation and processing, including physical treatment. Certain fabricating processes which impart flow during moulding improve the mechanical strength and shock resistance, and operations which stretch the polymer during cooling increase the impact strength and elongation before tensile break.

Since the commercial development of polystyrene resins a few years before the present war, it was recognised that owing to the absence of polar groups these high polymeric substances possessed very valuable dielectric properties, and that their amorphous and homogeneous character was superior to definite crystalline structures. On the other hand, it was soon found as the result of experience that polystyrene was difficult to fabricate owing to its inherent brittleness and that it had a tendency to craze and deteriorate on ageing. During the last few years, considerable progress has been made in this country and America in the development of less brittle and more stable types of polystyrene, and not only has the basic commercial grade been improved, but other new types have been produced, particularly flexible forms and grades with higher softening points, and possessing improved machinability.

Properties

The properties of polystyrene which are of particular importance to the radio and related industries may be summarised as follows:—

1. The dielectric constant varies little with frequency; there is very little dielectric absorption of polystyrene from zero frequency, even into the range of visible radiation. The dielectric loss of polystyrene at 1,000,000 cycles with a high voltage gradient across the sample was measured by H. H. Race and S. C. Leonard (*Elec. Eng.*, 55, 1347 (1936)) by means of a calorimetric method and they found a power factor of 0.04 per cent. as observed with lower voltage gradients.

The British material, *Distrene*, has the following properties:—

Frequency c/s.	S.I.C.	Loss Factor.
50	2.2	0.0002
1,000	2.2	0.0005
200,000	2.1	0.00044
650,000	2.1	0.0002
1,000,000	2.3	0.0002
40,000,000	2.3	0.0001

2. The moisture absorption of polystyrene is negligible, the figure for Distrene being nil for 48 hours. There is a slight deterioration in properties when moisture absorption does take place, but exposure to high humidity affects the properties of this dielectric less than water immersion, the latter being a condition less likely to be experienced in actual practice.

3. The thermal softening point (Martens) is 87° C., and the distortion temperature is 89-90° C. (ASTM. D48.37) Thermal expansion, 10^{-5} per ° C., is 7.2 (−10° to 45°). Cold flow is very low at room temperature and only appreciable at 60° C.

4. The mechanical properties of polystyrene are very satisfactory, and its relatively high Young's Modulus (modulus of elasticity, lb. per sq. in. $\times 10^6$ is 5.5) is an asset in such mouldings as must have dimensional stability in the presence of varying mechanical forces. The tensile strength of Distrene is 2.7-3.0 tons per sq. in.; compressive strength, 7 tons per sq. in. and impact strength, Izod, notched bar (ft. lb. 20° C.) 0.25-0.3.

5. The optical properties of polystyrene are exceptionally good. The light transmission of a 0.1 in. plate is 92 per cent. Refractive index η^{20}_{D} — 1.59. The material yellows slightly when exposed to U.V. light, but its stability under normal conditions is good. Like acrylic resin, polystyrene is capable of transmitting and carrying light through curved sections and use is made of this peculiar property in the production of aircraft instruments.

6. The lightness in weight of polystyrene is of considerable importance for a number of applications. The specific gravity varies slightly, 1.05 to 1.07, compared with 3.5 for Mycalex; 2.5 for steatite; 2.3 for porcelain; 2.8 for mica and 2.25 for glass.

7. The easy workability of polystyrene is an important factor from the production angle. The resin is available in powder form for moulding by the compression or injection methods, preferably the latter, and cast sheets, rods, bars and tubes rendered strain-free and stabilised, can be milled, turned and drilled without fear of crazing and cracking, provided special precautions are taken to guard against over-heating of the material. Polystyrene resin begins to soften at temperatures of 75° C. to 85° C. and a satisfactory supply of cooling and lubricating liquid must be available (soluble oil diluted with 20 parts of water). Standard milling machines and cutters can be used for working Distrene; speeds used should be similar to those for machining brass. Only thin turnings should be taken off in the lathe, otherwise chipping may result. Where small quantities of parts are required, or where simple shapes are needed, then fabrication of cast sheet, rod, bar or tube is recommended.

8. Parts moulded or machined may be cemented together with a polystyrene cement made by dissolving 5 parts of polystyrene, air polymerised form and 1.5 parts Aroclor 1254 in 93.5 parts of toluene.

This is claimed to be suitable also for a dope or impregnating solution for the coating of coils. High quality insulating finishes suitable for use in the high frequency field can be made by use of polystyrene, suitably plasticised, in solvent, such as benzene, toluene, xylene, solvent naphtha, carbon tetrachloride, etc., and the use of the monomeric styrene also possesses interesting possibilities as an additive to the solution of the polymer in solvent.

It is possible that a new dielectric may be produced by impregnation of veneers (hardwood) with a solution of the monomer and subsequent bonding and polymerisation in a hydraulic press. Some use is made of mica-filled polystyrene for use as temperatures higher than 70° C. and up to 85° C., but this filled material has a higher power factor and dielectric constant than pure polystyrene.

There is no doubt that polystyrene is a material of very considerable interest to radio manufacturers for normal production where the highest quality dielectric is required. In the evaluation of production costs of units moulded of polystyrene, it should be remembered that the use of this plastic permits considerable economies in both material and output. The specific gravity is the lowest of any plastic material; this means that more mouldings per lb. can be produced. Due to the free flowing properties of the British material at injection moulding temperatures, the moulding rate is approximately 25 per cent. faster than with other materials.

Polystyrene is also of interest to electronic engineers engaged on research as components can easily be machined and new designs developed very speedily. In this latter connection, however, it is worthwhile noting that owing to the relatively low softening point of this plastic coupled with its ability to burn at temperatures above 120° C., the consequences of a flash-over cannot be ignored. For experimental work in connexion with radio frequency heating, Mycalex* is sometimes preferred.

