The TRUE ROAD TO RADIO

FERRANTI

THE TRUE

ROAD TO

RADIO

FERRANTI

CONTENTS.

PREFACE

TRUE RADIO REPRODUCTION

*T*HE *two previous editions of this book fulfilled a very useful purpose judging by the demand for copies and the numerous requests for a third edition. Here is the answer to these requests.*

In designing commercial sets for Ferranti Ltd., the author has by research, measurement and thought, learned much which was not common knowledge; some must be original, and he feels that this work is a suitable means of making this public. Naturally, because of the crowded state of the ether, a state which is becoming alarming with the rapid growth of high powered stations, much of this new knowledge concerns the radio frequency side of receivers.

Despite this, he has attempted to write a popular consecutive story, illustrated by curves and illustrations, for the majority, with the more detailed technical portion filled in as separate sections to help the growing minority of amateurs, constructors and technical men interested in this science.

ALBERT HALL, A.R.C.Sc., M.I.R.E., Wh. Ex.

June, 1931.

FOREWORD

THE TRUE ROAD TO RADIO

THIS is a personal foreword, which the writer feels compelled to insert. On the opposite leaf, is the portrait of a great man who was taken from us in January, 1930, and who founded the firm of Ferranti in London in 1882.

We owe much to his inventions, foresight, and example, and those electrical comforts we all enjoy are, in part at least, due to him. This is fully recognised by the electrical and scientific world. But I, who was his personal assistant for twenty-eight fruitful, interesting, and happy years, must bring to your notice a phase of his early work which is little known, and bears very directly on the new science in which all are so interested. The commercial interest of our firm in Radio, commenced only eight years ago, during which time Dr. Ferranti played a very important part in its development, but he used his inventive skill to its advantage whilst still a very young man.

In 1885, when he was twenty years of age, Patent No. 14917 was granted to Sebastian Ziani de Ferranti and George Leonard Addenbrooke for improvements in telegraphy and telephony, and consisted of a telephone or speech transformer capable of transmitting speech as well as of ringing bells. A sample of this invention is preserved at the South Kensington Museum for the Nation. Of course, it was the logical result of discoveries on induction by Faraday, and Dr. Ferranti's transformer for the transmission of electrical power, but it was the beginning of a mighty industry. All the telephone and telegraph services in the world depend on it, and it appears in every radio receiver as an audio frequency or an output transformer. It would be difficult to estimate the millions that have been made.

Dr. Ferranti related to me, that when the Post Office started their telephone system, they used wire returns instead of the earth returns of their competitors, the National Telephone Company. It soon became necessary to interconnect the two systems, and he was asked to devise something to permit this, and at the same time permit of bell ringing. This patent specification was the result. A photograph of the patent drawings is reproduced here.

Fig 1.

Fig 4.   Fig 3.

Fig 2.
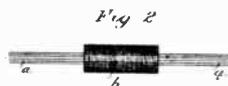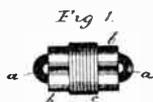
Another patent is perhaps of even more interest. Every wireless constructor, and a great many more people, have heard of Litz wire, or Litzendraht. With the introduction of the Regional Stations, its use for wave traps will probably be much increased. Its German name, and a comparatively recent use, leads to the assumption that it is a German invention of recent date. This is a mistaken idea. Specification 2315 of 1888, granted to Sebastian Ziani de Ferranti for Improvements in Electrical Mains, covers the construction of what is now known as Litz wire, or cable. The claims are :—

1. An electric main composed of a number of thinly insulated wires formed into a cable or rope substantially as described, for conveying electric currents of high tension.*

   *This should have read " high frequency." The mistake being due to the Patent Agent.

2. An electric main composed of a number of thinly insulated wires so put together that each wire is the same average distance from the centre of the main.

The second claim is very important for high frequency work. Each strand must pass from the outer surface to the inner surface at frequent intervals along the length so that each may pass through every portion of the magnetic field due to the current.
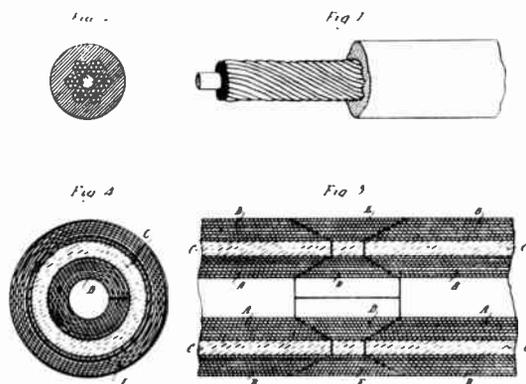
Dr. Ferranti at one time mentioned that William Thomson, later Lord Kelvin, told him of his discovery of the repulsion of the current by its own magnetic field, leading to the fact that the surface of the conductor is more important in carrying alternating current than the centre, so Dr. Ferranti decided that he had to make a cable which was to all intents and purposes all surface and every portion of which passed in and out of every part of the magnetic field.

In these days of 50 cycle current, one fails to realise the necessity of a cable so constructed. The Doctor stated, however, that in the early days, the frequency of alternating current was very high. 200 cycles was quite common, and some of the machines gave current at over 300 cycles per second, the machines being run as fast as possible, consistent with mechanical strength. Therefore, in those early days, the resistance of the cable would become quite important. Actually, a cable was used in the City of London Scheme in 1892 made of hemp core with insulated wires stranded round it. The frequency of current transmitted was 100 cycles.

There is another patent which has become of peculiar interest recently, because of the use of alternating current mains to supply power to radio receivers. Patent No. 2313 of February 15th, 1888, granted to Sebastian Ziani de Ferranti for improved means of obtaining uniform continuous current from Dynamo Electrical Machines. This, as far as is known, is the first smoothing or filter patent, and the claim is for :—

> " Obtaining a uniform or comparatively uniform continuous current from dynamo electric machines producing alternate currents, first changing the alternating current by a commutator into a current of one direction, and subsequently equalising the strength of this current by means of a self-induction coil or coils and condenser substantially as described."

Patent drawings are reproduced below.

The commutators spoken of, formed part of a rotary rectifier which was used in those early days.
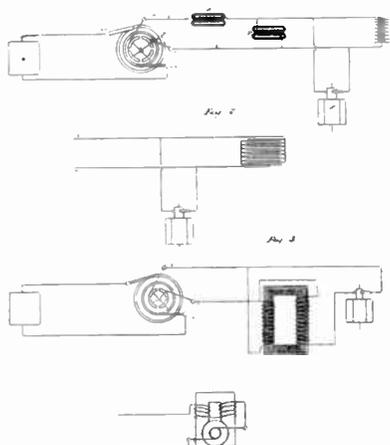
THE TRUE ROAD TO RADIO

FAITHFUL radio reproduction, in the strictest sense of the word, is not yet a possibility, since the various processes of conversion, transmission, and reconversion from the broadcast studio or other sources, to the listener, are not yet good enough to ensure that what is heard is an exact replica of the original.

It is difficult to say what the original is, since if one listens to an artist or orchestra, just what is heard depends on the type of room in which the performance occurs and the position of the listener therein.
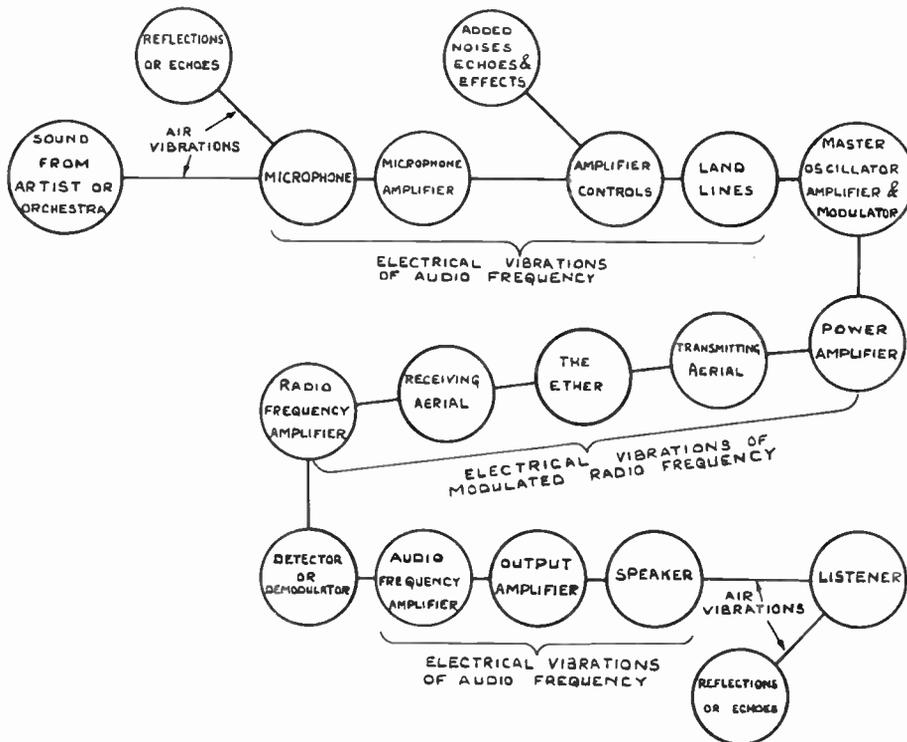
Fig. 1.

To understand the nature of the problem, it is necessary to consider the path of the sound or its electrical equivalent from the points at which it occurs to the point where the listener hears it.

Figure 1 shows in a diagramatic form the number of stages and the nature of the processes through which the matter broadcast has to pass in almost all cases before it reaches the listener, and a consideration of this figure shows that the reproduction obtainable, when the best apparatus is used, is very remarkable, when every link in the chain may produce some degree of distortion.

We may say that the ear is not equally sensitive over the frequency scale, nor linear for different amplitudes ; but this does not affect our problem of perfect reproduction. These variations are allowed for, in the music as played. In whatever way our imperfect ears appreciate the original, we wish the reproduction to be appreciated in the same way, therefore no allowance for the defect of the ears need be made.

For the purpose of this book, we have to assume what is not always correct, namely, that the output from the broadcasting station is free from distortion, since we are mainly concerned with the reception and conversion into sound of broadcast transmissions.

Carrier Wave. All broadcasting stations send out or transmit a rapid electrical alternation or vibration known as the carrier wave, and the number of alternations per second is known as the frequency of this carrier wave, and may be expressed as cycles per second, or kilocycles per second.

For example :—The frequency of the 5XX carrier wave is 193,000 cycles per second, or 193 kilocycles per second.

Because of the rise and fall of electrical strength (voltage) during each alternation in the transmitting aerial, an electro-magnetic field is created in space, which at any instant of time,

apart from the general attenuation, has a rise and fall in value with distance—is, in fact, a wave.   The distance between any two consecutive maxima or minima values is called the wave length of the carrier wave.   This is true at any instant of time, but the maxima are travelling through space with the velocity of light, i.e., the carrier wave has this velocity, namely, 300,000,000 metres (186,000 miles) per second.

Determination of Wavelengths.

Hence the relation exists :—

$$\text{Wave length in metres} \ldots = \frac{300,000,000}{\text{Frequency in cycles per second.}}$$

$$\text{Wave length of 5XX (frequency 193,000 cycles)} = \frac{300,000,000}{193,000} = 1,554 \text{ metres.}$$

Most readers will be familiar with both these methods of defining the carrier wave of a broadcasting station.
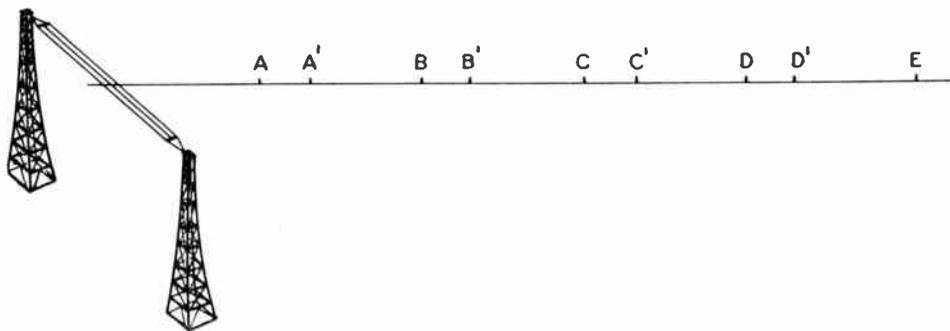
Fig. 2.

If we had suitable measuring means, we should find that at some given instant of time the electromagnetic field was at a maximum at consecutive points, A, B, C, D, and E, etc. Fig. 2, falling as one proceeded from A to a minimum at the centre of A B, and rising again to a maximum at B.   A B is the wavelength of the signal; in our example, A B is 300 metres, therefore the wavelength is 300 metres.   If we explore a $\frac{1}{1,000,000}$ part of a second later, we should find

SHORT WAVE scale (KILOCYCLES / METRES):
30000 — 10
20000 — 15
15000 — 20
— 25
10000 — 30
9000 — 35
8000 — 40
7000 — 45
6000 — 50
— 55
5000 — 60
— 65
— 70
4000 — 75
— 80
— 85
— 90
— 95
3000 — 100

SHORT WAVE
10 – 100 METRES

KILOCYCLES   METRES

MEDIUM WAVE scale (KILOCYCLES / METRES):
3000 — 100
2000 — 150
1500 — 200
— 250
1000 — 300
900 — 350
800 —
— 400
700 —
— 450
600 — 500
— 550
500 — 600

MEDIUM WAVE
100 – 600 METRES

KILOCYCLES   METRES

LONG WAVE scale (KILOCYCLES / METRES):
500 — 600
— 650
— 700
400 — 750
— 800
— 850
— 900
— 950
300 — 1000
200 — 1500
150 — 2000
100 — 3000
90 —
80 —
— 4000

LONG WAVE
600-4000 METRES

KILOCYCLES   METRES

FREQUENCY AND WAVELENGTH

the maxima now at points A^1, B^1, C^1, D^1—where A A^1 equals 30 metres : that is, the maxima are travelling through space with a velocity equal to $\dfrac{30}{\dfrac{1}{1,000,000}} = 300,000,000$ per second, which is the velocity of light.

The ordinary broadcasting stations of Europe have frequencies ranging between :—

Leeds ... 1,500,000 cycles ... Wave length, 200 metres.

Kaunas (Lithuania)—

155,000 cycles ... Wave length, 1,935 metres.

Short Wave Band.

There is, however, a recent development of transmissions on what is known as the *short wave band*—namely, between wave lengths of 15 metres (20,000 kilocycles) and 100 metres (3,000 kilocycles). These are not at present of general interest, being arranged for and best suited to reception at great distances, principally in the Colonies.

Figure 3 shows three scales, from which the frequency in kilocycles corresponding to any particular wavelength in metres, or vice versa, can be determined at a glance.

Modulation.

Frequencies of the carrier waves indicated above are far above the audible limit, but are necessary as a medium for carrying the audible sound frequencies. The latter, after being converted into their electrical equivalent, are superimposed on the former by a process called *Modulation*. It is not proposed to describe the technical details of the process of Modulation, but an examination of Figure 4 will enable the process to be visualised.
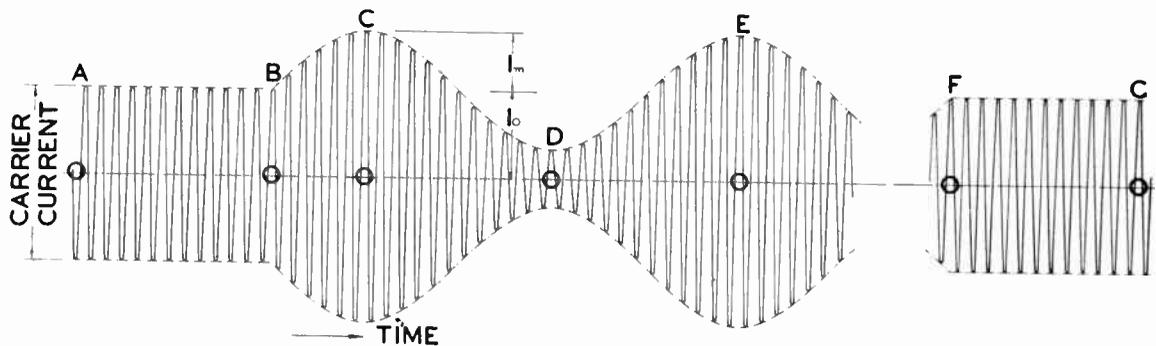
Fig. 4.

Here, horizontal measurements represent time, and vertical measurements, such as O A, the instantaneous value of the carrier current. From time A to B, when the microphone is not in use, O A is constant. When the microphone receives a sound pressure wave or a musical note, the electrical current equivalent generated by the microphone and amplified, is used to change the amplitude of the carrier wave current in accordance with the pressure changes of the sound wave. As the pressure increases, the amplitude of the carrier current increases to a value O C; as it decreases, the carrier current decreases to a minimum value O D; then it increases again and so on. The line joining the tips of the successive carrier current alternations B, C, D, E, F, represents the musical note. When the note is finished, the carrier current again becomes constant as from time F to G.

Overmodulation. The ratio $\dfrac{IM}{IO}$ multiplied by 100 represents the percentage modulation. The reader will have heard of and experienced the effects of *overmodulation*. This occurs when the value IM becomes so great relative to IO that the line B, C, D, E, F does not accurately repeat the original sound wave. The earlier processes of modulation were such that $\dfrac{IM}{IO}$ could not be very great without distortion. Recent improvements have enabled it to be much greater.

For a given station IO is fixed, and the signal received in an aerial is fixed by the aerial and set. IM is the variable which appears as sound in the speaker; hence the greater $\dfrac{IM}{IO}$ can be, without appreciable distortion, the louder will be the music for a given range, and the greater the range of the station, or alternately, the more efficient the station.
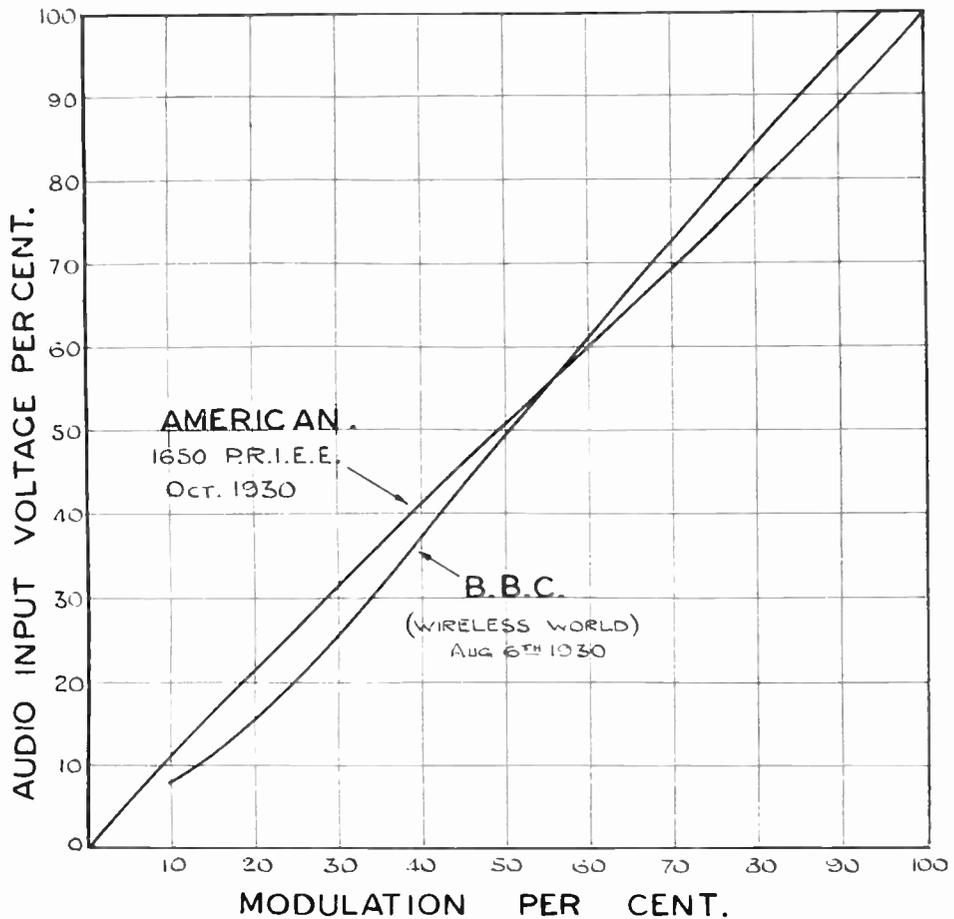
Fig. 5.

The ideal would be when at the peak passages of music or speech $\dfrac{IM}{IO} = 1$, a modulation of 100%.

Figure 5 shows the modulation curve of the best B.B.C. station and one of an excellent American station. It is clear that the modulation can approach 100% without serious distortion. The B.B.C. states that the modulation it uses, except accidently, is not allowed to exceed 80%.

Side Bands.   To appreciate how a special form of distortion can, and very often does, occur in the high frequency side of a receiving set, we must further consider Figure 4. It can be shown mathematically and experimentally that when the modulated carrier wave form B, C, D, E, F is that of a pure note, the station is in effect transmitting three frequencies. One is the frequency of the carrier wave, another the frequency of the carrier plus that of the pure note, and the third the frequency of the carrier minus that of the pure note. The two latter are known as side bands.

Example :—The London Regional carrier wave is 356·3 metres, a frequency of 842,000 cycles per second. If the microphone is receiving a pure note of 1,000 cycles, then the three waves transmitted are :—

$$842,000 - 1,000 = 841,000$$
$$842,000$$
$$842,000 + 1,000 = 843,000$$

When music or speech is being transmitted, the line B, C, D, E, F, representing the variations of the carrier current, is of a much more irregular shape than the pure sine wave shown in Figure 4. It can always be resolved, however, into a number of such pure sine waves which may range from 50 cycles to 8,000 cycles per second. For each, there will be a pair of associated side bands, so that a transmitting aerial may, and often does, send out a band of frequencies ranging from 8,000 cycles less, to 8,000 cycles more, than the carrier frequency. This fact introduces many difficulties, and the results in the receiving set will be dealt with later.

The energy of the music, or speech modulated carrier wave, spreads through the ether in all directions, and a very small portion of it is received by your aerial.

As a matter of interest, the signal voltages measured in a very ordinary aerial four and a half miles south of Manchester during September and November, 1930, are given below.

The units are microvolts, or millionth parts of a volt.

TABLE OF SIGNAL VOLTAGES IN AN AERIAL.

STATION.	DISTANCE. Miles.	POWER. K.W.	WAVE LENGTH. Metres.	DATE AND TIME. 1930.	SIGNAL AERIAL TO EARTH. Microvolts.
Hilversum	355	8·5	1,875	Sept. 3rd, light ...	1,050
Radio Paris ...	378	17	1,725	Sept. 3rd, light ...	1,000
Daventry 5XX ...	100	35	1,554	Sept. 3rd, light ...	15,000
Eiffel Tower ...	378	15	1,445	Sept. 3rd, light ...	1,000
Motala	750	40	1,343	Sept. 3rd, dark ...	300
Kalundborg ...	612	10	1,153	Sept. 3rd, dark ...	220
Huizen	298	8·5	1,071	Sept. 3rd, dark ...	420
Budapest	1,023	23	550	Sept. 3rd, dark ...	730
				Nov. 14th, 8.30 p.m.	1,490
				Nov. 16th, 10.40 p.m.	720
Munich	708	1·7	533	Nov. 16th, 10.42 p.m.	580
Vienna	880	20	517	Sept. 3rd, dark ...	650
				Nov. 14th, 8.40 p.m.	1,390
				Nov. 16th, 10.4 p.m.	720
Milan	766	8·5	501	Sept. 3rd, dark ...	700
				Nov. 18th, 7.25 p.m.	680
				Nov. 24th, 9.23 p.m.	1,320
Prague	750	5·5	487	Nov. 14th, 8.30 p.m.	910
Daventry 5GB ...	100	25	479	Sept. 3rd, light ...	5,450
				Nov. 14th, 8.50 p.m.	5,250
Langenberg ...	420	17	473	Sept. 3rd, dark ...	1,190
				Nov. 23rd, 2.37 p.m.	270
Lyons (La Doua)	378	5	466	Sept. 3rd, dark ...	1,560
Rome	1,051	50	441	Sept. 3rd, dark ...	1,300
				Sept. 3rd, dark ...	2,080
				Nov. 14th, 9.0 p.m. ...	5,800
Stockholm	873	75	436	Sept. 3rd, dark ...	2,480
				Mar. 18th, 9.0 p.m. ...	1,140
				Nov. 23rd, 2·35 p.m.	310

Station.	Distance. Miles.	Power. K.W.	Wave Length. Metres.	Date and Time. 1930.	Signal Aerial to Earth. Microvolts.
Witzleben	665	1·7	418	Nov 25th, 9.25 p.m.	440
Dublin	167	1	413	Sept. 3rd, light ...	1,090
				Nov. 16th, 10.7 p.m.	1,250
Katowice	945	10	408	Sept. 3rd, dark ...	495
				Nov. 16th, 10.10 p.m.	1,020
Berne	640	1·1	403	Nov. 16th, 7.0 p.m.	650
Glasgow	200	1·2	398·9	Nov. 22nd, 5.55 p.m.	1,090
Frankfurt	430	1·7	390	Nov. 16th, 6.55 p.m.	1,050
				Nov. 23rd, 2.32 p.m.	100
Toulouse	704	8	385	Nov. 16th, 6.53 p.m.	700
Manchester ...	4½	1·2	376·5	Sept. 3rd, light ...	117,000
				Nov. 14th, dark ...	106,000
Muhlacker	600	75	360	Nov. 23rd, light, 2.30 p.m.	360
				light, 2.40 p.m.	850
London Regional	180	30	356·3	Sept. 3rd, light ...	730
				Nov. 16th, 10 p.m. ...	1,420
Barcelona	860	8	349	Sept. 3rd, dark ...	625
				Nov. 16th, 10.15 p.m.	1,020
Strasbourg	554	15	346	Nov. 15th, 9.0 p.m.	4,350
				Nov. 16th, 6.50 p.m.	1,300
				Nov. 23rd, 2.15, day	160
Brussels (No. 2)	335	3	338·2	Sept. 3rd, dark ...	625
				Nov. 16th, 10.20 p.m.	1,320
Breslau	745	1·7	325	Sept. 3rd, dark ...	614
Goteburg	625	15	322	Sept. 3rd, dark ...	1,330
				Nov. 16th, 6.40 p.m.	1,450
				Nov. 23rd, 2.20 p.m.	240
Marseilles (PTT)	765	1·5	316	Nov. 22nd, 8.40 p.m.	1,040
Bordeaux (Lafayette ...	602	35	304	Nov. 15th, 9.0 p.m.	1,300
				Nov. 15th, 10.35 p.m.	950
Turin	760	8·5	296	Nov. 17th, 9.40 p.m.	920
				Nov. 17th, 7.30 p.m.	620
Nürnberg	650	2·3	239	Nov. 16th, 6.15 p.m.	1,320
Cologne	745	1·7	227	Nov. 15th, 9.10 p.m.	1,900
				Nov. 22nd, 8.20 p.m.	2,070
Cork	240	1·5	224·4	Nov. 22nd, 8.45 p.m.	2,050

All these signals are relatively strong, and are easily measured; in suitable locations, some can even be much stronger. Very much weaker signals can, however, be utilised effectively. Thus, in the Ferranti A.C. Receiver, Model 32, by suitable use of the reaction control, a signal of 25 microvolts, wave length – 350 metres – will give full standard loud speaker output (50 milliwatts at 30% modulation).

To be better able to judge the suitability of the signals for general use, the following facts must be noted :—

(a) That transmissions from stations on the long wave band —namely, 1,000 to 2,000 metres—do not attenuate with distance as rapidly as those on the medium band— 200 metres to 550 metres. It is known that 5XX (1,554 metres) can be received in distant parts of Great Britain when it is impossible to receive any other English station.

(b) That many long wave stations can usually be received both during daylight and darkness. The more distant vary somewhat in strength from time to time, but not excessively so.

(c) That the powerful English medium wave stations, such as the two Brookmans Park Transmitters and 5GB (no doubt the same will apply to other Regional stations when completed) can be received at quite good strength up to 180 miles (our personal experience is limited to this distance) during daylight. In some cases, however, quite a weak station can be received at good strength consistently during daylight. Near Manchester, Dublin, 167 miles away, a one kilowatt station, can be received during daylight or darkness stronger than the 30 k.W. London station (356·3 metres) without appreciable fading. This, however, is exceptional.

(*d*) That the medium wave stations more distant than 180 miles—such as the Continental stations—can rarely be received during daylight (exceptions are cases of the latest stations of very high power). During darkness, many can be received at great strength, but not consistently, as a station may be strong one night and weak the next. Further, all stations more than 120 miles distant are liable to fade after dark, often so badly as to render listening impossible.

To make use of the extremely small amount of energy received in the aerial, four operations are usually necessary in the receiver, and they are as follows :—

THE OPERATIONS OF A RECEIVER.

High Frequency Amplification. The first operation is the magnification or amplification of the very small signal received in the aerial. This is called high frequency or radio frequency amplification, and can be performed in a variety of ways, indicated later.

Detection. The second operation, which in the case of a receiver operated near to a powerful station, may be the first, is known as detection or rectification. A better term is *demodulation,* as it is the opposite of the operation performed in the studio. This consists of the separation of the audio frequency impulses corresponding to the sounds in the studio, from the radio frequency carrier wave which at this stage is no longer required. The audio frequency currents obtained by this process of detection may be utilised to operate telephones, but are not sufficiently strong to operate a loud speaker.

Audio Frequency Amplification. The third operation is the amplification of the audio frequency currents obtained by detection.

*Power
Amplification.*   The fourth and last operation is the conversion of the amplified signal, which is in the form of electrical pressure or voltage, into power, and thence into sound, through the medium of a speaker.

The Valve.   The component which has made these operations relatively easy is known as the Thermionic Valve. Suitably used, it can amplify radio frequency voltages, can rectify, detect or demodulate, and can amplify audio frequency voltages. Further, it can supply the energy which is converted into sound by the speaker.

For the valves of a receiver to function, three supplies of electrical energy are necessary :—

(1) Low tension current to heat the filament or cathode, enabling it to emit electrons.

(2) The high tension voltage to attract the electrons from the filament to the plate or anode of the valve.

(3) The grid bias voltage, by means of which the grids of the various valves are maintained at their correct mean potential.

These three supplies are known in many overseas territories as " A," " B," and " C " supplies respectively.

HIGH OR RADIO FREQUENCY AMPLIFICATION.

This is necessary where the signal received from the aerial is too small for satisfactory detection, i.e. when a distant or not very powerful transmitting station is tuned in.

Before dealing with the details of radio frequency amplification, it is necessary to state that this part of a receiver often causes more distortion than any other,

including the speaker. This arises from the crowded state of the ether, and the necessity of great selectivity to separate one station from another, often only 9,000 cycles apart.

This fact is now receiving recognition, and will be dealt with later. The following paragraphs, therefore, treat radio frequency amplification from the point of view of efficiency and stability (freedom from self oscillation) only.

Neutralising.   Until recently, radio frequency amplification has not been an easy operation, owing to the necessity for what is known as " neutralising." Before the advent of the " screened grid " valve, the valves used for radio frequency amplification had three electrodes : the filament or cathode, the grid, and the plate or anode. The anode and grid formed two plates of a small condenser. In utilising the valve for radio frequency amplification, two coils are used, each of which may be tuned, one being connected between the grid and filament and the other in some way between the anode and filament. When a station is tuned in, a portion of the magnified energy in the anode circuit is fed back through the above mentioned small " anode to grid " condenser, to the grid.

If the coil losses are sufficiently low, these are overcome, and oscillation is set up. In the early attempts to obtain radio frequency amplification, stability was attained by increasing the losses in the coils, resulting in stability but small amplification. The first really successful way of overcoming this oscillation and attaining a large degree of amplification was the use of what is known as " neutralisation."

Figures 6 and 7 show the theoretical circuits for what are known as neutralised tuned anode and neutralised H.F. transformer couplings. Anode bend rectification is shown, but grid leak rectification can be employed.
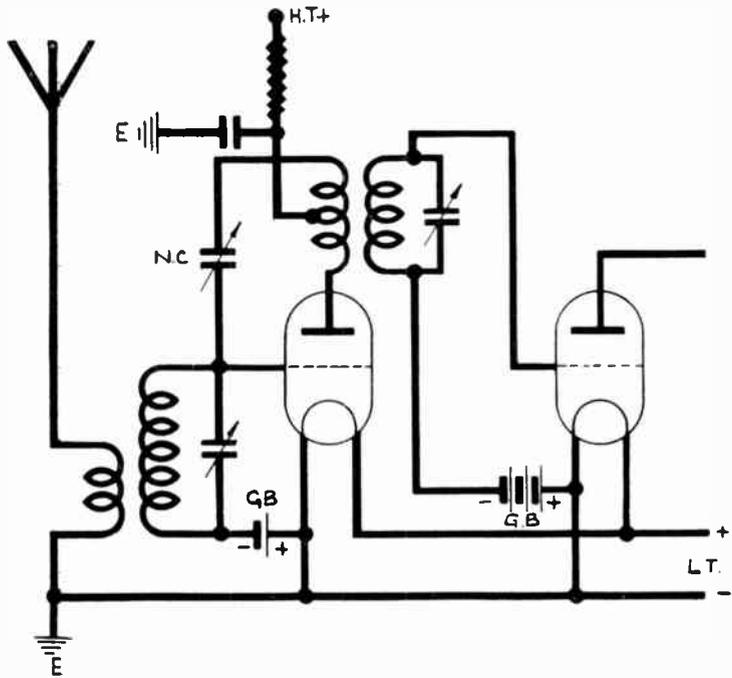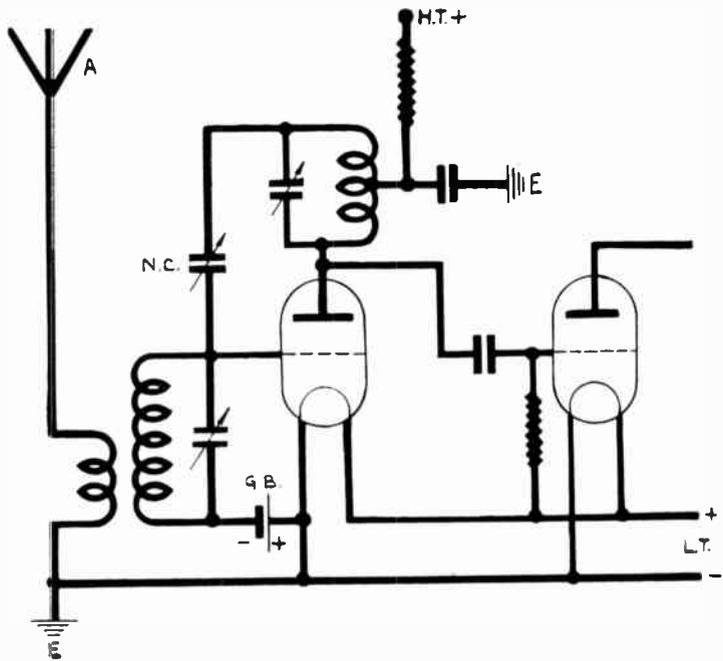
Figs. 6 and 7.

Neutralisation consists of feeding back a small portion of the energy in the anode circuit of equal and opposite value, or, as it is usually stated, in opposite phase to that unavoidably transmitted from the anode to grid, the two thus neutralising each other. This neutralising feed-back takes place via a small neutralising condenser NC, shown in Figures 6 and 7.
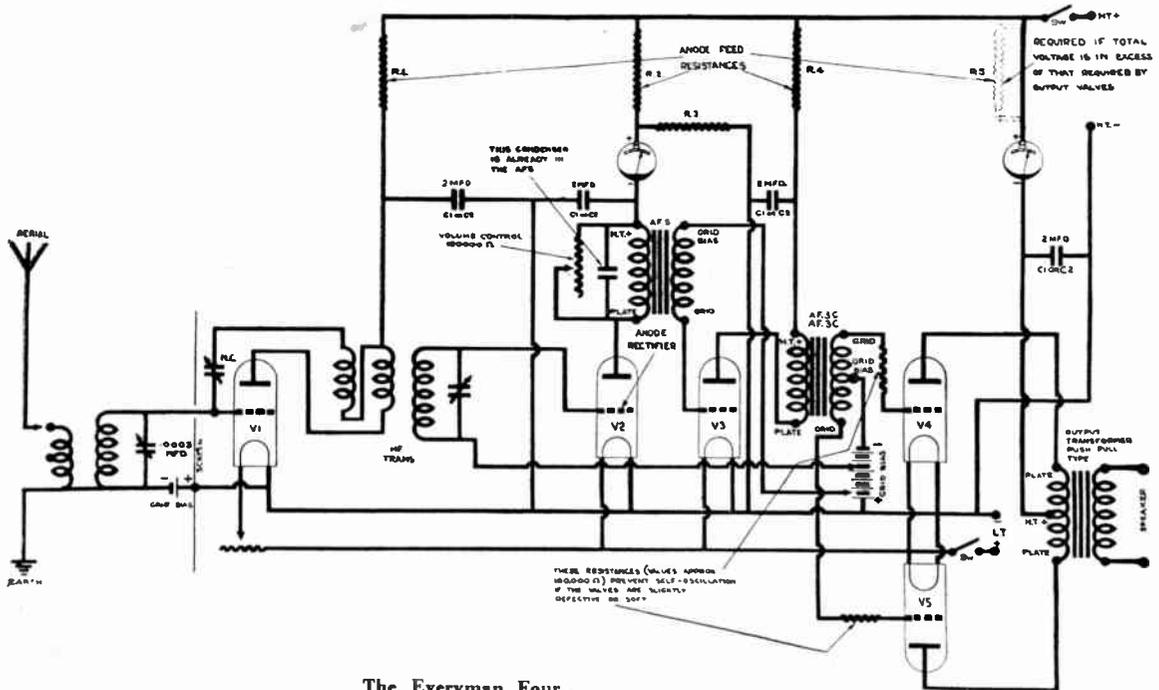
The Everyman Four.

Originally published by "The Wireless World."
Published by courtesy of "The Wireless World."

Modified by Ferranti Ltd.

Fig. 8.

Figure 8 shows the circuit of a modified form of the "Everyman 4" receiver.