Before leaving polystyrene, it is useful to discuss quite briefly the trend of modern research and industrial development. As mentioned previously, stretched polystyrene can be produced and, in America, a flexible material has been on the market for more than two years. The electrical properties of the flexible grades are not quite as good as the rigid dielectric, for instance at 1,000,000 cycles the dielectric constant is 2.5-2.7 and power factor 0.0001-0.0004, but they are very superior insulating materials for radio work and appear to possess many interesting possibilities. New types of polystyrene have also been developed for use at higher temperatures than the normal grade, and one fairly recent American dielectric is claimed to withstand about 20° F. higher than ordinary polystyrene without deterioration of machining properties.

Aniline Formaldehyde Resin

This is a plastic which has come into prominence during the last five years or so for specialised electrical applications. Although a thermoplastic, its softening point is so high that, for temperatures up to 100° C. it can be classed as a thermo-setting resin. Two main types of moulding are available.

1. The pure aniline formaldehyde resin, *i.e.*, without filler. This is a rigid dielectric with a Young's Modulus E (lb. per sq. in. $\times 10^6$) 0.6, and Izod impact strength, 0.5. moisture absorption, practically nil.

2. Special paper in which the resin is precipitated in the pulp prior to the manufacture of the paper. This paper is pressed up to form boards and moulded rods which have good mechanical and electrical properties and can be readily moulded or machined. The tensile strength varies from 15,000 to 20,000 lb. per sq. in. and the impact strength, Izod, 0.8; water absorption is a little more than the pure resin, the figure being 0.4 per cent.

The power factor of the British material, " Panilax " is tan d at 50 c/s., 1 per cent. to 2.0 per cent. at 25° C., and 2.0 per cent. to 12 per cent. at 90° C.; tan d at radio frequencies, 0.15 to 0.3 per cent. at 25° C. and 0.2 per cent to 1.0 per cent. at 90° C.

Phenolic and Melamine Resins

Special grades of phenol formaldehyde moulding powders are being manufactured to obtain good electrical qualities at normal low frequencies, but generally speaking P.F. powders are not suitable for use at high frequencies. The cast phenolic resins are more suitable and the type, *Catalin* mica filled, is likely to assume some importance where the highest electrical properties are not specified. The surface resistivity, ohms per sq. cm. of the mica-filled material is 1.6×10^6; volume resistivity, ohms per cm., 1.9×10^6. Unfortunately, the power factor (1 Mc/s.) of this particular grade is not yet available, but the power factor (1 Mc/s.) of the high strength *Catalin* is 0.014 (this is not the electrical grade). *Melamine* urea resins are of interest for certain components requiring high arc resistance and high dielectric strength. The American melamine resin (asbestos filled) has a power factor (10^6 cycles) 0.041, but new grades show an improvement on this.

Non-Rigid or Flexible Dielectrics

Polythene is of exceptional interest as a rubber-like material specially suitable for use at all frequencies between 50 and 5×10^7 c/s. The name polythene is the generic name given to the solid polymers of ethylene discovered by Imperial Chemical Industries, Ltd. It is a white translucent substance, tough and flexible, although without the elasticity of natural rubber. The chemical inertness, water resistance and dielectric properties are all exceptionally good. Volume resistivity, 10^{17} ohms per cm. cub.; surface resistivity, 10^{14} ohm. per sq. cent. and breakdown voltage, 1,000 volts per mil. The dielectric constant (2.3 at 20° C. and all frequencies between 50 and 5×10^7 c/s.) has a tendency to fall with increasing temperature, *e.g.*, to 2.0 at 100° C. Polythene has a softening point of about 115° C. for the hardest grade and below the softening point this material is remarkably free from cold flow. As the temperature is lowered, polythene gradually stiffens and finally breaks on bending at about −20° C. This material is of principal interest to radio manufacturers on account of its availability in extruded forms, such as tubes, tapes, foils, threads.

Rubberlike plastics from petroleum hydrocarbons, such as *Isolene* (polyisobutylene) are now assuming importance for insulation, power factor tg at 800 cycles is 0.0004 at 20° C. and 0.0005 at 85° C. Polyisobutylene is completely unaffected by immersion in water. It retains its mechanical properties up to nearly 100° C. It is frequently used in association with polythene and other thermoplastics and is easy to extrude.

The American material, *Styraloy*, which is a new synthetic elastomer, is of interest because it shows good dielectric strength at elevated temperatures and surprising resistance to corona discharge at even high temperatures. The power factor appears, however, to rise rather abruptly at frequencies higher than one megacycle, and it may well have an absorption peak around 100 Mc/s. and then fall off again at higher frequencies. Thus at 50 Mcs. the factor is 0.40-0.50, compared with 0.05-0.10 per cent. at 1 megacycle and 0.35-0.40 at 50 Mc/s.

* A full account of this material is given in the May, 1942, issue of this Journal, p. 749.

The Effects of Electric Shock

By H. A. POEHLER *

This article is of particular interest in view of the increasing use of high voltages in electronic equipment, and we are indebted to *Electronics* for permission to reproduce it in condensed form.

DEATH by electricity is due to one of three fundamental causes: a cessation of respiration due to a block in the part of the nervous system controlling breathing; a serious reduction of the circulation of the blood, due to unco-ordinated beating of the heart muscles (ventricular fibrillation); or an overheating of the body. Of the three, the second of these is the most dangerous, for there is no practical way of bringing a fibrillating heart into a normal beat. Of course, death may be the result of a combination of the above causes, or due to complications, such as a broken neck, etc.

The reason for the wide variation in voltage required to send a lethal current through a human body is that the resistance of the body varies from 1,000 ohms when wet to 500,000 ohms when dry.

The resistance of the body is made up of the skin resistance and the internal resistance. The former is large when the skin is dry (70,000 to 100,000 ohms per sq. cm.), but falls to less than a hundredth of this value when wet. The internal resistance is low because the tendons, muscles and blood are relatively good conductors.

It is understandable that the effect a given current will have depends on the current path through the body. It is found that the heart, the brain, and the spinal column are the three most critical regions.

Numerous studies[1] have shown that the threshold of perception is 1 mA. In other words, currents less than 1 mA. are not even felt, provided abnormally large current densities, as result from pin-point contacts, are not produced.

Currents from 1 to 7 mA. are perceptible, but not yet painful. When the currents reach a value of 8 to 15 mA. they are painful, and cause an involuntary contraction of the muscles affected. Muscular control, however, can still be exercised. Currents of 15 to 20 mA. are painful, cause involuntary contraction, and muscular control is lost. Currents of 20 to 50 mA., passed between arms, or an arm and a leg, involve the chest muscles and breathing becomes difficult. Currents of 100 to 200 mA., when passed through the body in a path that involves the heart regions, produce ventricular fibrillation.

Currents in excess of 200 mA. produce burns; if they take a path involving the heart region, the heart action is suspended for the duration of the current passage, but generally is resumed at the end of this period.

If the path involves the part of the nervous system controlling respiration (such as hand to hand, hand to foot, head to hand, etc.) a block in the respiratory system is produced. If artificial respiration is applied, the body may resume its own breathing after as long as 8 hours; if the damage to the respiratory-controlling nervous system is severe, however, breathing may be suspended indefinitely.

The variation of the percentage of shocks causing fibrillation with the magnitude of the current passed through the body of a sheep is shown in Fig. 1.[2] Each point represents about 75 trials. Note that the susceptibility increases with current up to a maximum, and then decreases as the current is increased further. This is in agreement with observed data on man, for it has been observed that as the voltage increases on high-voltage shocks, the per cent. that can be resuscitated increases.

For shocks short in duration compared to a heart cycle, the probability of producing fibrillation varies with the part of the heart cycle in which the shock occurs. This is shown by the dash-dash curve superimposed on the electrocardiogram in Fig. 2. This sensitive phase represents the decreasing contraction of the heart muscles. At any other time, the heart is quite insensitive to shock.

Duration of Shock

Finally, the effect of shock length was studied. The results are plotted in Fig. 3. Note the sudden increase in susceptibility to fibrillation as the shock length approaches the length of the heart cycle. What happens to this curve as the shock length is decreased to much smaller values, say one microsecond, is an interesting question, but no authentic data are available on this subject.

In many cases of electric shock the victim becomes unconscious and stops breathing, but his heart keeps on beating. This is due to a break in the nervous system controlling respiration. The nerves are paralysed by the currents and no longer transmit stimuli to the lungs.

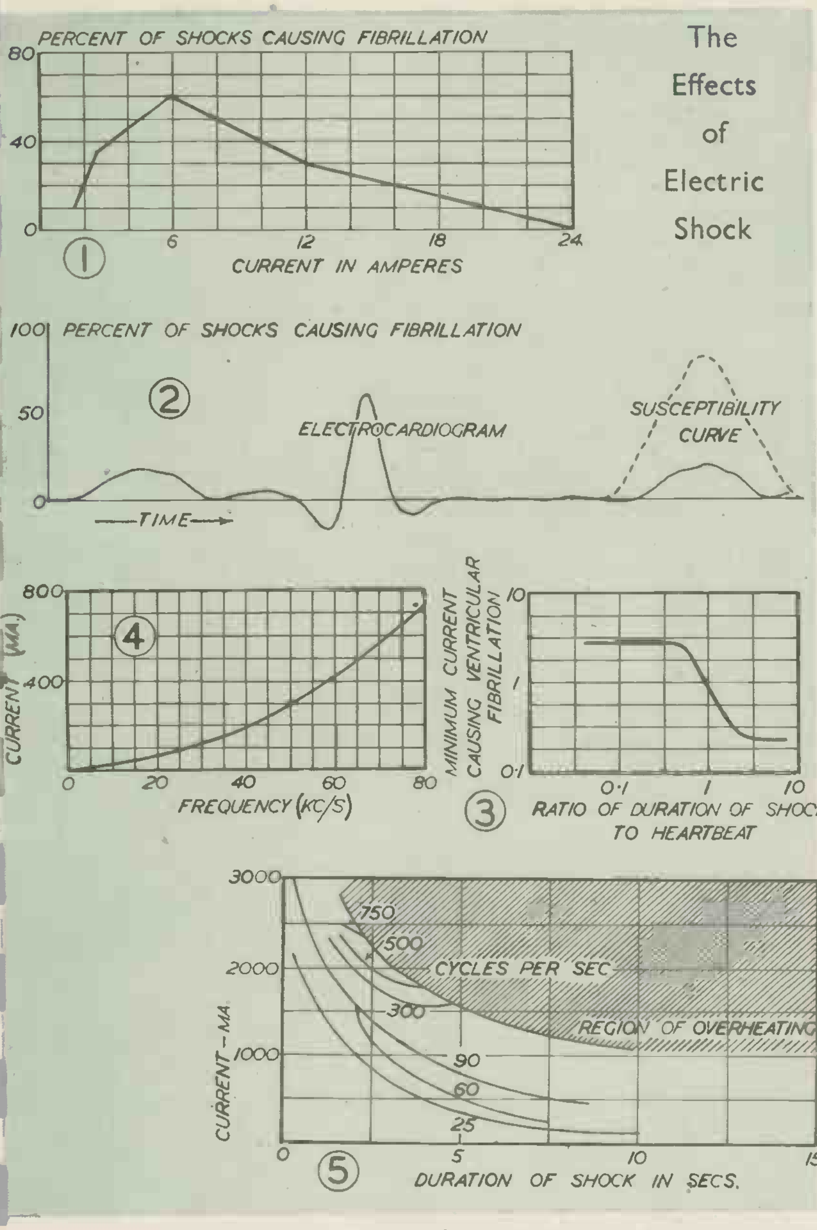
If the damage to the nervous system is not too severe, the block will pass away (0 to 8 hours) and the person will resume breathing of his own accord, provided the person has been kept alive by supplying the vital cells of the body with oxygen in the meantime through artificial respiration. This explains the prescribed procedure in all cases of electric shock: apply artificial resuscitation immediately and continue until *rigor mortis* sets in.