"Everyman 4." The best form of receiver using neutralised H.F. transformer coupling is the well known " Everyman 4," which appeared several years ago in the *Wireless World,* and provided that the neutralisation is carried out accurately, and highly efficient coils are used, as in this receiver, very satisfactory results can be obtained. The disadvantage of neutralisation is, that when efficient coils are used, without which the amplification is not high, re-neutralising has to be carried out when using new valves, and also when changing from the medium wave to the long wave broadcast band.

Moreover, present day valves have such high amplification that neutralising now presents difficulties, especially if an attempt is made to use more than one high frequency stage.

Screened Grid Valve. With the introduction of the Screened Grid Valve, the problem of satisfactory radio frequency amplification has been greatly simplified. This valve has four electrodes, the fourth being a screen of wires very closely spaced and interposed between the grid and the anode, and often continued round the back of the anode. This screen is connected to some point on the H.T. supply to maintain it at the desired potential, but where radio frequency is concerned, it is earthed. This reduces the feed-back or the capacity from anode to grid in the case of a well made valve, to one thousandth part of that in a normal three electrode valve ; could the screen wires touch, which is impossible, the capacity would be zero.

The introduction of the screen changes the properties of the valve. The amplification factor varies in different makes from 200 to 1,500, as compared with 20 to 35 in the case of the three electrode valve.
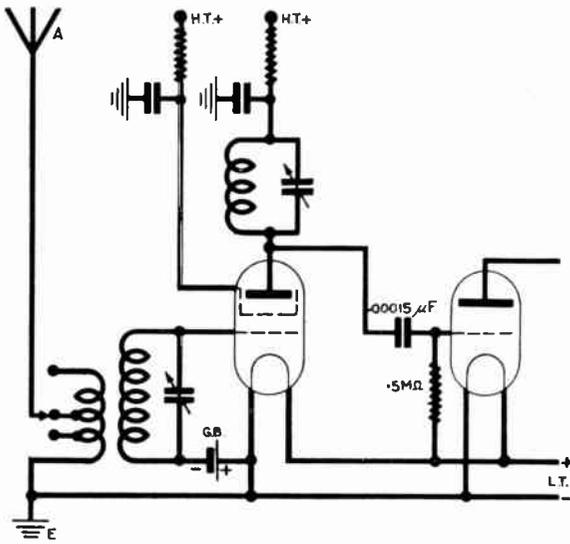
Fig. 9.
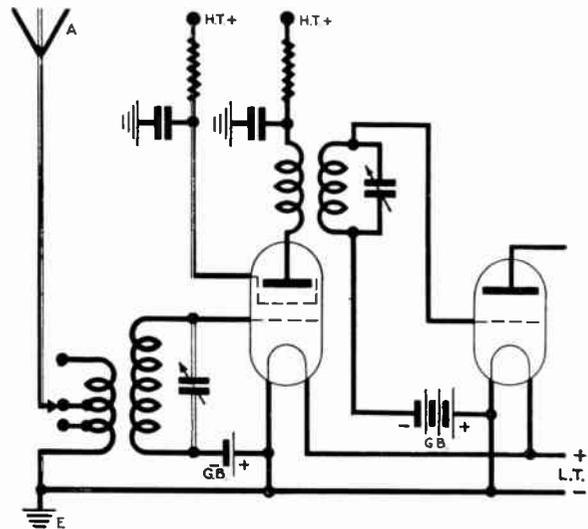
Fig. 10.

Figures 9 and 10 show the theoretical circuits for tuned anode and H.F. transformer coupling, using the screened grid valve.

At first sight, it would seem, from these figures, that extraordinary amplifications would be attainable, but unfortunately, the resistance of the screened grid valves vary from 300,000 ohms to 1,000,000 ohms, compared with 20,000 to 30,000 ohms for the ordinary valve. Since the amount of impedance we can put in series with the anode by our tuned circuits is very limited, we can only step off a portion of the very high amplification. In practice, with good coils and valve holders, etc., we can get an amplification of, from 40 to 50 for a single stage, somewhat greater than the best attained by neutralising and the use of highly efficient Litzendraht coils in the " Everyman 4 " popularised by the *Wireless World*.

Such an amplification per stage as 50 for two stages could only be obtained with special precautions. The capacity anode to grid, though it is only one thousandth part that of an ordinary valve, still plays a part and introduces considerable reaction effect if one attempts to get too great an amplification per stage. Those interested in the technical discussion of this are referred to :—

(1) *Experimental Wireless* and the *Wireless Engineer,* January, 1928, page 3. " The Stability of the tuned grid tuned plate H.F. amplifier," by R. T. Beatty, which deals with a single stage.

(2) *Proceedings of Institute of Radio Engineers.* Paper by Nelson.

The stability under good conditions, however, and the efficiency of the screened grid valve, are such that really satisfactory commercial receivers involving radio frequency amplification have only been possible since its introduction.

It should be observed that, when using either neutralised or screened grid valves, it is necessary to employ screening between the anode circuit or coils and those in the grid, as otherwise there will be a direct feed back between these portions of the circuit, notwithstanding the removal of the feed back through the valve. It is evident that it would be useless to have reduced the latter without also reducing or preventing the former.

Screens.   This feed back from the anode circuit to the grid circuit can be prevented by the use of earthed metal screens. Considerable screening can be obtained by the use of a simple sheet of copper or aluminium from ·032 to

·04 inches thick placed across the set between the two circuits, the screened grid valve being located partly through the screen, especially if the coils on the two sides be placed with their axes at right angles. Attempts have been made, as in the original Ferranti Screened Grid 3 and 4 Constructional Receivers, to reduce the feed back by using so-called " astatic coils "—i.e. a coil in which one half is wound in the opposite direction from the other. There is little doubt that the gain in stability when using such coils is not due to their so-called " astatic " nature, but to the increased resistance, since much more wire has to be used to get the same inductance as in a plain coil. The same result could be obtained by a plain coil of much smaller dimensions wound with finer wire to have the same inductance and resistance.

Astatic Coils.

Toroid Coil. One form of coil which has no external field and so cannot in itself cause feed back, is the toroid coil, a coil wound like an anchor ring. It has been used in at least one make of commercial set. The objections to it are :—

1. It is large.

2. It is somewhat difficult to wind.

3. It has a rather large resistance for a given inductance.

The only wholly satisfactory method of screening is to enclose completely one or all of the circuits within a metal box. In the case of a single screened grid valve, it has been found sufficient and best, to enclose completely the anode circuit within the box ; this must include the anode coil or H.F. transformer, the tuning condenser (unless otherwise screened), the *detector valve* and the H.F. by-pass condenser connected to the plate, so that there is little chance of appreciable H.F. voltage leaving the screen box.
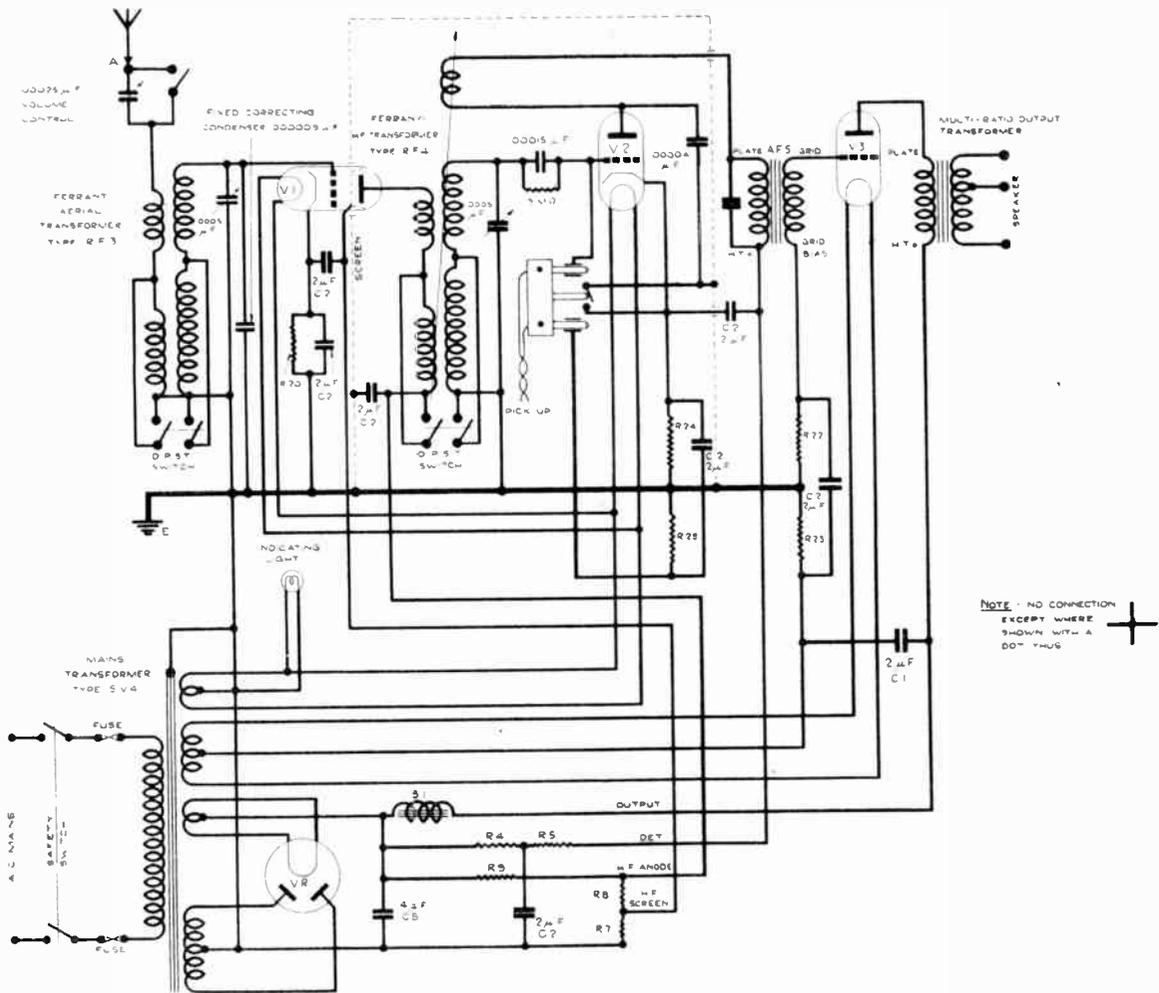
Fig. 11.

Figure 11 shows the circuit diagram of the Ferranti A.C. Mains Set, Models 31 and 32. The dotted line shows the extent of the screening box. The only additional precautions are the by-pass condensers from the H.T. connections to the screen and the anode of the screened grid valve.

HF VALVE S4VA
REACTION COIL
DETECTOR MH4.
H.F TRANSFORMER
PICK-UP SOCKETS
OUTPUT TRANSFORMER
MAINS PRIMARY TAPPING
RECTIFIER U10
SAFETY SWITCH SOCKETS
LOUD SPEAKER SOCKETS
PARALLEL GRID CONDENSER

AERIAL - GRID TRANSFORMER
AERIAL - SERIES CONDENSER
L.F TRANSFORMER
OUTPUT VALVE P625

INDICATING LAMP CARRIER
H F TRANSFORMER
TAPE CONDENSER DRIVE
TUNING CORRECTOR
TUNING CONTROL
WAVE - CHANGE CONTROL
REACTION CONTROL

INPUT VOLUME CONTROL
BY-PASS CONDENSERS
FEED RESISTANCES
"ON" AND "OFF" SWITCH
MAINS TRANSFORMER

HIND

Fig. 11a.

Constructional Charts.

Figure 11a is representative of a popular type of set, and will be described more fully later, and figures from a large number of measurements made will be used to illustrate the principles involved.   For those desirous of constructing such a set, a constructional chart will be supplied on application, also a constructional chart of the same type of set supplied with energy from batteries.

In a set as shown in Figure 11 fitted with switches to change the wave band in which it may be tuned, the portion of the inductance wanted for the longer wave band should not be placed too close to the other, i.e. they should not couple too closely, and should have its ends shorted by the switch arms as shown, otherwise the unused inductances, having self capacity and therefore a natural tuning frequency, will act as absorption wave traps (see pages 83 and 84) of this frequency and greatly reduce the amplification of the set at frequencies bordering on this.

A well known commercial set has greatly reduced amplification from 390 metres to 410 metres due to this effect.

There is an additional precaution necessary in the use of such a set.   The aerial lead and the loud speaker leads should not cross, or be brought near to each other ; it is better that the aerial lead should be brought to the set on the left, and the speaker leads be taken direct away to the right.   The reason is, that despite good filtering after the detector valve, there may be, when receiving a high power station, a slight trace of radio frequency left in the speaker leads which may feed back from these to the aerial and cause an oscillation.

Where two screened grid valves or two stages of radio frequency amplification are used, greater precautions will have to be taken.   Primarily, for the reasons given on page 27, the full amplification possible per stage cannot be

utilised without a grave risk of oscillation, especially on the shorter wave lengths from 200 to 300 metres. The circuits connecting the valves, therefore, must not be the most efficient possible. All the circuits must be completely screened by earthed metal boxes. The joints of the screening boxes must be good ; covers should have $\frac{3}{8}''$ of clean and good fitting overlap, to earth them effectively. The grid bias, screen, and anode voltage leads should be shunted to the metal by a ·1 mfd condenser at the point where they leave, while still within their respective boxes, and outside the box a resistance of 500 to 1,000 ohms should be inserted in each, to make the shunting condenser more effective. The negative of each valve, in the case of D.C. valves, or the cathode in the case of indirectly heated valves, should be directly earthed where the circuits permit of this ; if not, each point should be shunted by a ·1 mfd condenser to earth.
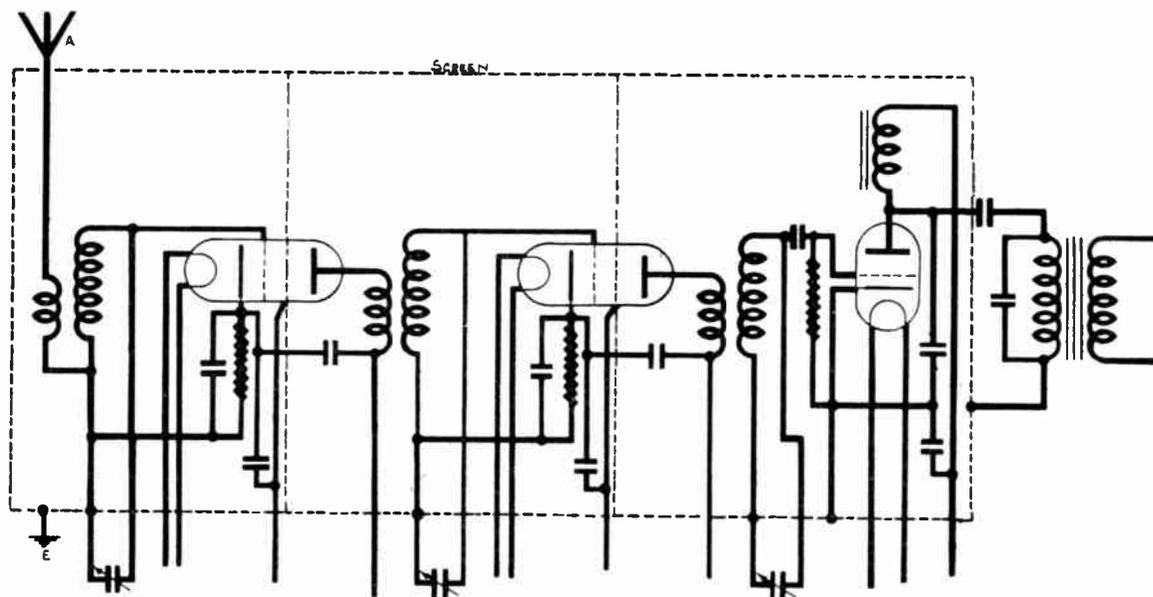
Fig. 12.

Figure 12 shows a circuit diagram with the above precautions.

LONG WAVES

Ferranti A.C. Mains Receiver.

Further, in the latter case, each heater must be earthed at some point, often the centre point of the supplying mains transformer. Finally, where a common ganging shaft passes from box to box to drive the condensers, this shaft must be earthed if possible, direct in each screen box.

All these precautions show that, unless a constructor follows exactly a design which has been well thought out and well tried, he is likely to have trouble.

So far, we have only treated radio frequency amplification from the view point of efficiency and stability. We will now show how it usually introduces distortion.

The introduction of distortion by Radio Frequency Amplification. On page 16 it was shown that, when a station is transmitting music or speech, it sends out in addition to the carrier frequency, other frequencies known as side bands, spaced equally on each side of the carrier frequency by an amount equal to the pure musical note concerned. These side bands are caused by the music, and in effect carry the music. As explained, for good transmission they may reach 8,000 cycles each side of the carrier frequency. Hence, a receiver may receive a band of frequencies approximately 16,000 cycles wide when tuned to the carrier frequency, and unless it can amplify all these frequencies equally, the music reproduced cannot be like the original. Receiving a carrier wave of 357 metres or 839,000 cycles, the band of 16,000 cycles is equivalent to the range from 353·63 metres to 360·44 metres.

Radio Frequency Amplification and the properties of the circuit. Now, the amplification of the radio frequency depends on the properties of tuned circuits containing inductance, capacity, and resistance, and can be calculated or measured for signals of different frequencies.

C
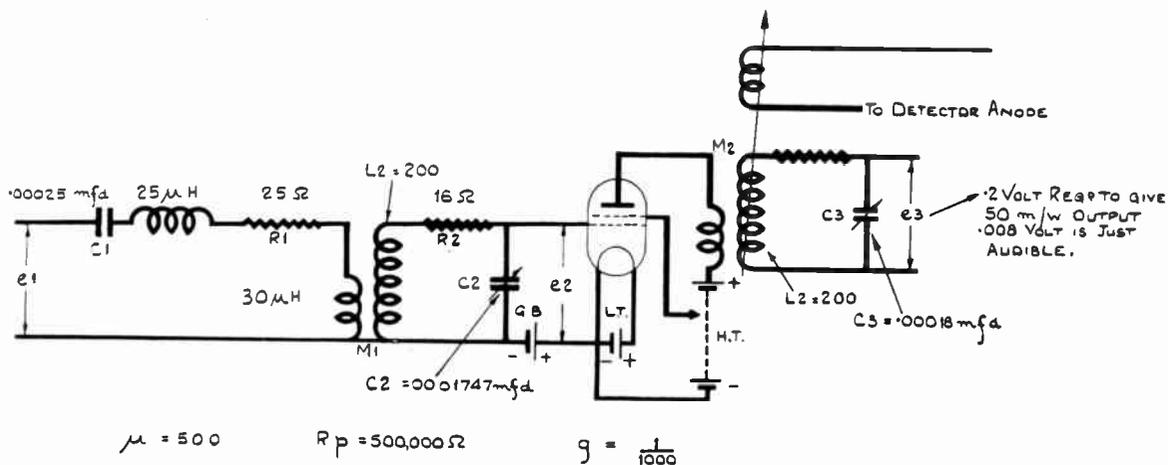
Fig. 13.

Figure 13 shows an aerial, coupled to a tuned circuit in the grid of a screened grid valve, in the plate circuit of which is connected an H.F. transformer with tuned secondary. The latter is connected to the grid of another valve, either an H.F. or detector valve. A weak signal voltage e_1, is received in the aerial and we wish :—

(a) To find the value of e_2, the voltage obtained at the grid of the first valve, or rather to find the ratio $\dfrac{e_2}{e_1}$.