A further characteristic of current that determines its effect on an organism is its frequency. A ready example is that of direct current and 60-cycle alternating current. The bearable direct current is about three times that of the 60-cycle current. This problem has been studied from two angles; one, the maximum current which a person could stand before distress was caused, and second, the amount of current required to kill laboratory animals.

The results as well as those of W. Kouwenhoven, D. Hooker and E. Lotz[3] show that the frequencies that are the most dangerous are those in the neighbourhood of 60 c/s.

W. McLachlan[4] gives detailed information based on studies of 475 cases where electricity alone was the cause of death, not lack of resuscitation, broken necks, burns, etc. McLachlan's figures are based on U.S. and Canadian industrial accidents, and divide the accidents according to the potential of the circuit involved:—

* Westinghouse Elec. & Mfg. Co., Bloomfield, N.J.



Record of Accidents by Potential of Circuit Involved.

Volts.	Total Cases.	Successful Revival. per cent.
0—749	65	63
750—4,999	212	65
5,000—39,999 ...	167	69
0,000 and over ...	26	88

Note that the danger does not necessarily increase with the voltage. This is due to two reasons: first, the muscular reaction is more pronounced, at high voltages, making it more likely that the person will be thrown clear of the circuit; secondly, as data on animals have shown, the heart is not thrown into fibrillation by very large currents (greater than 250 mA.).

The data by McLachlan show that if resuscitation is instituted soon after the accident, the fatality can be reduced to 33 per cent.

REFERENCES

1 Dalziel, C. F., and Lagen, J. B., Effects of Electric Current on Man, *Elec. Eng.*, 60, No. 2, p. 63.

2 Ferris, L., King, B., Spence, P., and Williams, H., Effects of Electric Shock on the Heart, *Elec. Eng.*, 55, p. 498, 1936.

3 Kouwenhoven, W., Hooker, D., and Lotz, E., Electric Shock, Effects of Frequency, *Elec. Eng.*, 55, p. 384.

4 McLachlan, W., Electric Shock; Interpretation of Field Notes, *Jrl. Industrial Hygiene*, XII, No. 8, p. 291, Oct., 1930.

Electronics, Vol. 17, No. 7, p. 140, 1944.

NOTES FROM THE INDUSTRY

The English Electric Co.

At a recent Press luncheon, Sir George Nelson, the chairman of the English Electric Co., spoke of the company's activities in war-time, particularly in the building of Hampden bombers, tanks, diesel sets, and smaller electrical equipment, such as high frequency alternators, cathode-ray ignition testers, and radio transformers for Service requirements.

It is the company's policy to remain in the field of aircraft construction in addition to the resumption of their other well-known electrical activities.

They are particularly interested in the technical training of their employees and have a Central Education Officer at Stafford Works whose main work is to secure the continued maintenance of the high standard of British technicians.

T.E.M.A.

At the annual general meeting of the Telecommunication Engineering and Manufacturing Association, Mr. T. A. Eades was elected chairman and Mr. F. T. Jackson vice-chairman for the ensuing year. The formation of this Association has proved of considerable value to the industry which it represents and has amply justified the anticipations of its founders.

S.I.M.A.

The Bowen Trust for Prizes

Mr. W. Bowen, governing director of the Bowen Instrument Co. and Cables & Plastics, Ltd., has presented to the Scientific Instrument Manufacturers' Association a Trust Fund for the awarding of annual prizes of £25 each for:—

(a) A new invention.
(b) An improvement of design.
(c) An improvement in manufacturing technique.
(d) A new development of new process arising from research.

The competition is open to any employee of members of the S.I.M.A., and full particulars and terms can be obtained from the secretaries, Messrs. Binder, Hamlyn & Co., River Plate House, South Place, E.C.2.

R.C.A. Victor

The appointment of D. F. Schmit as Director of Engineering is announced by Frank M. Folsom, Vice-President in charge of the RCA Victor Division. Mr. Schmit, who was formerly assistant chief engineer, will fill the post vacated by Dr. J. B. Jolliffe, who recently was elected Vice-President of the Radio Corporation of America, in charge of RCA Laboratories.

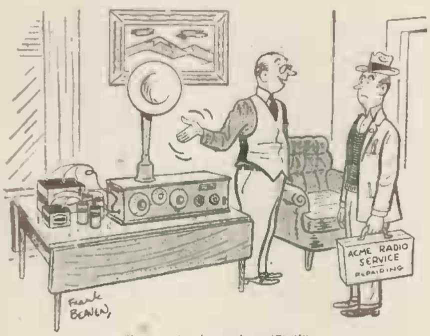
Mr. Folsom also announced the appointment of George L. Beers as Assistant Director of Engineering in charge of Advance Development. He was formerly on the engineering administrative staff.

Mr. S. B. Turner

Mr. S. B. Turner, who has been with Messrs. B.X. Plastics since 1931, has now joined the staff of the Expanded Rubber Co. (a subsidiary of B.X.) as sales manager.

Mr. O. S. Puckle

Mr. O. S. Puckle, M.I.E.E., has left Messrs. A. C. Cossor to join R.F. Equipment, Ltd., as Chief Engineer, with a seat on the Board.



"I want it changed to 'FM'!"

—*Radio News.*

B.B.C. Standard Frequency Transmissions

The B.B.C. announces that certain of its transmission frequencies are now controlled within ±1 part in 1 million of the nominal frequency. Whenever these frequencies are used by B.B.C. transmitters, they can therefore be employed as a reference standard.

The frequencies in question are 200, 6,180, 9,510 and 17,810 kc/s. Transmission hours will change with alterations in B.B.C. services, but the present schedule of transmissions is as follows (times in G.M.T.):—

(a) 200 kc/s. (European Service) 04.00-13.30, 14.30-00.45.
(b) 6,180 kc/s. (Home Service) 05.00-13.30, 14.30-22.15; (European Service) 04.00-05.00, 22.30-00.45.
(c) 9,510 kc/s. (Overseas Service) 04.00-08.00, 17.30-02.15.
(d) 17,810 kc/s. (Overseas Service) 09.00-15.15.

Rust Removal

Details of an interesting new product known as Manganesed-Phosphalene No. 7 have been received from the Leonard L. Minthorne Co., Inc., of New York. This is a non-inflammable chemical solution for removing and preventing rust and corrosion on iron, steel, zinc and alloys. It can also be used for loosening tightly rusted bolts, nuts, springs, etc., and for bonding paint, varnish, lacquer, enamel and plating to iron and steel. It is claimed that this product will dissolve rust no matter how heavily encrusted, and that by its action all possible rusting agents are thoroughly destroyed and removed.

Catalogues Received

List "B" of Messrs. Daly's, Ltd., describes their range of standard block condensers, and List "T" the tubular type ranging from 2 μF, 25 vp., to 8 μF, 350 vp. Copies can be obtained from the company at The Green, Ealing, W.5.

Three additional pages to the S. S. White book on Torsional Remote Control describe special problems in coupling operating shafts to knobs some distance away and the types of fitting recommended. The complete book is available to *bona fide* engineers on request, and inquiries should be addressed to the S. S. White Co., St. Pancras Way, London, N.W.1.

Bulletin 11a from Messrs. Handy & Harman, 82 Fulton Street, New York, describes a special application of their Easy-Flo brazing alloy to the brazing of carbide tool tips. Easy-Flo No. 3 has a flow point of 1,270° F. and is available in wire or strip.

Low-Voltage Soldering Irons

The Acru Electric Tool Mfg. Co. have now produced a low-voltage soldering iron similar in appearance to the Wireless Model but having a rating of only 25 W at 6 V. This should be specially useful to users having battery supply only available.

Particulars from the Acru Electric Tool Mfg. Co., Hyde Road, Ardwick, Manchester 12. Price 22s.

A low-voltage iron has also been produced by W. T. Henley—makers of the "Solon" brand. This can be supplied with either the oval tapered bit or a replaceable round pencil bit and is suitable for 12 V or 24 V. Details from W. T. Henley's Telegraph Works, 51 Hatton Garden, E.C.1.

CORRESPONDENCE

Photo-Mechanical Oscillators

SIR,—In a recent article in this Journal (Vol. XVII, page 326) G. R. Baldock and W. Grey Walter gave the description of a "Low Frequency Photo-Mechanical Oscillator." In this instrument a complex wave is made up by superposing the electrical outputs of a number of selenium photocells with apertures as shown in Fig. 5 of the quoted paper in front of which equidistant illuminated circular holes in a rotating disk are moved with constant angular velocity.

It can be shown by a simple elementary calculation that the total amount of light falling on one cell as a function of time is given by the formula

$$\phi(y)=\frac{1}{\pi}\left\{\arc\cos y-y\sqrt{1-y^2}\right\}$$

where

$$y=\frac{R}{r}\sin\omega t \quad -1<y<+1$$

Here ω is the angular velocity of the disk, r the radius of the hole and R the distance of the centre of the hole from the centre of the disk; the maximum light intensity is taken as unity.

In the case where $R\gg r$, y is approximately proportional to time and $\phi(y)$ represents directly the time dependence of the photo-current. Comparing this function with a harmonic function of the same period, amplitude, and phase one finds a considerable distortion (*see* Fig. 2 of the quoted paper) which amounts to as much as 7 per cent. This means that the generated function can be regarded as the sum of a harmonic function of the same frequency plus higher harmonics whose frequencies are integer multiples of the fundamental frequency and whose amplitudes are of the order of magnitude of about 10 per cent. of the amplitude of the fundamental.

If r and R are comparable in magnitude the distortion will be different for each row of holes, *i.e.*, for each component, but it will again be of the same order of magnitude.

Apart from the distortion due to the geometry of the generating mechanism another distortion of a physical nature will be added due to the fact that the sensitivity of a selenium photocell is in general not uniform throughout its whole surface. The photo-electric current will then not be precisely proportional to the illuminated area.

As long as the device is only used for the production of low-frequency complex electrical waves for, say, physiological purposes, this does not matter much. But if it is to be used with any degree of accuracy as a "harmonic synthesiser" or, by way of imitation of a given wave form as a "harmonic analyser," the components must be accurate harmonic functions. Assume, for example, that the given wave contains a second harmonic of 10 per cent. amplitude with respect to the fundamental; the distortion mentioned above will then be of the same order of magnitude as the quantity to be measured, and the same argument applies to the higher harmonics as well. Hence the amplitudes and phases of the components obtained experimentally will be widely different from the true Fourier coefficients and phases.

A combined photo-mechanical harmonic synthesiser and analyser which, of course, can also be used as a low-frequency generator, has been described by the present authors recently in an article in the *Phil. Mag.* (October, 1944). In this device, just as in that of Baldock and Grey Walter, the components are produced photo-electrically, thus avoiding the use of complicated valve generators. The components produced in our instrument, however, are purely harmonic, and no distortion is involved either in harmonic synthesis or analysis. Furthermore, only one photocell is used in the former operation and only two cells in the latter one as compared with one photocell for each harmonic component in Baldock's and Grey Walter's instrument. Moreover, by the replacement of a disk carrying the shape of the component functions by another one our instrument can be readily used for synthesis and analysis with respect to any set of orthogonal functions.