(b) To find the ratio $\dfrac{e_3}{e_2}$ where e_3 is the voltage obtained at the grid of the second valve.

(c) And so obtain the value of $\dfrac{e_3}{e_1}$

Further, since we now realise that a band of frequencies is entering the aerial, we wish to see how the different frequencies affect these ratios. The various values of

inductances and capacities are given in the figure. The circuits are tuned to a carrier wave of 357 metres, or 839,000 cycles, the capacity in the first circuit being slightly less than in the second because of the effect of the aerial capacity.

At the carrier frequency the value of $\frac{e_2}{e_1}$, the other constants being fixed, depends on the resistance of R2 of the first tuned circuit. The curves in Figure 14 show the values of $\frac{e_2}{e_1}$ for two values of R2—namely: R2 = 16 ohms and R2 = 20 ohms.
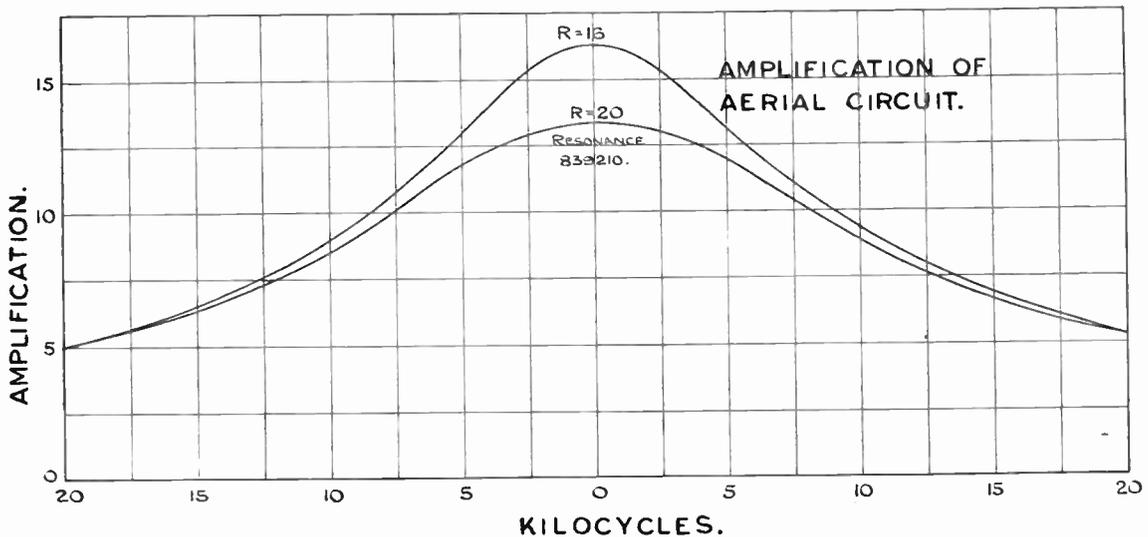
Fig. 14.

When the in-coming signal is 357 metres, that is the carrier wave, for R2 = 20 ohms, we have $\frac{e_2}{e_1}$ = 13·4, but if the signal is 8,000 cycles more (353·63 metres) or 8,000 cycles less (360·44 metres), $\frac{e_2}{e_1}$ is only 9·9. Hence the side bands

Reduction of amplification of sidebands.

corresponding to the 8,000 cycle note are only amplified 74% of the carrier wave. The side bands corresponding to notes of frequency less than 8,000 cycles are amplified more than this, but always less than the carrier wave or side bands corresponding to very low notes. Clearly then, the higher notes in the reproduced music will not be at their full value.

In the second curve of Figure 14 the 8,000 cycle side band is only amplified $\frac{10 \cdot 4}{16 \cdot 3} = 64\%$ of the carrier wave. The reduction of the higher notes is greater because the resistance of the coil is less—namely, 16 ohms.

Now consider the value of $\frac{e_3}{e_2}$.
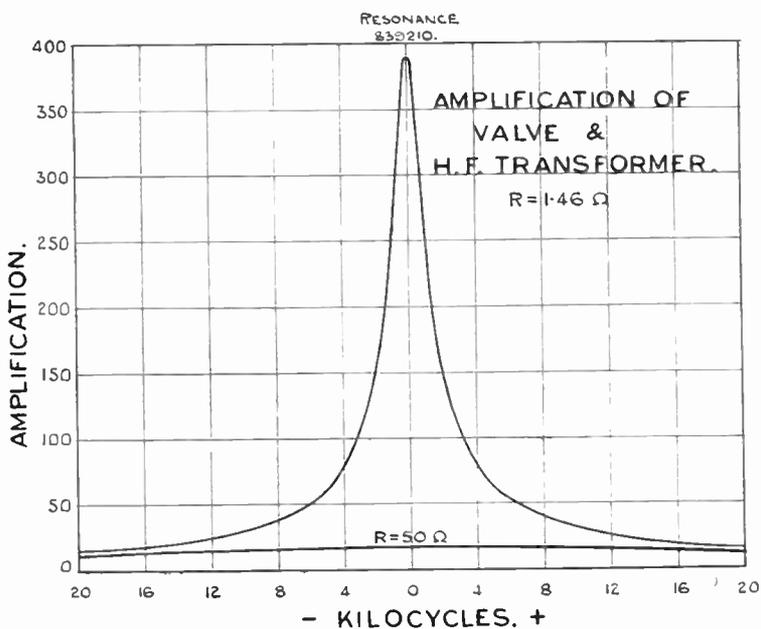
Fig. 15.

Figure 15 shows this for two values of R3, the resistance of the second tuned circuit. When R3 = 50 ohms, $\frac{e_3}{e_2}$ for the carrier wave = 15·6—for the 8,000 cycle side band it equals 14·8. The 8,000 cycle side band is amplified only 95% of the carrier wave.

When R3 = 1·46 ohms (such as can be obtained by the use of reaction), $\frac{e_3}{e_2}$ for the carrier wave = 389, and for the 8,000 cycle side band = 32, which is only $8\frac{1}{4}$% of that for the carrier wave.

Our main interest is, however, in the values of $\frac{e_3}{e_1}$, the total amplification ratio. These are given for two cases.
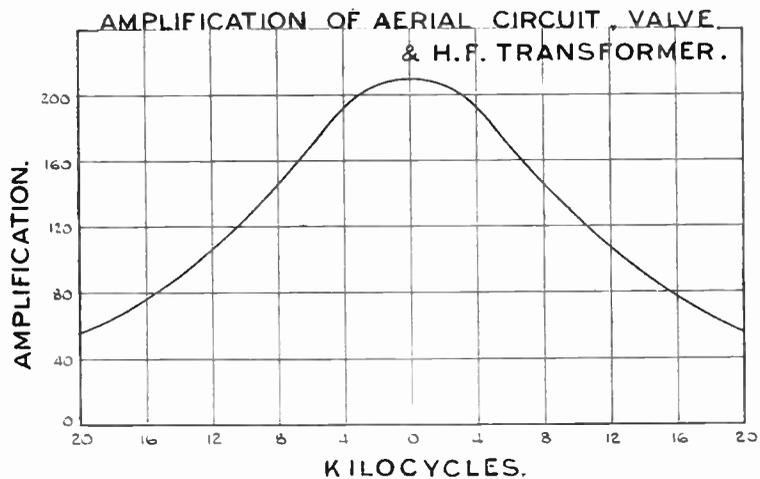
Fig. 16 (a).

(a) When the resistance R2 of the first circuit is 20 ohms and R3 of the second is 50 ohms.

The curve of Figure 16A shows these values. You will note that the 8,000 cycle side band is amplified 71% of the carrier wave.
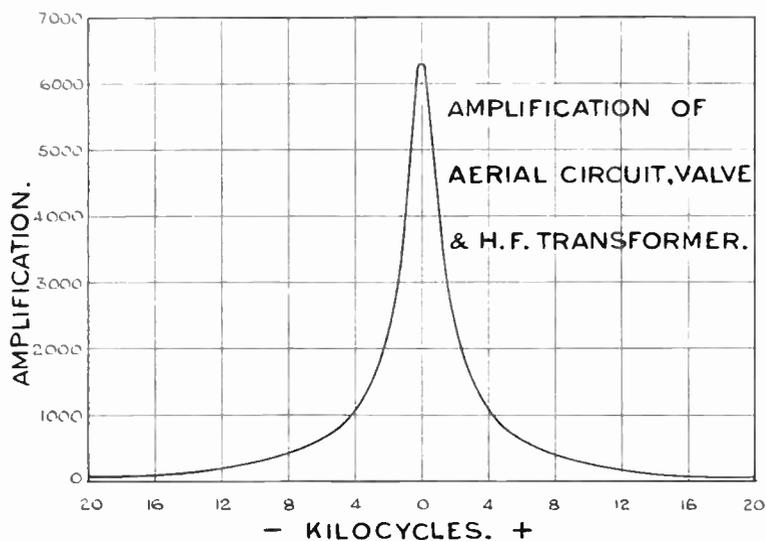
Fig. 16(b).

(*b*) When R2 is 16 ohms and R3 is 1·46 ohms.

The curve of Figure 16B shows the values. In this case, the 8,000 cycle side band is only amplified $5\frac{1}{2}\%$ of the carrier wave.

Clearly in the case (*a*) there is some, but not a serious, loss of the higher notes. In case (*b*) the loss is very great, so much so, that music under those conditions would lose its " brilliance."

The extent of the defect is perhaps better shown in Figure 17, which gives the strengths of the various notes from 50 cycles to 8,000 cycles relative to their proper value—namely, 100%—which would obtain in the receiver presuming that there were no further distortion. Curve 2 corresponds to case (*a*) and curve 1 to case (*b*). The same two curves are given in Figure 18 in the form of decibels loss for each note from its proper level or value.

AVERAGE MALE SPEAKING VOICE
AVERAGE FEMALE SPEAKING VOICE
HARMONIC RANGE
PIANOFORTE
CELLO
VIOLA
CLARIONET
CORNET
VIOLIN
OBOE
FLUTE
HUMAN VOICE
HORN

AMPLIFICATION PER CENT.

II

I

100
96
92
88
84
80
76
72
68
64
60
56
52
48
44
40
36
32
28
24
20
16
12
8
4
0

50   100   300   500   1000   2000   4000   6000   8000

FREQUENCY IN CYCLES PER SECOND

Fig. 17.

HARMONIC RANGE
ORGAN AND PIANO
CLARIONET
OBOE
FLUTE
CORNET
HORN
HUMAN VOICE

AVERAGE FEMALE SPEAKING VOICE
AVERAGE MALE SPEAKING VOICE

VIOLIN
VIOLA
CELLO

LOSS IN DECIBELS.

0

10

20

30

II

I

CYCLES   PER   SECOND

25   50   62½   100   150   300   500   1000   2000   4000   6000   8000

Fig. 18.

The figures given above are for one stage of radio frequency amplification, but all that has been said applies to two or more stages. *Low resistance circuits, unless other correcting devices are used, mean the loss of the higher notes.*

Reasons for use of low resistance circuits.

The reason for the use of low resistances which lead to such distortion will be evident from the comparison of Figures 16A and 16B. When the resistances of the two circuits are respectively 20 and 50 ohms, the maximum amplification of the weak aerial signal is 209 times; but when the resistances are respectively 16 and 1·46 ohms, the maximum amplification is 6,320 times. The desire to receive weak stations, therefore, leads to the use of low resistance circuits with a much greater amplification, despite the distortion.

Further, when receiving a station, say, on 357 metres (839,100 cycles), in the present crowded state of the ether there will be stations 9,000 cycles and 18,000 cycles greater or less than this in frequency. Assume that when the circuit is tuned to the desired station at 357 metres, the signal of another station 18,000 cycles distant (either 364·8 or 349·5 metres) is entering the aerial at the same time.

From Figure 16A (R2 = 20 ohms, R3 = 50 ohms), we see that the unwanted station is amplified 65 times compared with 209, the figure for the wanted station, a reduction of $3\frac{1}{4}$ to 1. Hence, it is likely that if the two stations are of anything like equal strength in the aerial, both would be heard together, but if R2 = 16 ohms and R3 = 1·46 ohms, we see a different state of affairs. The unwanted signal 18,000 cycles off tuning, would only be amplified 70 times, against 6,320 for the wanted station, a reduction of 90 to 1.

Hence, it will be easy to receive the 357 metre station without a trace of the other, even though this is of great strength.

*Remedy for
H.F. distortion.
Band Pass
Filters or
Correctors.*

The low resistance circuits are said to give greater " selectivity," and this is an additional reason for their use. What, then, is the remedy for this distortion? The use of " band pass " filters in the H.F. circuits and " correctors " in the audio frequency side, both of which will be described later, can give considerable improvement, but have not been adopted in most commercial sets.

Most manufacturers have spent all their efforts in obtaining a good degree of amplification and great selectivity—which a portion of the public demands—and have paid little or no attention to the distortion caused thereby. These special devices mentioned are costly, necessitate the use of an additional valve, and add complication and difficulties in manufacture. Also, of greater importance, most of the people demanding this selectivity do not yet know the bad effect it has.

The following general statement is true of all British commercial sets (some American sets are now using band pass filters).

No selective set, or one used in a selective condition, can reproduce the higher musical notes effectively, and the greater the selectivity, the greater the loss of these higher notes. A set advertised to have " razor edge selectivity " must produce music which has lost most of the higher notes and its brilliance.

In the Models A.C. 31 and 32 manufactured by Ferranti, the difficulty has been partly met, by making reaction reversible. Low resistance, with its great selectivity, is always obtained by reaction, intentional or unintentional, but if the effect can be reversed, the resistance of the circuit is increased, not reduced. In these receivers, when receiving moderately strong stations not requiring great

amplification, the reaction control is turned backwards to increase the resistance so that there is only a small loss of the high notes. On the other hand, by turning the control forward, as great a degree of selectivity and amplification can be obtained as in any other three valve set with no greater loss of high notes than is necessary in any set of a corresponding degree of selectivity.

In other sets, the difficulty has been met by making the coils of the lowest resistance possible and inserting a resistance in series with each which can be shorted by a switch. When receiving strong stations, the switch is open, resulting in a high resistance coil and quality. When receiving distant stations, the switch is closed, resulting in a coil of low resistance, loss of quality, but much greater amplification and selectivity.

We have described generally how radio frequency amplification is obtained, the difficulties involved, and precautions necessary, and the manner in which it introduces distortion.

For many readers, this general survey will suffice to give them some real insight into this part of their receiver. For others who wish to delve deeper, much additional information is given in the various paragraphs of the next section.

Limitations of Radio Frequency Amplification. From the foregoing, it might appear that in the screened grid valve we have a means of obtaining any desired degree of amplification and could receive any signal, however weak, by using sufficient stages. There are, however, definite limitations. On page 27 the possibility of self oscillation when the amplification is great enough is mentioned, due to the residual anode to grid capacity which, though so minute, has still some regeneration effect. With care this does not prevent an amplification of a million, or even two millions.

The actual limitation comes from an unexpected quarter. We are apt to look upon valves and copper wire coils as very stable things ; that no variations occur in them except such as we apply to the grid of the valve, presuming energy supplies are steady. There are, however, in both, variations occurring which, though minute, become important when the amplification is great enough.

The emission of electrons from the hot filament is analogous to ebullition or boiling, occurring in gusts at extremely short intervals of time, and though the effect must be averaged because of the large area of the filament, there is still left superimposed upon the mean flow of electrons to the plate a minute irregular variation of very high frequency.

Schottky flicker effect.   This is known as the " Schottky flicker effect."

Schottky temperature effect.   There is a similar fluctuation of current in the electrical conductors, due to the thermal agitation of its ions and electrons (temperature is due to this agition and consequent collisions.) This is known as the " Schottky temperature effect." For details, reference should be made to a Paper by Ballantine, page 1,377 of *Proceedings of the Institute of Radio Engineers,* August, 1930.

The combined effects occurring in the first radio frequency valve and its associated circuits are amplified. In the detector circuit, the components of the disturbances beat with each other and appear in the speaker as noise, usually a hissing noise. When receiving a carrier wave, the trouble is much worse. There seems to be a beating with this, or, in other words, some modulation, which greatly increases the noise in a sensitive set, so that there is a very unpleasant background to the music.

Sets have been built in America which, on a wave length of 214 metres (their most sensitive point), give full standard output of 50 milliwatts (noise, not music) with no input ;

obviously, such a set is unsuitable for musical reproduction. The above phenomena definitely limit the radio frequency amplification that can be used, and from the rather limited experience we have had at the time of writing, we think that the amplification should not exceed 75,000.

DEFINITIONS OF SELECTIVITY AND QUALITY IN A SET.

We have chosen the following definitions as being most convenient.

Selectivity. Selectivity is the ratio of a tuned signal to that received when the receiver is detuned 10,000 cycles from the tuned frequency. Selectivity is a number greater than one.

Quality. Quality is the ratio of the signal received when the set is detuned 5,000 cycles to that received when fully tuned.

It is a number usually less than one—alas, often very much less—and we naturally desire it to be as near one as possible.

Referring to Figure 16A, giving values of $\frac{e_3}{e_1}$ when R2 = 20 and R3 = 50 ohms, we see that at the tuned point $\frac{e_3}{e_1}$ = 209 ; at 5,000 cycles off tuning $\frac{e_3}{e_1}$ = 100, and at 10,000 cycles off tuning $\frac{e_3}{e_1}$ = 125, hence, according to the definition :—

$$\text{selectivity} = \frac{209}{125} = 1\cdot67$$
$$\text{quality} = \frac{180}{209} = \cdot86$$

Hence, in this case, there is little selectivity, but the quality is high.

Now consider the case of R2 = 16 and R3 = 1·46 ohms (Fig. 16B), $\frac{e_3}{e_1}$ at resonance tuning = 6,320, at

5,000 cycles off tuning $\dfrac{e_3}{e_1} = 800$, at 10,000 cycles off

tuning $\dfrac{e_3}{e_1} = 210$:—

$$\text{selectivity} = \frac{6,320}{210} = 30$$
$$\text{quality} = \frac{800}{6,320} = \cdot 127$$

Now, there is a great degree of selectivity since an unwanted signal 10,000 cycles away from the wanted one would be reduced 30 times. The quality is bad, since the 5,000 cycles note is reduced to ·127 of its proper value.

Difficulties when near to powerful Stations.

What has been said in the last paragraph would be sufficient if the signal voltage e_1 received in the aerial, was never very large. But the introduction of the high power stations, and in particular the new Regional Stations, has changed the problem of selectivity. A great many people will be receiving in their aerial (good outdoor aerials) a signal voltage e_1 of the order of ·500 volt (500,000 microvolts), while the number of those who will receive $e_1 = \cdot 2$ of a volt will be very large. It now becomes necessary to consider the amplification of the receiver at wave lengths differing very much more than 10,000 cycles from its tuned position.

Figure 19 shows the values of $\dfrac{e_3}{e_1}$ for signals of wave lengths from 250 metres to 500 metres for the circuits of Figure 13 used previously for illustration, the circuits being tuned to 358 metres. The various formulæ and means for calculating these and those previously used are given below for those further interested. They show that when the incoming signal is more than 3% off resonance—that is, when the tuning point is 358 metres—the signal is less than 347 metres or more than 370 metres, *the value of* $\dfrac{e_3}{e_1}$ *does*

AT 358 METRES = 6340

AMPLIFICATION OF AERIAL CIRCUIT, VALVE & H. F. TRANSFORMER.