In our instrument, if used as a synthesiser, the synthesised function appears as a stationary curve on the screen of a cathode-ray oscillograph. It could, of course, as well be produced by an ink writer although it is generally recognised as an advantage for research and also for teaching to be able to produce the whole curve representing the investigated functional relation instantaneously instead of having to watch the curve being traced out by a pen; but perhaps this point of view is not shared by physiologists.

R. FURTH, Dr.Phil., F.Inst.P.
R. W. PRINGLE, B.Sc., Ph.D., A.Inst.P.

Departments of Natural Philosophy and Mathematical Physics,
Edinburgh University.

Gas-Filled Relays

SIR,—My attention has been drawn to two points in my recent note on gas-filled relays published in your February issue. They concern the question of the life of the GT1C run as shown in the figures.

First, to clarify any doubt which may have arisen it is emphasised that the values of grid voltage shown in the table for the GT1C, as well as the circuit arrangement of Fig. 1, refer only to the purpose of test. The relays should not be run continuously under these conditions of extreme negative grid voltage, which should in general not exceed −10 v. The extreme values recorded were included to draw attention to the persistence of the discharge when once established.

Second, it has been pointed out that the negative grid supply voltage of 48 shown in Fig. 2 is excessive. In these circuits the effect of positive ion grid current must be considered; the actual potential at the grid is approximately −23 v, which would result in a life of only a few hours' continuous running. I had used these circuits only for intermittent counting, and had chosen the voltages having in mind the stability of the whole counter arrangement. For continuous running, the voltage should be reduced, to give a value −10 v at the grid.

It is understood that the makers refer in the latest editions of their data charts to optimum values of grid voltage, and are glad to remark on circuits using gas-filled relays from the aspect of valve life.

D. W. GILLINGS.

MAY MEETINGS

*NOTE.—In general, visitors are admitted to the meetings of scientific bodies on the invitation of a member, or on application in writing to the Organising Secretary at the address given. In certain cases (marked *) tickets may also be obtained on application to the Editorial offices of this Journal.*

Institution of Electrical Engineers

All meetings of the London Section will be held at The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.

Ordinary Meeting

Date: May 10. Time: 5.30 p.m.
Annual General Meeting.
(Corporate Members and Associates only.)

Measurements Section

Date: May 18. Time: 5.30 p.m.
Lecture:
"Magnetic Materials."
By:
Sir Lawrence Bragg, O.B.E., M.A.

Radio Section

Date: May 2. Time: 5.30 p.m.
Lecture:
"Notes on the Stabilities of L.C. Oscillators."
By:
N. Lea, B.Sc.

Date: May 15. Time: 5.30 p.m.
Discussion on:
"The Characteristics of Luminescent Materials for Cathode-Ray Tubes."
Opened by:
C. G. A. Hill, B.Sc.

Date: May 22. Time: 5.30 p.m.
Discussion on:
"Non-Ferrous Contact Springs."
Opened by:
H. G. Taylor, D.Sc.(Eng.), and L. B. Hunt, Ph.D., M.Sc.
The Secretary:
The Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.

Cambridge Radio Group

Date: May 8. Time: 6 p.m.
Held at:
Cambridgeshire Technical College.
Lecture:
"Carrier Protection on Overhead Transmission Lines."
By:
D. H. Towns, B.Sc., A.M.I.E.E.
Group Secretary:
D. I. Lawson, c/o Pye Ltd., Radio Works, Cambridge.

The Television Society*

Date: May 29. Time: 6 p.m.
Held at:
The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Lecture:
"The Human Eye and the Photo-Cell."
By:
Dr. W. Sommer.
Lecture Secretary:
G. Parr, 43 Shoe Lane, London, E.C.4.
General Secretary:
O. S. Puckle, 8 Mill Ridge, Edgware, Middlesex.

Electronic Music Group

Date: May 12. Time: 3 p.m.
Held at:
The Northern Polytechnic, Holloway, London, N.7.
Lecture:
"A Homophonic or Single-Note Electronic Musical Instrument with a Photo-Cell as Playing Manual."
By:
W. Saraga, Ph.D.
Note.—A demonstration of an experimental model will be given.
The Secretary:
L. N. Bennett, 9 Senga Road, Hackbridge, Wallington, Surrey.

The Association for Scientific Photography*

Date: May 24. Time: 7 p.m.
Held at:
The Caxton Hall, Westminster.
Annual General Meeting followed by Lecture:
"Photography Applied to Research in the Steel Industry."
By:
S. H. Thorpe, A.R.P.S.
The Secretary: A.S.P., 34 Twyford Avenue, Fortis Green, London, N.2.

Kingston-upon-Hull Electronic Engineering Society

Date: May 11. Time: 7.30 p.m.
Held at:
The Electricity Showrooms, Ferensway, Hull.
Lecture:
"Reproduction of Sound."
By:
E. A. Buckland, B.Sc.
The Secretary:
H. W. Akester, 720 Anlaby Road, Hull.

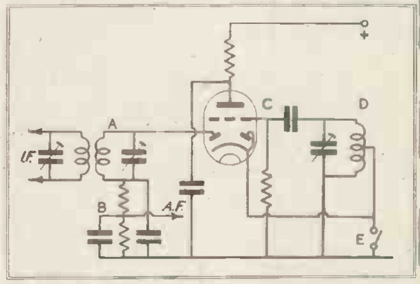
British Kinematograph Society

Date: May 16. Time: 6 p.m.
Held at:
Gaumont-British Theatre, Film House, Wardour Street, London, W.1.
Symposium:
"Auditorium Requirements in Sound-Film Presentation."—Part I.
Organising Secretary:
R. H. Cricks, Dean House, 2 Dean Street, London, W.1.