WAVE LENGTH IN METRES.
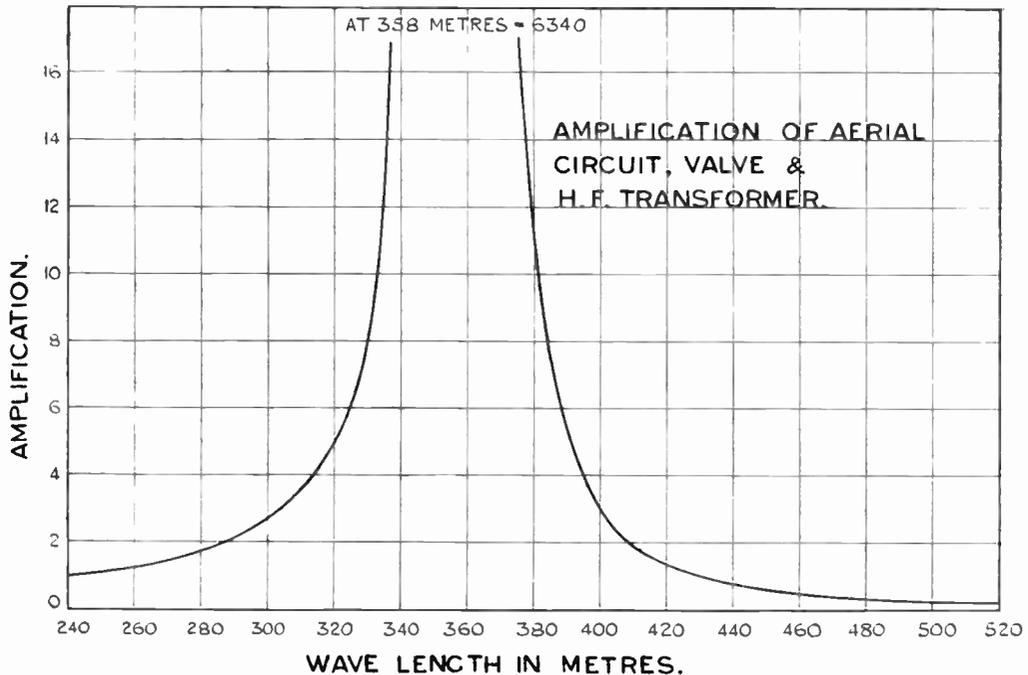
Fig. 19.

not depend in any way on the resistance of the coils, but only on inductances and coil couplings which are fixed for the set and the capacities which are fixed by the tuning. The curve of Figure 19 is therefore true when the amplification at tuning is 209 (R2 = 20, R3 = 50) or 6,320 (R2 = 16, R3 = 1·46). At first sight this seems a surprising result, and is probably not generally recognised, but nevertheless it is certain.

It will be noted that when the incoming signal is 250 metres the amplification is 1·2, rising as the signals approach tuning frequency, to 209 or 6,320, according to the resistances, falling again until, at 430 metres, it is 1 again, and at 500 metres it is still ·25.

It is found that in the A.C. Receiver, Model 32, with the volume control full in, and tuned to 358 metres, ·2 volt (radio frequency with 30% modulation) across the grid of the detector valve gives full standard output, and that ·008 volt gives a signal still audible but not enough to be annoying when listening to another station. Since a 500 metre signal is amplified one quarter, were it of strength equal to ·03 volt in the aerial, it would still be heard, since one quarter of ·032 = ·008, which is audible.

Similarly, a signal of ·0067 volt of 250 metres wave length would be heard. Thus anywhere near a strong station, this station is likely to be heard over the whole of the tuning range. Apparently the set is useless in the circumstances; but it contains, however, in its volume control the means of overcoming this difficulty. This is a series condenser in the aerial connection, which is shorted out automatically when full in. When not just shorted, its value is ·0028 mfd.

Turning this volume control to the left, the value is reduced.

Operation of Receiver A.C. 32.

The effect is to reduce all signals entering the aerial. This is the first step; but we now desire to strengthen the particular station we wish to receive. The solution is indicated in the previous paragraphs, where we show that decreasing the resistance of the coils only affects the amplification of the signal to which the circuits are tuned. Hence, turning the reaction control to the right to decrease the resistance of the coil, will only increase the wanted station.

The best way to operate this set is, therefore, to turn the volume control to the left to reduce all signals, turn the reaction control to the right to strengthen the station to which you are tuned and which you desire, but the operations must not be carried too far, since the excessive use of reaction must result in deterioration of quality.

This method has proved very good in practice. Near Manchester, where 2ZY on 377 metres gives a signal strength of 117 millivolts in the aerial, one can easily receive London (356 metres) and stations on or above 400 metres. It must be confessed, however, that measurements with the receiver and the known properties of the circuits would not lead one to expect such results. When two stations come in together into the aerial and receiver, there is something quite independent of the H.F. circuits which increases the selectivity. This will be dealt with in the next section.

Mathematical Section.

Formulæ used for previous calculations. Refer to Figure 13 for the meaning of symbols.

For the loosely coupled aerial circuit :—

$$(1)\ \frac{e_2}{e_1} = \frac{\dfrac{M_1}{C_2}}{Z_2\sqrt{\left\{R_1 + \left(\dfrac{\omega M_1}{Z_2}\right)^2 R_2\right\}^2 + \left\{\left(\omega L_1 - \dfrac{L}{\omega C_1}\right) - \left(\dfrac{\omega M_1}{Z_2}\right)^2\left(\omega L_2 - \dfrac{1}{\omega C_2}\right)^2\right\}}}$$

$$\text{Where } Z_2 = \sqrt{R_2{}^2 + \left(\omega L_2 - \dfrac{1}{\omega C_2}\right)^2}$$

For the H.F. transformer in the plate circuit of the valve whose plate resistance is R_p and mutal conductance g.

$$(2)\ \frac{e_3}{e_2} = \frac{\dfrac{M_2}{C_3}g}{Z_3\left(1 + \dfrac{\omega M_2{}^2}{Z_3{}^2}\dfrac{R_3}{R_p}\right)}$$

$$\text{Where } Z_3 = \sqrt{R_3{}^2 + \left(\omega L_2 - \dfrac{1}{\omega C_3}\right)^2}$$

From these $\dfrac{e_3}{e_1}$ can immediately be calculated.
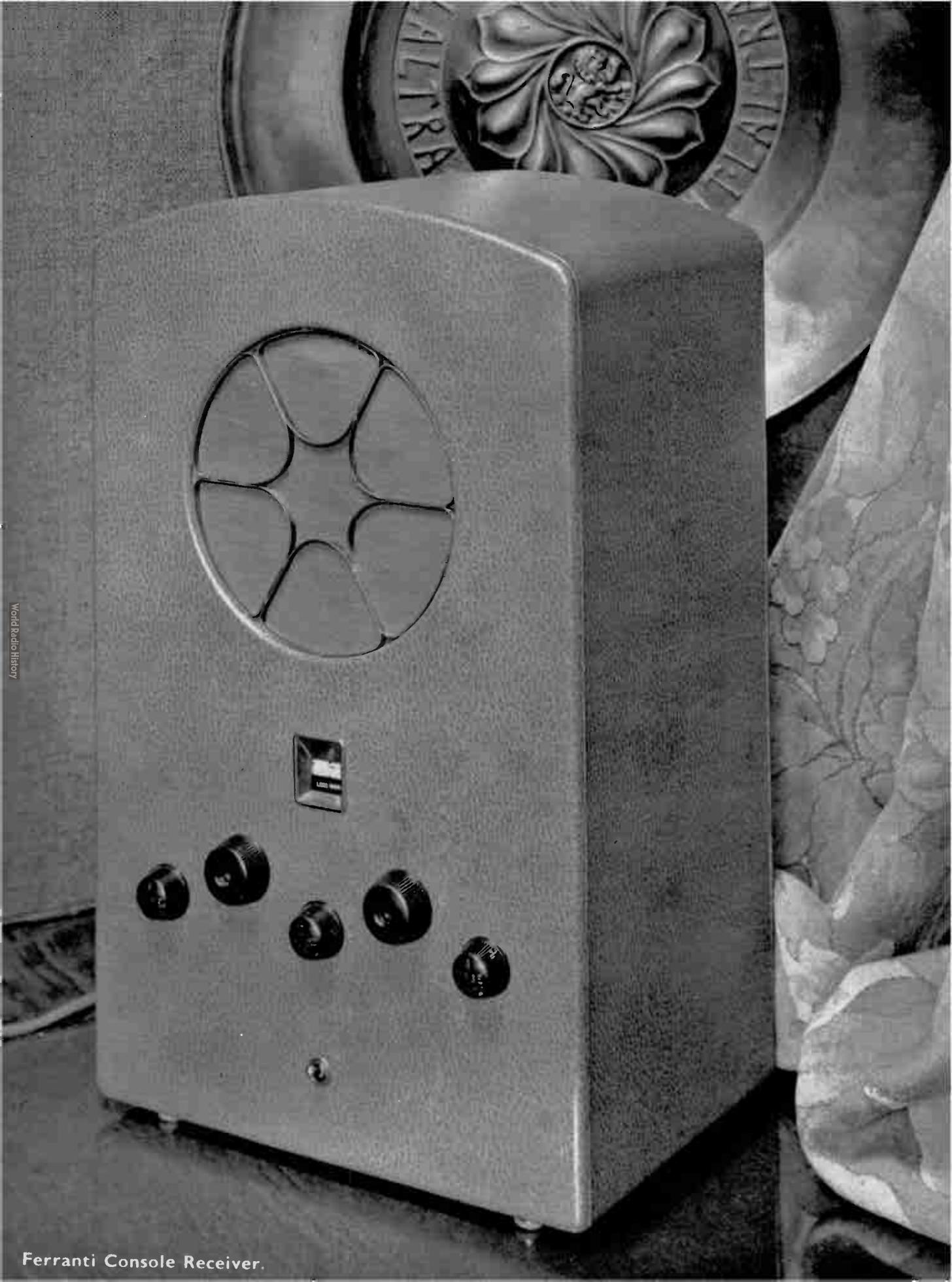
Ferranti Console Receiver.

When the incoming signal is more than 3% off the resonant frequency the above formulæ simplify

$$(1) \quad \frac{e_2}{e_1} = \frac{\dfrac{M}{C_2}}{\left[\omega L_2 - \dfrac{1}{\omega C_2} \right] \left\{ \left[\omega L_1 - \dfrac{1}{\omega C_1} \right] - \dfrac{\omega^2 M_1^2}{\omega L_2 - \dfrac{1}{\omega C_2}} \right\}} \quad (3)$$

$$(2) \quad \frac{e_3}{e_2} = \frac{M_2 g}{C_3 \left[\omega L_2 - \dfrac{1}{\omega C_3} \right]} \quad (4)$$

$$\text{and} \quad \frac{e_3}{e_1} = \frac{\dfrac{M_1}{C_2} \times \dfrac{M_2}{C_3} \times g}{\left[\omega L_2 - \dfrac{1}{\omega C_3} \right] \left[\omega L_2 - \dfrac{1}{\omega C_2} \right] \left[\omega L_1 - \dfrac{1}{\omega C_1} - \dfrac{\omega^2 M_1^2}{\omega L_2 - \dfrac{1}{\omega C_2}} \right]} \quad (5)$$

It will be noted from these formulæ that when off resonance the amplification does not depend in any way on the resistance of the coils, only on the inductances, couplings and capacities of the circuits. Only when very near resonance, or exactly at resonance, do the resistances and, therefore, the amount of reaction used, play a part.

This is the justification of the instructions for separating stations and the statements used earlier in this section.

Demodulation. THE DEMODULATION OF A WEAKER BY THE CARRIER WAVE OF A STRONGER STATION, or increase of selectivity when two stations not differing greatly in wave length are received together.

This action is discussed by Beatty, page 300 of *Experimental Wireless,* June, 1928, who, because of an error in his assumptions, makes the effect appear even more wonderful

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than it really is. A complete theory for a perfect linear rectifier is given by Butterworth on page 619 of *Experimental Wireless,* November, 1929, though his mathematics are somewhat difficult for the ordinary man to follow.

In popular language, the carrier wave of the strong station heterodynes with that of the weak station and transfers a very large part of the modulation of the latter to the intermediate heterodyne or supersonic frequency, equal to the difference of the frequencies of the two stations, which, since the set, unlike the superheterodyne set, has no second detector, never becomes audible.

Recently, accidental circumstances have proved how great the effect can be. The receiver on which the results were noted was made up of two stages of H.F. amplification with screened grid valves, anode bend detector with a milliameter in plate circuit. This was receiving Manchester and owing to its great strength the aerial circuit was detuned and the other two circuits detuned $3\frac{1}{2}°$, one above and the other below resonance to get band pass effect. The milli-ammeter was reading 1 ma, corresponding to 4 volts R.M.S. on the grid of the detector, so that rectification approached linearity. On many occasions, at 8.55 p.m., because of the early finish of its programme, Manchester has shut off its carrier wave. As long as the carrier wave is on, there is no sign of another station, but as soon as it was shut off, Radio Toulouse came in so loud as to be heard with ease in another room. As the carrier wave was put on again, Radio Toulouse disappeared. The carrier wave of Manchester sufficed; it was possible to see in the milliammeter Manchester's carrier wave fall, and when the needle was quite low down, Radio Toulouse could be heard. After a few minutes, the needle again rose as Manchester's carrier wave grew in strength and the Radio Toulouse signal became inaudible.

The effect has been reproduced by tuning Manchester to moderate strength on an indoor aerial and then operating an unmodulated oscillator. Though the latter could not give a strong signal in the aerial, it silenced the Manchester signal for 9,000 cycles less and 18,000 cycles more than the carrier frequency.

This demodulation effect is useful in another direction. A receiver with a high degree of amplification brings in a lot of noise. Some, as explained, is due to the first circuit and valve, and the remainder comes into the aerial. The effects, however, are due to irregular modulation of carrier waves. If such a receiver with good rectification is tuned to a strong station, it is found that as the station tunes in, all noise disappears. The effect can also be produced by feeding an unmodulated oscillator strongly into the aerial. As the carrier of the oscillator is tuned in, the noise disappears.

It would be useful to get a quantitative measure of this demodulating effect. Mathematics help us to the extent that if the detector rectifier follows a square law, such as it must if the signals are small and the audio frequency amplification great, and two stations not differing greatly in frequency, and modulated to the same extent, be received together, the audio output of each is quite separate and proportional to the square of the relative strengths of the signals at the grid of the detector valve. Here there is no demodulating effect ; the output for each station is the same as if the other did not exist—if one station is $\frac{1}{10}$ of the strength of the other, its output is only $\frac{1}{100}$.

On the other hand, if the rectifier is linear and perfect, then there is a considerable demodulating effect. Thus, if a station is $\frac{1}{10}$ of the strength of another at the grid of the detector, then if each came in separately the relative strengths of the audio output would also be 1 to 10 ; but if they come

in together, then the outputs of the weaker to the stronger are in the ratio of 1 to 167. The weaker station has been demodulated to $\frac{1}{16\cdot7}$ of its proper value, while the stronger station can be shown to have been demodulated less than 1%. By graphical comparison with the case of a perfect rectifier, we can see that the result for anode bend rectification with large signals (4 volts R.M.S. or more with modern valves), and for grid leak rectification for signals of ·3 volt and above, for valves of the MH4 and MHL4 class, must be very nearly the same as with the perfect rectifier.

This demonstrates clearly the gain, from the viewpoint of selectivity, of using a moderate degree of audio frequency amplification so that the rectification is of a sufficiently large radio frequency signal to come into the above ranges.

Beatty, without sufficient justification, makes the following general statement, which is true of many good modern sets or can be made true by proper design : " The unwanted station is silenced, if its carrier wave does not exceed one tenth of the wanted station at the grid of the detector valve."

RECEPTION OF TWO STATIONS TOGETHER IN A RECEIVER.

Mathematical Section.

The two stations represented by (1) $E_0 \sin \omega t$.
(2) $E_1 \sin (\omega + \delta)t$.
where δ is small compared with ω.
E_0 and E_1 contain the modulation effects at audio frequency which are small compared with $\frac{\delta}{2\pi}$

$$\text{Thus } E_0 = (1 + M \cos pt) e_0$$
$$\text{and } E_1 = (1 + m \cos qt) e_1$$

First Case.

SQUARE LAW RECTIFIER, following the law
current $Z = K(e + \alpha)^2$
Let the single station $e = E_0 \sin \omega t$ be received.

Then the mean rectified current over one complete cycle ω is $\frac{1}{2}(E_0)^2 \times K$.

With a suitable plate load we can say that output voltage is proportional to E_0^2 is, say, BE_0^2.

Similarly, when station $E_1 \sin (\omega + \delta)t$ is received alone, the output will be BE_1^2.

Now, if both stations be received together, we get at the grid of the detector

$$e = E_0 \sin \omega t + E_1 \sin (\omega + \delta)t$$
$$= (E_0 + E_1 \cos \delta t) \sin \omega t + E_1 \sin \delta t \cos \omega t.$$
$$= R \sin (\omega + \Phi)t$$

where $R \cos \Phi = E_0 + E_1 \cos \delta t$, $R \sin \Phi = E_1 \sin \delta t$

and $R^2 = E_0^2 + E_1^2 + 2 E_0 E_1 \cos \delta t$

$$R = \sqrt{E_0^2 + E_1^2 + 2 E_0 E_1 \cos \delta t}$$

The combined signal represented by
$$e = R \sin (\omega + \Phi)t.$$

This is a radio frequency signal of varying frequency, rising a little above $\dfrac{\omega}{2\pi}$ and falling a little below it, by the amount $\Phi = \tan^{-1} \dfrac{E_1}{E_0}$
and having the varying amplitude R, the variations being slow compared with $\dfrac{\omega}{2\pi}$.

The curve $R = \sqrt{E_0^2 + E_1^2 + 2 E_0 E_1 \cos \omega t}$ represents the envelope of the tips of the carrier waves.

Now as above, the output voltage is
$$B R^2 = B (E_0^2 + E_1^2 + 2 E_0 E_1 \cos \delta t)$$

But the portion $2 E_0 E_1 \cos \delta t$ has a frequency of $\frac{\delta}{2\pi}$ and is above audio frequency; hence it does not affect the loud speaker. The output voltage which appears is $B(E_0^2 + E_1^2)$. Clearly each station affects the speaker and has the same strength as if the other did not exist. There has been no demodulation, but by the very nature of the rectifier, one station is $\frac{1}{10}$ as strong as another at the grid of the detector valve, then its output is only $\frac{1}{10} = \frac{1}{100}$ of that of the stronger.

Second case. PERFECT LINEAR RECTIFIER. Because the relation between volts and rectified current follows the straight line law and zero grid signal is adjusted to the point where this line cuts the zero current line, it follows that the peaks of the pulses of the rectified current follow an exactly similar form to the peaks of radio frequency variations of the grid volts; but when both signals are received this form has been shown to be

$$R = \sqrt{E_0^2 + E_1^2 + 2 E_0 E_1 \cos \delta t}$$

where $\frac{\delta}{2\pi}$ is the difference of the frequencies of the two stations. If both stations are modulated

$$E_0 = (1 + M \cos pt) c_0$$

$$E_1 = (1 + m \cos qt) e_1$$

where $\frac{p}{2\pi}$ and $\frac{q}{2\pi}$ are audio frequencies.

Hence, output voltage is proportional to the mean of

$$\sqrt{E_0^2 + E_1^2 + 2E_0 E_1 \cos \delta t}$$

FIRST : If each is received separately, the output voltage for the first is equal to KE_0 and for the second is equal to KE_1

If each were modulated to the same extent, the output in the speaker would be just in proportion to the strength at the grid of the detector valve.

SECOND : Combined signals.

The form $\sqrt{E_0{}^2 + E_1{}^2 + 2\,E_0 E_1 \cos \delta t}$ introduces serious mathematical difficulties into the solution of our problem. Butterworth gives a solution and obtains numerical values (see *Experimental Wireless*, 619, November, 1929), using elliptic integrals, true only for very small percentages of modulation. The writer prefers the following method.

Following Aiken (*Proc. I.R.E.*, January, 1931) by a long and wearisome use of the binomial theorem, we can expand

$$R = \sqrt{E_0{}^2 + E_1{}^2 + 2E_0 E_1}, \cos \delta t \text{ using}$$

$$E_0 = (1 + M \cos pt)\, e_1$$
$$\text{and } E_1 = (1 + m \cos qt)\, e_0$$

into (provided $\dfrac{e_1}{e_0}$ is less than 1)

$$R = \left\{ Me_0 - e_1 g\left[a_0 M - a_1 + a_2\,\frac{M}{2} \right] + \frac{m^2 e_1{}^2 M g^2}{2e_0} \right\} \cos pt.$$

$$+ \left\{ me_1 - e_1 g\left[a_0 m - a_1\,\frac{Mm}{2} + \frac{m e_1 g}{e_0} \right] - \frac{3e_1{}^2}{2e_0} g^3 b_0 m \right\} \cos qt.$$

$$+ \left[\frac{m^2 e_1}{2e_0}\, g^2 - \frac{b_0\, e_1{}^2\, g^3\, m^2}{4e_0} \right] \cos 2\, qt.$$

$$+ \left\{ e_1 g\left[a_0 - \frac{a_1 M}{2_1} - \frac{m^2 c_1 g}{2c_0} \right] + \frac{b_0 e_1{}^2 g^3}{2c_0} (2 + m^2) \right\} \cos \delta t.$$

$$- \frac{b\, e_1{}^2\, g^3}{4e_0} \cos 2\, \delta t + e_1 g\left[\frac{a_0 M}{2} - \frac{a_1}{2} + \frac{a_2 M}{4} - \frac{m^2\, M e_1 g}{4e_0} \right]$$
$$\cos (p \pm \delta)\, t.$$

$$+ \left\{ e_1 g\left[\frac{a_0 M}{2} - \frac{a_1 Mm}{4} - \frac{m e_1 g}{2c_0} \right] + \frac{b_0 e_1{}^2 g^3 m}{c_0} \right\} \cos\left[q \pm \delta \right] t.$$

Where $a_0 = 1 + \dfrac{M^2 g^2}{2} + \dfrac{3 M^4 g^4}{8}$

$a_1 = Mg + \dfrac{3 M^3 g^3}{4} + \dfrac{5 M^5 g^5}{8}$

$a_2 = \dfrac{M^2 g^2}{2} + \dfrac{M^4 g^4}{2}$

$b_0 = 1 + 3 M^2 g^2$

$g = \dfrac{e_0}{e_0 + e_1}$

neglecting further terms in the series which only affect the result to less than 1%.

Now, again, the terms involving δ, $p \pm \delta$, $q \pm \delta$ are in our case above audio-frequency and do not appear in the speaker. The output voltage at the speaker is proportional to—

$$R = \left\{ Me_0 - e_1 g \left[a_0 M - a_1 + a_2 \frac{M}{2} \right] + \frac{m^2 e_1{}^2 M g^2}{2e_0} \right\} \cos pt$$

$$+ \left\{ me_1 - e_1 g \left[a_0 m - a_1 \frac{Mm}{2} - \frac{me_1 g}{e_0} \right] - \frac{3 e_1{}^2 g^3 b_0 m}{2e_0} \right\} \cos qt.$$

$$+ \left\{ m^2 \frac{e_1{}^2 g^2}{2e_0} - \frac{b_0 e_1{}^2 g^3 M^2}{4e_0} \right\} \cos 2 qt.$$

Clearly, the two modulations are intermixed, and a second harmonic of the weaker station is introduced.

Now consider the weaker station to be $\frac{1}{10}$ of the stronger or $\dfrac{e_1}{e_0} = \frac{1}{10}$ and that each is modulated 20% or $m = M \times \cdot2$, we find then that the factor of $\cos pt$ is within $(1 - \frac{1}{120})$ of the value Me_0. Hence, the modulation of the wanted station is reduced less than 1% and the effect of the terms

in it involving the modulation of the unwanted station is exceedingly small. The factor of the term cos qt is equal to $me_1 \times \cdot 06$ or $\frac{1}{16 \cdot 7}$ th part of the modulation of the wanted station; we have, therefore, reduced the unwanted station to $\frac{1}{16 \cdot 7}$ part of the value it would have had if it came in alone. This is a measure of the demodulating effect. The second harmonic term is only $\frac{1}{5000}$ part of the modulation of the wanted station, and can be neglected. The method of rectification in modern sets is not, or should not be, square law, and unfortunately cannot be linear.

Anode Bend Rectification. Using strong signals, the portion of the curve on which the variations occur of either the modulation (if not too great), and the beat note due to the combination of the two stations is straight. Hence, the tips of the pulses of rectified current follow the same law as the tips of the grid voltage R.F. variations, so this can be expanded into just the same terms as before. The percentage of the maximum height of the pulses to give the mean rectified current is not constant for all values of this height, due to the bend at the lower end of the valve curve; but if the weak station is only $\frac{1}{10}$ of the strong one and the average modulation not more than 20%, then the factor is sufficiently constant to ensure that the demodulation is of the order just calculated. The qualitative experiments described indicate this. The exact percentage of demodulation cannot be calculated.

Grid Leak Rectification. It has been shown, page 112, that when the R.F. signal exceeds $\cdot 25$ volts R.M.S. at the grid of the detector, the ordinary average modulation variations occur on a part of the grid current curve which enables the mean rectified current to be an accurate copy of the peaks of the radio frequency grid volt variation. The same will apply to the beat variations due to two stations, if one is much weaker than the other. Hence, in grid leak rectification, under the conditions specified above, the demodulation is very nearly that calculated for a perfect linear rectifier.

Cross Talk or Modulation. This is a form of interference between two stations which can, in special circumstances, occur. If, when tuned to a moderately weak station, a very strong station—say 50 kilocycles away—is received, and at the grid of one of the H.F. screened grid valves its signal reaches a strength of $\frac{1}{2}$ volt or more, whilst this valve has too much grid bias, then some small rectification of the unwanted strong signal occurs, due to curvature of the valve characteristics, and modulates the wanted signal (it increases and decreases its amplitude slightly, just as the music in the studio does).

Hence, no amount of selectivity afterwards will separate the two programmes. It would be difficult to design a set in which it could occur, since in most cases you would hear the unwanted station, due to its mere strength, even if the weaker one did not exist. It is conceivable, if the aerial was connected direct to the grid of the first valve through a resistance or inductance and other working conditions were wrong in the set. It would not have been mentioned here if the Press had not been dealing with it.

The self evident remedy is to design the set properly, using a good tuned loose coupled circuit between the aerial and the first screened grid valve, and suitable means—a series aerial condenser, or a wave trap—to reduce the signal input to this first tuned circuit.

Correction for loss of higher Audio Frequency Notes due to selectivity. Here we will deal with the possibility of correction on the H.F. side of a receiver.

For such correction, one ought to have available two stages of radio frequency amplification, since any arrangement for correction involves loss of amplification. The extra valve is necessary to make up for this loss. In the descriptions to follow, we shall assume that the receiver has two stages.

It will be realised from an earlier section, that the ideal receiver, when tuned to a carrier wave, should give uniform

amplification for 8,000 cycles each side of the carrier frequency, to give perfect quality, and then at 10,000 cycles each side to give no amplification, to achieve perfect selectivity. But even if such an ideal could be realised, it would not be satisfactory at present. Stations are only 9,000 cycles apart. Hence, the side bands of the two stations will overlap, and there is bound to be interference.

The best compromise would be uniform amplification for 5,000 cycles each side of the carrier frequency and then a rapid fall of amplification to zero at 10,000 cycles each side of the carrier frequency. Even this compromise cannot be realised in practise.

Partial Correction by suitable Detuning.

Amateurs have for some years known of a method of reducing high note loss in such a receiver as shown in Figure 12, where there are three tuned circuits in series. It can only be done when the station wanted is much stronger than necessary. The first circuit is tuned to the carrier wave, and the other two are detuned equal amounts each side of the carrier wave. A description of the actual process carried out on a receiver may be useful. When receiving Manchester, the first circuit is well detuned, since the station is so powerful. The other two circuits are then sharply tuned, as judged by the maximum reading on the detector valve milliammeter. The readings of the two condensers are then $116°$ and $118\frac{1}{2}°$. The first is adjusted to $119\frac{1}{2}°$ and the other to $115°$—that is, $3\frac{1}{2}$ degrees each way.

This is equivalent to 10,000 cycles each way. This figure can be obtained from the curve connecting degrees of condenser setting to wave lengths. Now, the first condenser is adjusted till the milliammeter reads 1 milliamp., giving in this case the best detection. At the same time the audio frequency volume control is adjusted to give the necessary volume in the loud speaker.
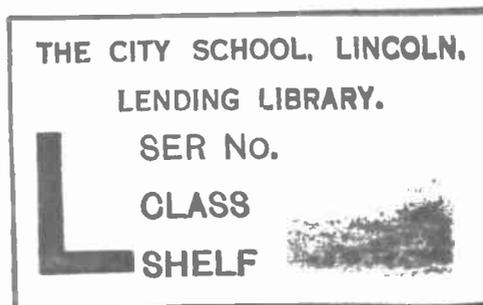
For 5XX, the last condensers are adjusted 5° each side of the tuning point, corresponding to 6,000 cycles.

Gain in brilliance and volume.
There is definitely a gain in brilliance of the music when this method of operation is used. Curiously, there is also a definite gain in volume, than when the circuits are sharply tuned to give the same radio frequency voltage at the detector grid.
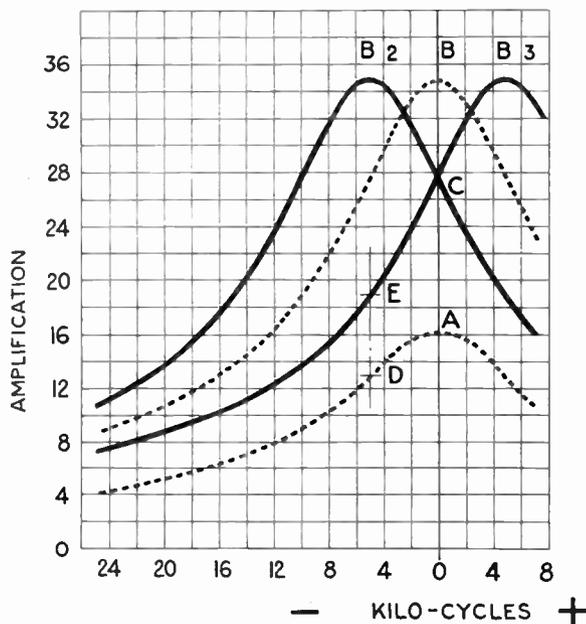
Fig. 20.

M = 50mH.   Wavelength = 357 metres.
R = 16 ohms.   ω =   5·29 × 10⁶.
g =   2.
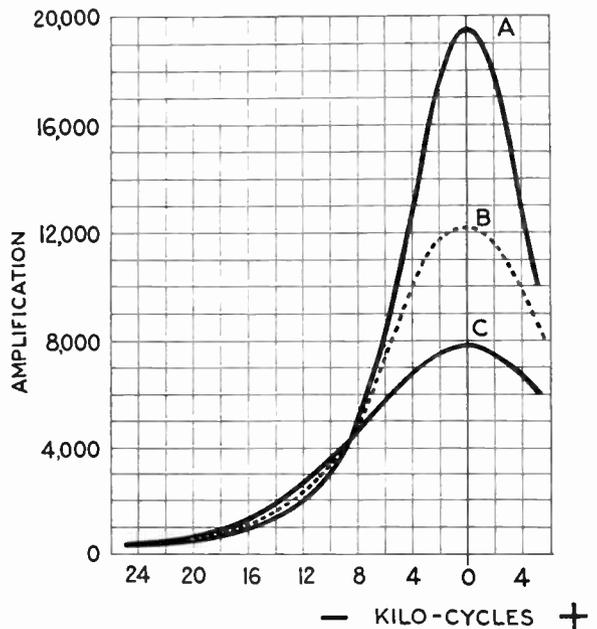
Fig. 21.

M = 50 mH.   Wavelength = 357 metres.
R = 16 ohms.   ω =   5·29 × 10⁶.

Figures 20 and 21 will enable the reader to visualise the process. In Figure 20, A is the amplification curve of the first circuit ; B is that of each of the other circuits. But the

second circuit is detuned 5,000 cycles lower than the carrier wave frequency ; therefore, B2 is the amplification curve.

The third circuit is tuned 5,000 cycles above the carrier frequency, B3 is the amplification curve. These curves are for coils of 16 ohms resistance each. To get the total amplification due to the three circuits, we multiply the ordinates of the three curves. For example, the total of the carrier frequency will be $0A \times 0C \times 0C$. For 5,000 cycles off this, $5D \times 5E \times 5B_2$, and so on. Thus we obtain the
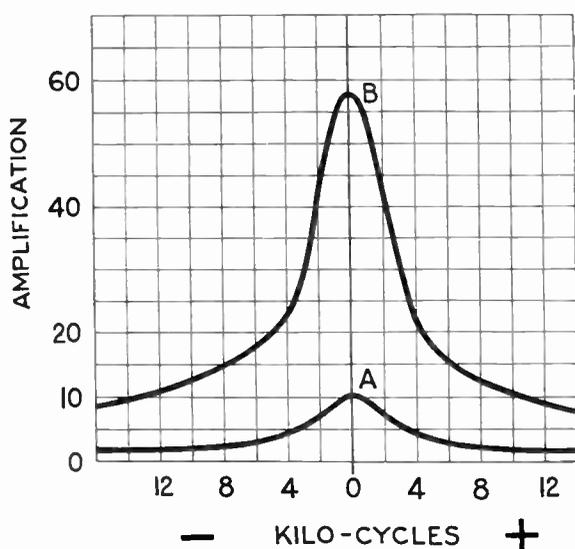
Fig. 22.

M = 500 mH.   L = 3,000 mH.   C = ·0002317 mfd.
R = 75 ohms. L₁ =   600 mH.   = ·00025 mfd.
g =   2.
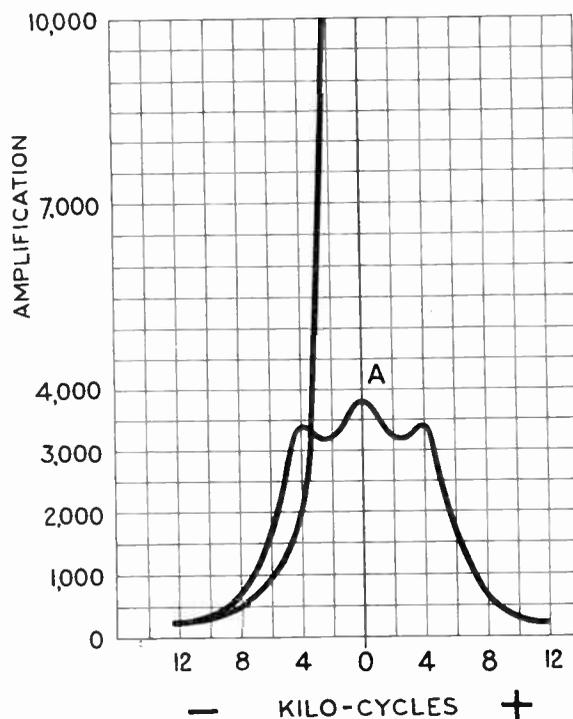
Fig. 23.

M = 500 mH.   Wavelength 1,570 metres.
R = 75 ohms.

curves in Fig. 21. A is the curve when all the circuits are tuned to the carrier frequency, B is the curve when the last two are tuned 5,000 cycles above and below, C when they are tuned 8,000 cycles above and below. The reader will notice the amplifications for curves B and C are less than for A. Also, that the 5,000 cycles note is amplified a greater proportion of the carrier frequency in B and C than in A.

Figures 22 and 23 are similar curves for 75 ohm circuits tuned for 1,570 metres. In the first, A is the aerial amplification curve and B that of each of the other two circuits. In Fig. 23, A gives the total amplification when the second and third circuits are tuned 5,000 cycles above and below the carrier respectively. The amplification is only about one-ninth of that when all the circuits are sharply tuned, but the 5,000 cycles note is almost gone in the latter case, whereas in the former it is at its full value.

Curve A, it will be noted, approaches out ideal compromise specified at the beginning of this section.

Correction by Band Pass Filters.

One can obtain, under suitable circumstances, something approaching our ideal compromise by utilising the properties of coupled circuits. If two circuits are near each other, or a portion of one brought near a portion of the other, then the flow of an alternating current in A, because of the magnetic field created in space, linking up with some portion of B, will transfer energy into the latter, causing a similar current to flow from it. This fact is the basis of all transmitters and receivers. But the transfer of energy into B from A modified the properties of each circuit; the current in A transfers, as it were, some portion of its inductance and resistance into B, and the current caused to flow in B transfers some portion of its inductance, etc., back into A. The effect naturally depends on the nearness of the coils to each other or the degree of coupling.
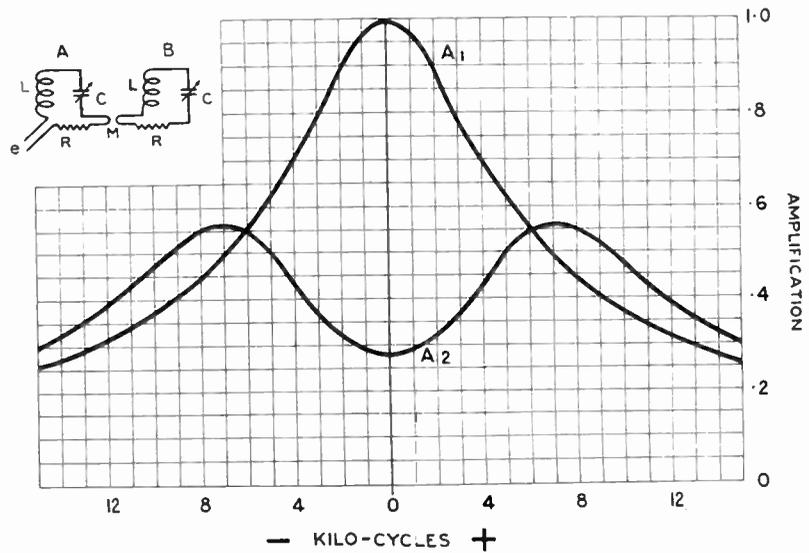
Fig. 24.