A Note on B.F.O's

By K. E. MARCUS

DEAR SIR,—I have noticed that Beat Oscillation Injection has so far mainly been achieved by coupling the beat oscillator magnetically or capacitively to the demodulator. This method is sometimes rather critical and it might therefore interest readers to know that electronic injection works quite satisfactory and is less involved.

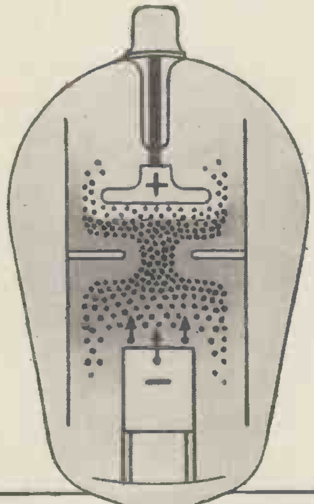


The figure shows the circuit using a double-diode-triode as demodulator and beat frequency Oscillator. The triode portion serves as an electron coupled oscillator, working at a frequency of IF+ Beat Note. D is the circuit determining this frequency, C is the conventional grid complex to maintain the oscillations. As it is known the cathode of such an osillator is on HF potential, and as the diode is worked as a conventional series demodulator for the IF across A, the frequency produced in the triode portion is injected into the diode portion. Consequently the beat note appears across the diode load complex exactly at the same point where the AF is usually taken off.

The injection is easily interrupted by closing the switch E which grounds the cathode.

B T-H

RESEARCH



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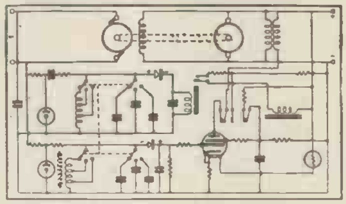
PROBLEMS WE HAVE SOLVED—NO. 1

THE ELECTRONIC GOVERNOR



CIRCUIT OF S.E.M. ELECTRONIC GOVERNOR (Pat. No. 16640/41)

S.E.M. research engineers were asked to design a governor with an accuracy of at least that of the best centrifugal governor when working under the same variation of load and supply voltage, and they developed the Electronic Governor. The circuit comprises a resonant circuit tuned to a frequency corresponding to the normal speed of the machine. The alternating current from this circuit is applied, through a rectifier, to the grid of an amplifying valve, the output from this valve being fed to an auxiliary field coil on the motor from which the main speed control is obtained. Energization of this coil varies with the speed of the motor, so that any change is immediately compensated by the Electronic Governor.

The governor gives a speed control of less than 0.5%, and the intermittent variation or hunting is less than 10% of that experienced with centrifugal governors.

This is only one of the many technical problems which we have been called upon to solve. The entire resources of our research laboratories are available to manufacturers who have a special problem involving the use of non-standard electrical devices.

SMALL ELECTRIC MOTORS LTD.

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- in fact before doing anything at all about Transformers you should consult a specialist. You know, or at least you you ought to know, that we planned and equiped a whole factory just to produce Transformers (power, audio frequency and chokes up to 10 k.v.a.) and specialised!

The dictionary will tell you that a 'micron' is the 'millionth of a metre' - but a pukka Transformer expert uses it as a 'measure of vacuum' and you know what a difference an odd spot of 'vac.impreg.' can make to the long life and performance of a Transformer.

perhaps you have never realised how much money we can save you ... we specialise in this job of producing Transformers ... savings are only to be expected.

PARMEKO LTD., LEICESTER. *Makers of Transformers*

ABSTRACTS OF ELECTRONIC LITERATURE

RADIO

Frequency Modulation
(*K. R. Sturley*)

Frequency modulation is likely to play an important part in post-war radio developments. The present position of this method of communication is reviewed; a brief discussion of the features of the three methods of modulation—amplitude, frequency and phase—is followed by a survey of the history of frequency modulation, with a consideration of its limitations, advantages and present applications.

The first of the two main sections of the paper is concerned with the production of a frequency-modulated signal: direct and indirect (integrated phase-modulation) methods are examined with particular reference to the variable-reactance valve modulator with automatic mean-frequency correction, and the indirect transposed-sideband modulator with a crystal-controlled master oscillator. Measuring and monitoring equipment is also described.

The second main section deals with frequency-modulated reception, and indicates the essential features of a frequency-modulated receiver, which, except for the amplitude limiter and frequency-to-amplitude convertor (examined in detail) is similar to its ultra-short-wave counterpart for amplitude modulation. Also included are three sub-sections dealing with tuning indicators, frequency-deviation compression, and distortion and interference. Possible future developments are discussed and there is a bibliography of the most important contributions to the literature of frequency modulation made over the past twenty years.

—*Jour. I.E.E., to be published.*

The Action of a Direct Radiator Loudspeaker, with a non-linear cone suspension system
(*H. F. Olson*)

During the past few years a number of mathematical investigators have directed their efforts toward the solution of differential equations with variable coefficients. These analyses are useful in explaining some of the phenomena which occur in electro-acoustic vibrating systems with non-linear elements. In particular, this mathematics may be used to explain the various phenomena exhibited by a direct radiator loudspeaker with a non-linear cone suspension system. One of the effects is a jump phenomenon in the response frequency characteristic. Another effect is the production of harmonics and subharmonics.

—*Jour. Acous. Soc. Am.** July, 1944.