M = 3·02 mH and   R = 10 ohms.   W = 5·29 × 10⁶.
       ·1335 mH.   L = 200 mH.   C = ·00018 mfd.
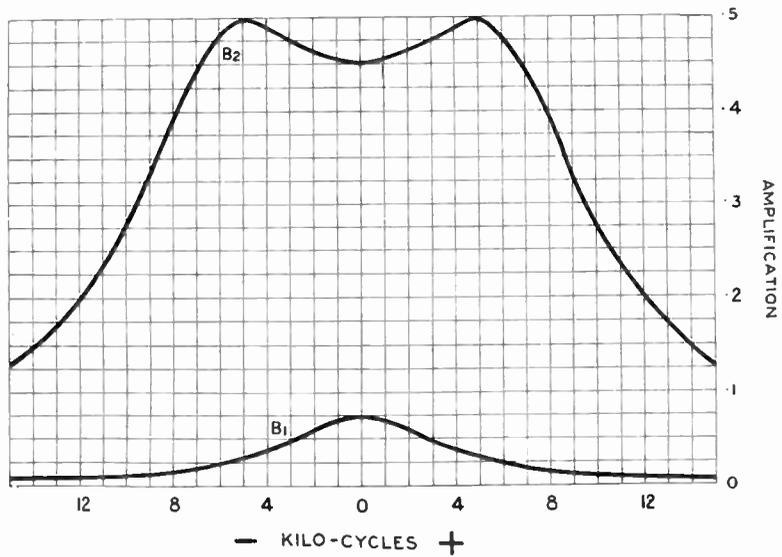
Fig. 25.

This is illustrated by curves in Figures 24 and 25. First, the coils A and B are far apart, so that A is hardly affected by the current induced in B. Curves A1 and B1 show the currents in the two circuits when 10 volts is produced in A. Notice that A1 and B1 are similar to standard resonance curves; in fact, A1 departs by only $\frac{1}{2}\%$ from this, because the mutual inductance between A and B (the measure of the amount of the coupling) is only ·133 microhenries.

Curves A2 and B2 are the current curves when the mutual inductance is increased to 3·02 microhenries. Notice that they are of quite a different shape; there are two frequencies for which the current is a maximum, and these are both different to the resonance frequency of each circuit if the other had not been coupled to it. The position and the height of the humps depend on the coupling, resistance, etc., of each circuit.
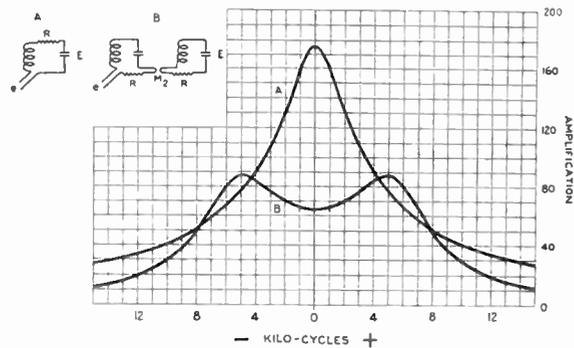
Fig. 26.

Figure 26 shows the result in another form. (A) is a normal resonance circuit, resonant frequency = 842,000 cycles. Curve A gives the ratio of voltage E across the tuning condenser to e, the voltage injected into the circuit. It is the

standard resonance curve. (B) are two coupled circuits and $\frac{E}{e}$, where E is the voltage across the second tuning condenser, is shown by curve (B). Such a circuit as (B) is the simplified diagram of what is known as a " band pass filter."
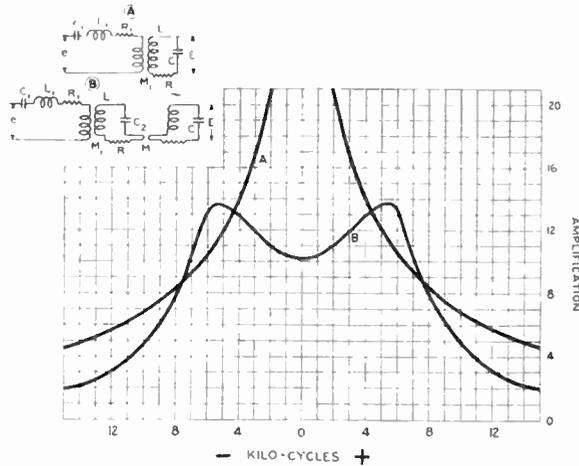
Band Pass Filter.

Fig. 27.

For use in an aerial circuit, it is arranged as shown in (B) of Figure 27, where C^1, L^1, and R^1 represent a standard aerial connected to a coil coupled to the " Band Pass Filter " by mutual inductance M_1, in this case 14·4 microhenries. (A) shows the same standard aerial connected to a simple resonance circuit.

Curves (A) and (B) are the ratios of $\frac{E}{e}$ obtained when e is applied across the standard aerial to earth.

You will notice the hump in (B), and that the 5,000 cycles to 6,000 cycles notes are amplified rather more than the low notes. When using the simple resonance (A) we obtain, it is

true, about double the maximum amplification, but the 5,000 cycles note is only amplified one half of its proper value. Further, for frequencies 15,000 cycles and above, off resonance, the amplification in (B) is relatively less than in (A), hence there is some gain in selectivity using the band pass filter, apart from the gain in the higher audio notes. In obtaining these results, we have taken the radio frequency resistance of the coils to be 6 ohms at 357 metres. This is an extremely low value, and could only be obtained by great care.
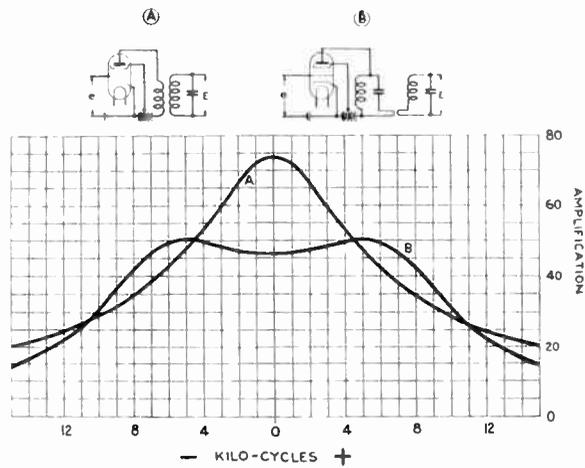
Fig. 28.

Band Pass Filter between two valves.
A band pass filter can also be used in the plate circuit of a screened grid valve. Figure 28 shows (A) a simple H.F. transformer and (B) a band pass filter in the plate circuit of a screened grid valve and the curves give the ratio of the voltage E across the grid of a second valve to that e applied to the grid of the screened grid valve. Resistance of coils = 10 ohms.

You will notice that the results are much the same as with the aerial band pass filter.
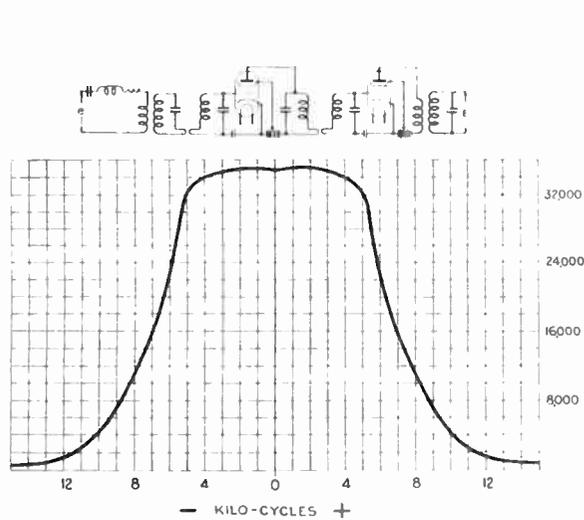
Fig. 29.
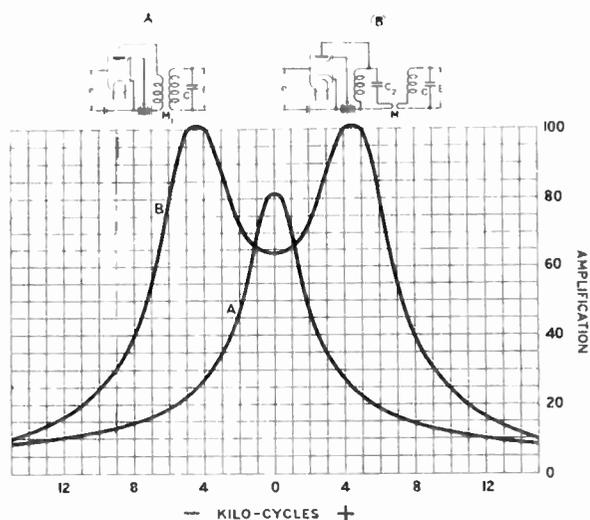
Fig. 30.

We will now consider a complete receiver made up of the above described aerial band pass filter, the valve plate band pass filter, and between the second valve and the detector valve, the simple H.F. transformer. Figure 29 shows this, and the curve for $\frac{E}{e}$ the total amplification. We have clearly, in such a combination, obtained something very closely approaching our ideal compromise.

Figures 30, 31, and 32 show a similar set of curves for the long wave band, the wave length of the carrier wave being 1,570 metres. In this case, the resistance of the coils and the degree of coupling have not been so happily chosen, since for the complete receiver, the 6,000 cycles note is amplified more than is desirable relative to the lower notes. A slight change in the values would give a curve more approaching the ideal.

At first sight it seems that we have, in the above arrangement, an ideal solution of our troubles, and can get selectivity
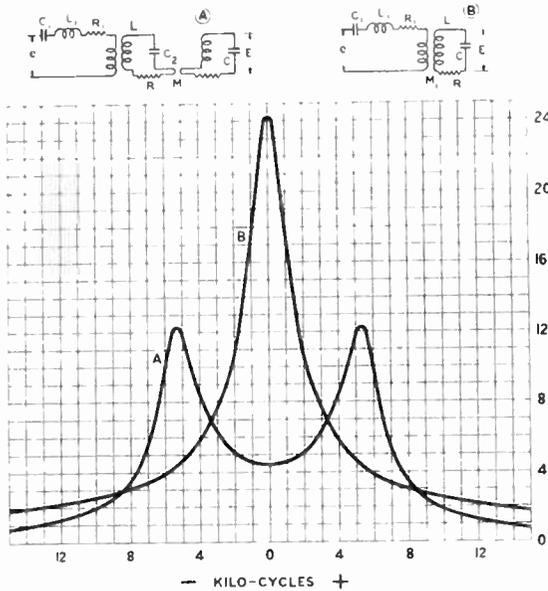
Fig. 31.
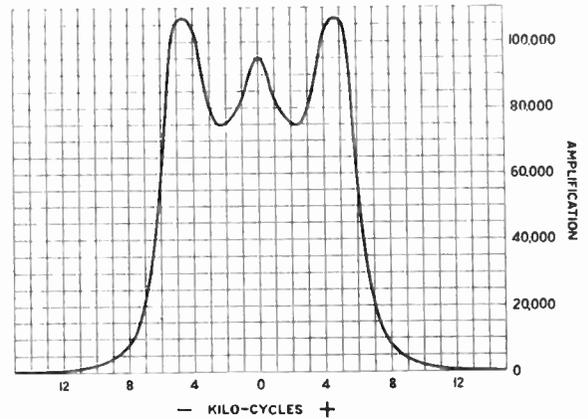
Fig. 32.

without great loss of high notes. But there are difficulties. First, due to the necessity of the fixed coupling of the band pass filter coils, and the varying resistance of the coils with frequency, we find that the curve which seems so ideal at 357 metres, becomes too broad at 200 metres with loss of selectivity and too narrow at 500 metres with some loss of quality. The advantage of the arrangement, despite this, is such that its use would soon become general but for a greater difficulty, which is mechanical. There are five condensers, which must be ganged much more accurately together all round the scale. The coils must be accurate to within $\frac{1}{4}\%$. This is not difficult. The condensers moving round must be accurate to within $\frac{1}{6}\%$ all round the scale until near the minimum, where the error might be $\frac{1}{3}\%$, which is not easy to obtain with the uncertain capacities of wiring and valves.
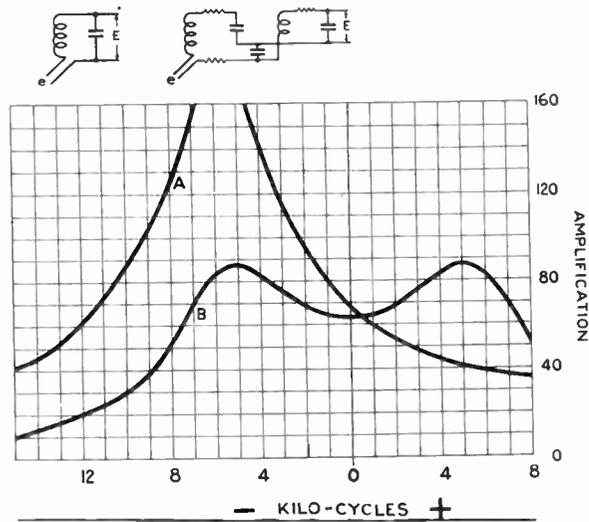
Fig. 33.

Capacitative Coupling in Band Pass Filters.

So far we have only considered the case where the coupling between the two tuned circuits was through space, or inductive. The two circuits can be coupled by means of a condenser, ·0135 mf, as shown in Figure 33. The two curves give the ratio $\frac{E}{e}$ where e is the voltage injected into the first circuit, for (A) a single tuned circuit and (B) for the coupled circuits. The latter shows the double hump seen in the previous figures. But there is a difference. In the previous figures the humps came equally spaced each side of the tuning of the single circuit, or the frequency of the received signal. Here, however, with capacitative coupling, one hump coincides with the tuning of the single circuit. This means that if the receiver contains single tuned circuits as well as band pass filters, as previously shown, different capacities will be wanted to tune the latter to the incoming signal than in the case of the former. This introduces further difficulties to the ganging of the condensers.

Band Pass Filter in the case of the Super-Heterodyne Receiver.

In such a receiver, the incoming signal is reduced to a lower or intermediate frequency for amplification by a local oscillator. This intermediate frequency is always constant, so that the design of the band pass filter in this amplifier is easy, since there is no need to change the tuning of the circuits. There is the further advantage that the intermediate frequency is low—in the case we have in mind, 125,000 cycles or 2,400 metres wave length—hence the ratio of inductance to resistance of the coils can be very large, leading to great selectivity.
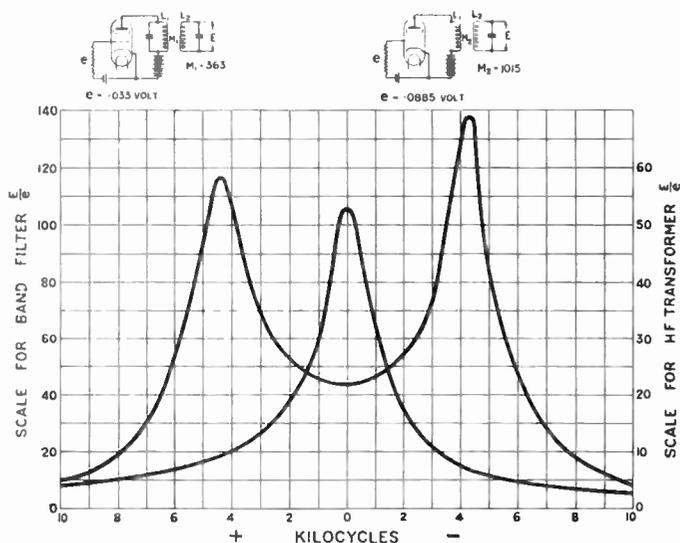
Fig. 34.

Wavelength 2,270 metres.   S4V Valve.

Anode Volts = 162.	G.B. Volts = $-1\frac{1}{2}$.	L_1 = 4,610 μ.
Screen Volts = 70.	g ... = ·95.	L_2 = 4,730 μ.

Figure 34 shows diagrams and curves of a first attempt to design an intermediate frequency stage for a super-heterodyne. Two stages were proposed, one band pass coupled and the other plain H.F. transformer coupled. In the Figure,

circuits and full particulars are given. The valves had a mutual conductance of ·95 milliamps. per volt. The curves give the *measured* values of $\frac{E}{e}$ in each case. Due to the fact that one coil of the band pass filter was of slightly larger inductance than the other, the humps are not exactly of the same height. From these we can estimate the combined amplification of the two in series.

For the tuned frequency (2,270 metres) = 44 × 53 = 2,330.
For 4,700 cycles ... ... = 137 × 6·3 = 861.
For 10,000 cycles ... ... = 2·7 × 8 = 21·5.

Hence, we have reduced the station 10,000 cycles away 108 times, a wonderful selectivity, and yet only reduced the 4,700 cycles note 2·7 times, giving quite a good quality. If one could use two stages of the band pass filter and one of the H.F. transformer (there would be two difficulties, owing to the great amplification; firstly, self oscillation, and secondly the great noise level) at tuning, amplification would be 44 × 44 × 53 = 102,500.

At 4,700 cycles away 137 × 137 × 6·3 = 118,000.
At 10,000 cycles away 2·7 × 8 × 8 = 172.

We would have reduced the station 10,000 cycles away 600 times, a marvellous selectivity, and the 4,700 note would be slightly above its proper value.

Though this case, with this degree of amplification, is hardly practical, yet it will enable the reader to get a glimpse of the possibilities of such a system.

Minor difficulties of Radio Frequency Amplification. The first difficulty experienced is the unequal amplification over each frequency band (viz.: 200 to 330 metres and then 1,000 to 2,000 metres). In receivers, the tuning inductance for each band is fixed, and the variation over the band is obtained by changing the capacity. For the longer wave lengths (say 550 metres) the capacity is large, and for the shorter (say, down to 200 metres) is small.

The amplification of each tuned circuit is approximately proportional to $\dfrac{M}{RC}$ where M is a fixed mutual inductance, C the tuning capacity, and R the inductance of the circuit at the tuning frequency. Though R increases with the reduction of wave length, the capacity C decreases much faster, hence the amplification for the short wave lengths of one band is greater than for the longer. An additional cause of the above effect is that unintentional reaction has a greater effect at the short wave lengths. This is only another way of saying that the effective R is less at these wavelengths than is warranted by the copper and other losses.

Figs. 39 and 40, page 77, giving the sensitivity curves of the A.C. Mains Receiver, Model 32, shows the variations obtained in a well designed set. If one tries to increase the sensitivity at 500 metres, the set will amplify so much at 200 metres that it will oscillate. The trouble is greatly increased in sets using more than one stage of H.F. amplification.

Remedies.

The best remedy is to vary the mutual conductance g of the screened grid valve or valves, which is a factor of the amplification. This can be done in two ways.