Single- and Double-Stub Impedance Matching
(*A. H. Wing and J. Eisenstein*)

Single- and double-stub impedance matching on transmission-line feeders are described. Formulæ determining the position and the length of a single matching stub for any load impedance or any observed standing-wave voltage distribution are derived, neglecting losses in the matching sections. Conditions for minimum length of stub and minimum distance between load and stub are specified. For the double-stub arrangement, expressions for the lengths of and distance between stubs are derived for any possible combination of load and distance between stubs. A double-stub tuner with any fixed distance between stubs cannot match all loads to the feeder, and the limits are specified. The optimum location on the feeder of the double-stub tuner is discussed. A graphical method employing a circle diagram for computing the lengths and position of the stubs is explained. The graphical determination of the stub lengths of the double-stub tuner is facilitated by auxiliary circular loci which also indicate the range of admittances over which an impedance match may be effected.

—*Jour. Appl. Phys.* Aug., 1944.

A Cyclotron which allows the Accelerated Particles to Emerge in a Direction Parallel to the Dee Interface.

P. Gerald Kruger, G. K. Groetzinger, J. R. Richardson, E. M. Lyman, W. E. Ogle, M. E. Nelson, G. Schwarz, J. B. Greene, N. C. Colby, R. W. Lee, C. E. McClellan, D. Scag, Lloyd Smith, and F. K. Talmadge.

Rev. Sci. Inst., Dec. 1944, *p.* 333.

Old Uncle Tom Cobley and All.

The Theory of Transmission Lines
(*E. N. Dingley*)

The purpose of this article is not to present new material but only to describe, in step-by-step fashion, the derivation of the formulas necessary to the solution of transmission-line problems and to demonstrate, by examples, how to use these formulas. This treatment will be helpful to students and to engineers not regularly confronted with transmission-line problems.

—*Proc. I.R.E., Vol.* 33, Feb., 1945, p. 118.

MEASUREMENT

Calibrated Response Curve Tracer
(*G. L. Hamburger*)

An apparatus is described which greatly facilitates the alignment of R.F. and I.F. band-pass amplifiers. With its aid the response curve of the amplifier to be aligned is continuously traced on the screen of a cathode-ray tube in conjunction with a co-ordinate system of frequency and amplitude derived from a standard signal generator.

—*Wireless Engineer*, Vol. 22, (1945), p. 170.

The Measurement of Balanced and Unbalanced Impedances at Frequencies near 500 Mc/s. and its application to the determination of the propagation constants of cables
(*L. Essen*)

Apparatus is described suitable for the measurement of balanced and unbalanced impedances at frequencies above 400 Mc/s. It consists of an air-spaced concentric line for unbalanced impedances and a screen twin line for balanced impedances. The length of line can in each case be varied by means of a movable bridge carrying a thermo-junction unit. The component to be measured is connected to the open end of the line, which is then adjusted to current resonance by the movement of the bridge. The impedance is evaluated from the readings of resonant length and the width of the resonant curve.

The results of measurements of a series of small ceramic condensers and carbon and metallised resistors are included in the paper.

—*Jour. I.E.E.*, Part 3. June, 1944, p. 84.

* Abstracts supplied by the courtesy of Metropolitan Vickers Electrical Co. Ltd., Trafford Park, Manchester.

A New Quartz Photocell for the Ultraviolet

Developed by Cinema-Television, Ltd.

THE most frequently used photocell for the detection of ultraviolet light has hitherto been the "sodium cell," consisting of a glass bulb incorporating a sodium photocathode and provided with a quartz window to transmit radiation of wavelengths below 3,500 Å.

The sodium-photocell has two features which are inconvenient for many practical applications: Firstly, the sensitivity, expressed in quantum yield, is low so that with the small amount of energy available from light sources in the wavelength range between 2,000 and 3,00 Å only extremely small photocurrents are produced. Secondly, the sodium cell has hardly any sensitivity for visible light, which necessitates the use of a second photocell of different colour response for measurements ranging from the ultraviolet into the visible spectrum.

Both these difficulties have been overcome by the use of the antimony-caesium photocathode which has been developed in recent years for the detection of visible light. In actual manufacture it has been found very difficult to produce a glass cell with antimony-caesium cathode and quartz window because there does not seem to exist any cement which makes a vacuum-tight seal between quartz and glass and which at the same time does not react with caesium vapour during the sensitising process. The simplest solution proved to be a cell made entirely of fused quartz.

Such a quartz photocell with antimony-caesium cathode is shown in Fig. 1. The cell has a total length of 10 cm. (excluding base) and a diameter of 3.6 cm. The cathode is of semicylindrical shape and has an area of approximately 10 cm.^{2}

The antimony-caesium cathode is sensitive to wavelengths between 2,000 and 6,500 Å, with a peak sensitivity at between 4,000 and 4,600 Å. In the ultraviolet range the cell is superior to the sodium cell by a factor of the order of 100 to 500. These figures refer to vacuum cells; the antimony-caesium cell can also be made as a gasfilled cell, but in view of its high primary sensitivity and the additional difficulties introduced by gasfilling, the use of the vacuum cell is, as a rule, advisable for precision measurements.

CLASSIFIED ANNOUNCEMENTS.

The charge for miscellaneous advertisements on this page is 12 words or less 4/- and 4d. for every additional word. Box numbers count as four words, plus 1/- extra for replies. Remittance should accompany advertisement. Cheques and P.O.'s payable to Hulton Press Ltd., 43 Shoe Lane, E.C.4. Press date—15th month for following issue.

FOR SALE

IN STOCK, Rectifiers, Accumulator Chargers, Rotary Converters, P.A. Amplifiers, Mikes, Mains Transformers, Speakers of most types, Test Meters, etc., Special Transformers quoted for.—University Radio, Ltd., 22, Lisle Street, London, W.C.2. Ger. 4447.

LOUDSPEAKERS

LOUDSPEAKERS—We carry on. Sinclair Speakers, 12, Pembroke Street, N.1.

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