1st: If the grid bias is constant and obtained from an independent source, vary the screen voltage. Figure 35 shows the relation between the mutual conductance of the S4V valve with the screen voltage for a constant grid bias of $-\frac{3}{4}$ volts. The variation is usually carried out by connecting a 50,000 or 100,000 ohm potentiometer across the voltage supply, the screen being connected to the movable point. In commercial sets, where the controls must be few, this potentiometer can be driven from the tuning condenser, so that the voltage is approximately correct for each position. Thus, one can obtain an approximately flat sensitivity curve.
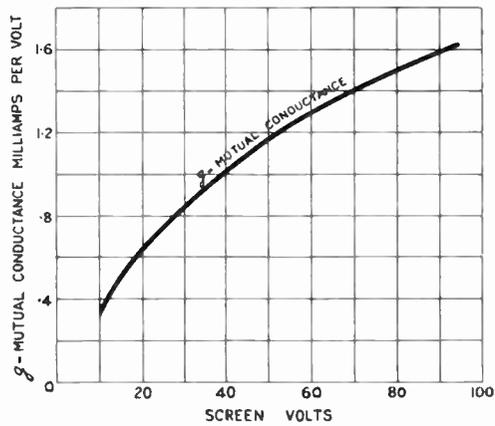
Fig. 35.
S4V Valve.   Plate Volts 153.   Grid Bias – ¾ Volts.

2nd : Most modern sets obtain the bias by the fall of potential caused by the plate current passing through a resistance (see Figure 11). In this case, the above method is of no use, since the bias falls as the screen voltage falls, thus giving approximately constant amplification. In this case, the best solution is to make the resistance causing the bias, variable, thus varying the bias.

The curve in Figure 36 shows, first, the plate current and the value of g corresponding to various grid biases for a valve S4VA. Between grid bias –1 and –2 the value of g varies between 2·43 and ·47, the plate currents from 1·74 milliamps. to ·32 milliamps., hence, resistance must vary from 6,250 ohms to 575 ohms to give the bias. One cannot use a bias less than –1 volt, because grid current begins and introduces damping into the tuned circuit ; it is not advisable to go below –2 volts, as there may be some slight rectification due to curvature of plate current curve. Here again, one can change the resistance automatically by the moving of the tuning condenser.
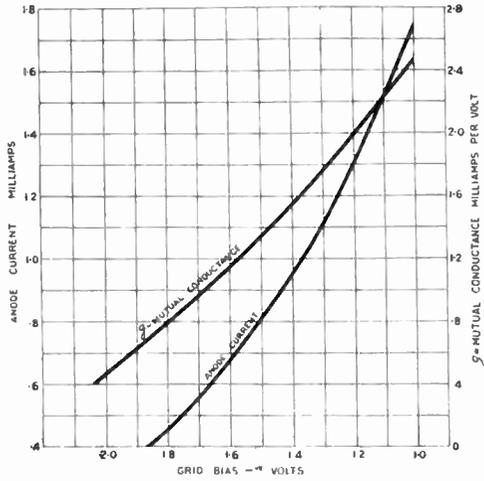
Fig. 36.

S4VA Valve. Plate Volts 153. Screen Volts 73.
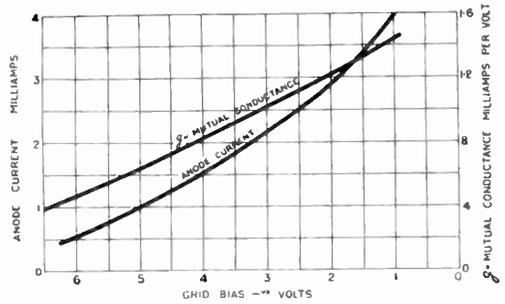
Fig. 37.

MS4 Valve. Plate Volts 152. Screen Volts 77.
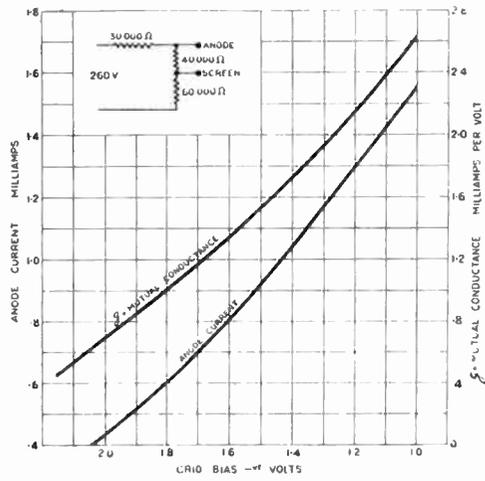
Fig. 38.

S4 VA Valve.

In many American sets, the H.F. transformers have primaries of large inductance, part of which is bye-passed by a fixed condenser. By suitable adjustment of the two, one can get a nearly flat sensitivity curve.

In other cases, an untuned H.F. transformer stage which self tunes at 550 metres is used, thus giving a descending amplification curve for this stage, correcting for the reverse effect in the other stages.

Effect of aerial on tuning of the First Circuit. In receivers where the first circuit is tuned by a separate condenser from that used for the other circuit or circuits, this effect is not important—one can always suitably adjust the condenser.

Modern convenience, however, demands that the condensers be " ganged," and the tuning of all, carried out by moving one control knob. Now, the aerial effect is serious. Due to the coupling, a portion of the aerial capacity is transferred to the tuning circuit, and reduces the tuning capacity wanted, this effect varying with different wavelengths.

If all aerials were alike in capacity, one could fit an automatic correcting device which would enable the condensers to be ganged with accuracy. But the capacity of aerials varies enormously, making ganging difficult.

Remedies. 1st : In such sets as the A.C. Mains Receiver, Model 32, the condensers are made accurate and ganged ; but the " stator " (the fixed blades) can be moved a small amount independently of the ganged " rotors " (moving blades) by a separate control to correct the first condenser for aerial effect. This arrangement, in some form or other, is now coming into general use for sets, but can be troublesome unless suitable precautions be taken, since, except in the case of weak signals, there is more than one combination of the setting

of the two controls, which apparently gives the maximum signal; it can, therefore, never be certain when the receiver is accurately tuned to the signal. This is met in the case of the above set, by making the wave length scale accurate to the tuning of the second circuit, so that the scale can be set to the desired wave length and the corrector control used to sharpen the tuning of the first circuit. This arrangement, although not ideal, works well in practice.

2nd : Loosening the coupling of the aerial to the first tuned circuit (Fig. 13) by using less turns in the aerial coil, or by a suitable series condenser in the aerial circuit, will finally render the aerial effect small enough to enable the condensers to be ganged accurately. But, unfortunately, this reduces the amplification of the set. When two stages of H.F. amplification are used, this loss can be compensated, and one obtains an efficient one knob tuning receiver.

With a single stage of H.F., the result can only be a compromise; to obtain good ganging one must lose amplification, though some of the loss can be made up by increasing the ratio of the H.F. transformer, or by adjusting the screen potential or grid bias as just described.

Valve capacities.
These very appreciably affect ganging, since they are large and are a part of the tuning capacity. Therefore, the minimum values of each condenser must be adjusted by very small fixed condensers to be equal. The types of valves used cannot be changed, and should the valve makers change the properties of the valves appreciably, the ganging may be upset.

Figures and Curves obtained from the A.C. Mains Receiver, Model 32.
This type of set, with one stage of H.F. amplification, detector, and power valve is, at the moment, very popular, and the information previously given, naturally, has been utilised in its design.

Figure 11, page 29, shows the complete circuit diagram of this set, with values for all the components.

Figure 11A, page 30, shows the structure. Except for the method of screening the second H.F. circuit, the layout does not appreciably affect the electrical performance ; other constructions would give the same result.
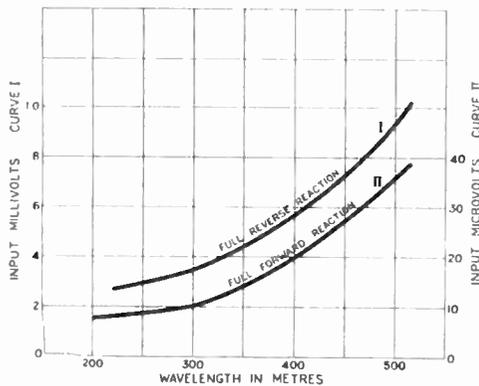
Fig. 39.

Sensitivity Medium Wave Band.
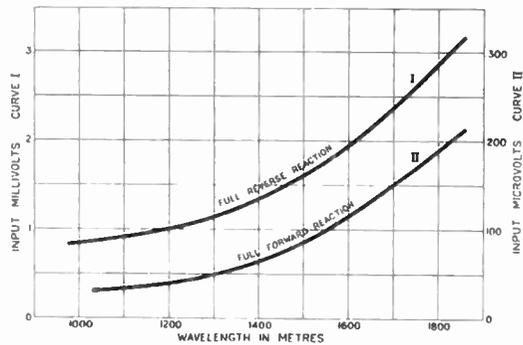
Fig. 40.

Sensitivity Long Wave Band.

A.C. Mains 32. Input to Standard Aerial Modulation 30% to give 50 Milliwatts Output.

Figures 39 and 40 show the " sensitivity curves " or the input, in millivolts or microvolts as the case may be, to a standard aerial connected to the set, to give standard output, viz.: 50 milliwatts, or, in this case, 15·7 audio frequency volts, across a resistance of 4,900 ohms, the radio frequency being modulated 30% with a 400 cycles note. Figure 39 is for the medium wave band, and Figure 40 for the long wave band. On each, curve I is the input in millivolts, when full reverse reaction is used, the condition for the least amplification and the best quality. Curve II is when the set is used with full forward reaction. This is an indefinite state for the set ; in fact, no doubt, more reaction could be used, but measurements would be difficult because of the fluctuations of the output instrument caused by variation in the A.C. mains voltages. On days when the mains are very steady, smaller inputs than those shown will give standard output.
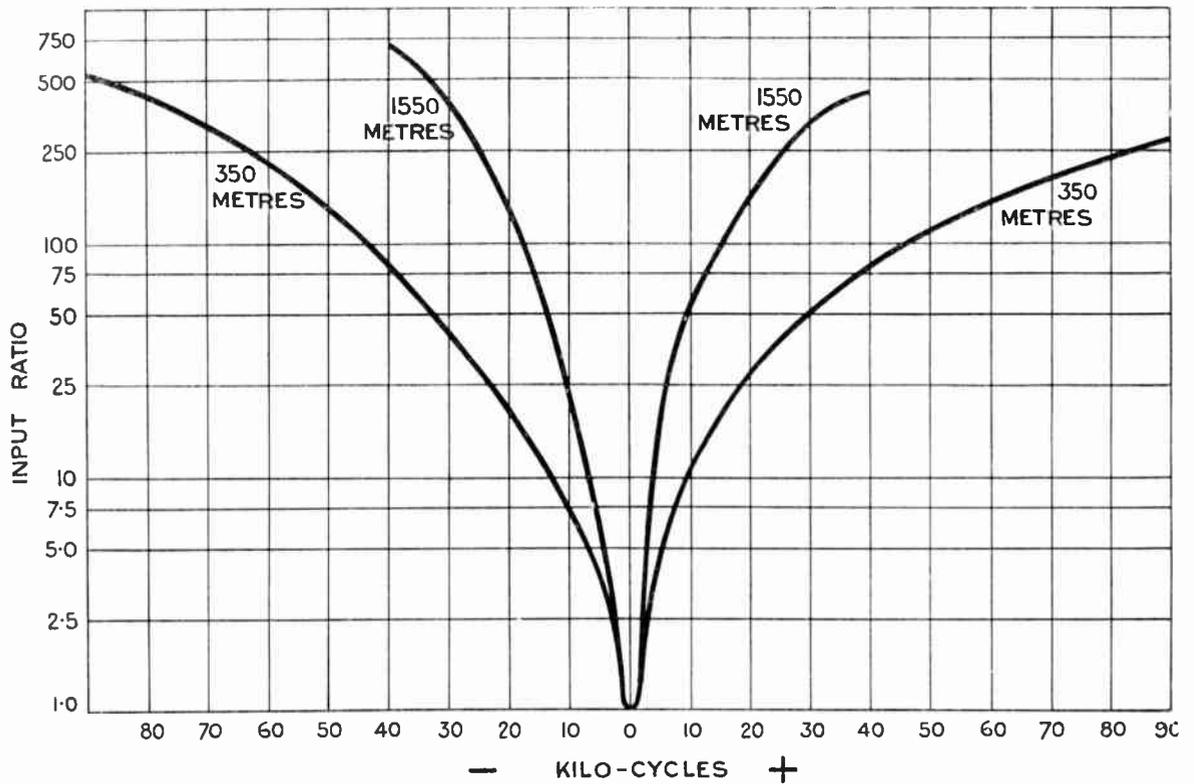
Figure 41 shows the *selectivity curves* for 350 metres and 1,550 metres when considerable, not the maximum, forward reaction is used.

Fig. 42.
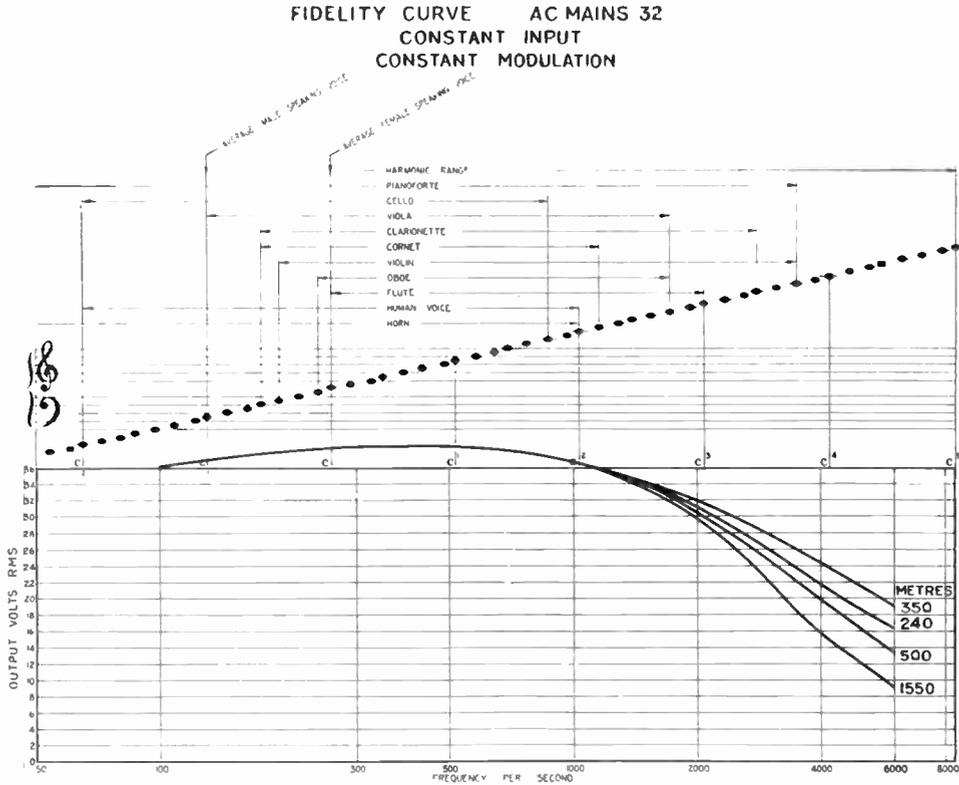
Figure 42 shows the fidelity curves for 240 metres, 350 metres, 500 metres, and 1,550 metres, using full reverse reaction to get the best quality. A constant input to the standard aerial is modulated 30% with notes from 100 cycles to 6,000 cycles and the output volts measured.
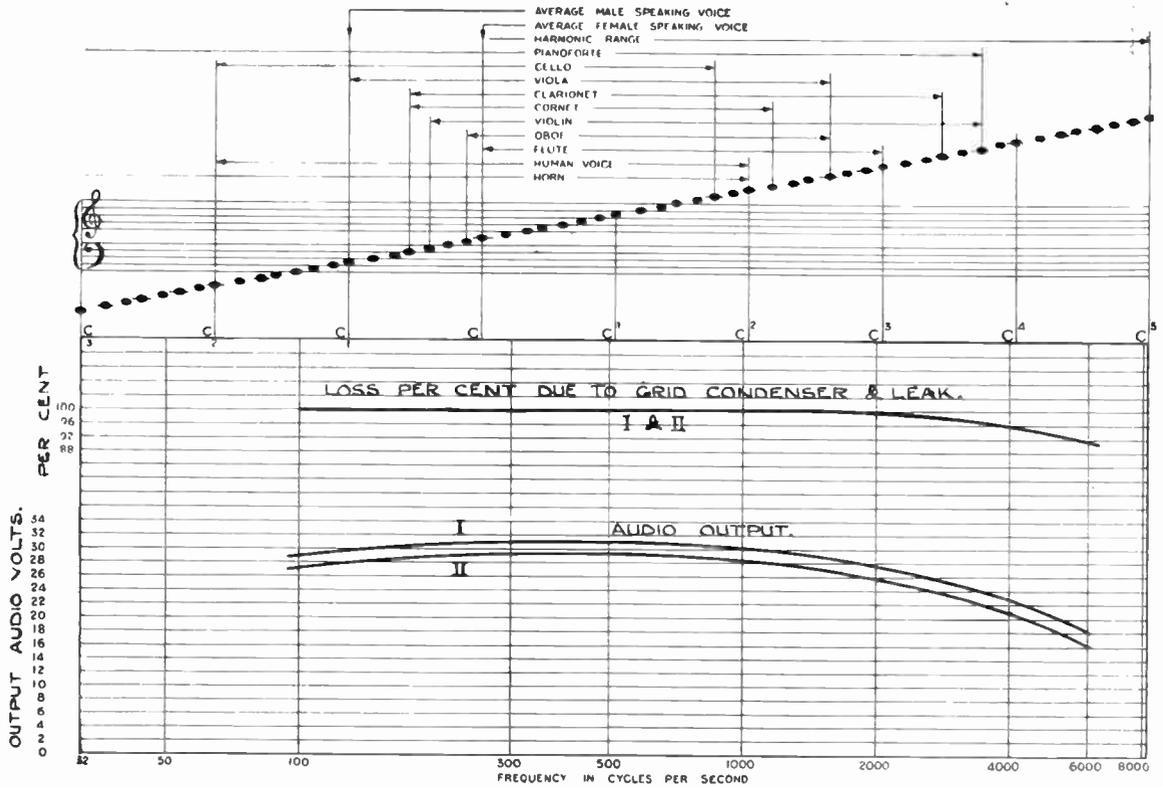
Fig. 43.

Figure 43 shows the fidelity curves when a constant input is applied to the grid of the detector valve, thus avoiding the losses in the H.F. stages. There should be only one curve for all wave lengths; two are shown, one for 250 metres and other for 1,500 metres, because in the measurements different percentages of modulation were used. Allowing for this, they become one curve. The droop at the high audio frequency end is due to (a) losses in the audio frequency and output stage, (b) the reduction of impedance at the high frequencies of the grid leak condenser

combination. Measuring the audio frequency and output loss directly, one can obtain the curve giving the loss (b) as a percentage of the 100 cycles note.

From all these curves, one can analyse what amount of high note loss occurs at each stage. We will compare the 5,000 cycles note with the 500 cycles note.

WAVE LENGTH 350 METRES.

Overall fidelity. The 5,000 cycle note is $\frac{21}{39}$ of
the 500 cycles note ... ...   53·8%
Due to grid leak and condenser it is 94%
Due to the audio stages 69%
Due to grid leak and audio stages, ·94 $\times$ ·69 $=$ 65%
Due to selectivity of H.F. circuits it is ...   82·7%
    TOTAL ... ... ·94 $\times$ ·69 $\times$ 82·7 $=$ 53·8%

WAVE LENGTH, 1,550 METRES.

Overall fidelity 29%
Due to grid leak and condenser 94%
Due to audio stages 69%
Due to H.F. circuits ... ...   44·6%
    TOTAL ... ... ·94 $\times$ ·69 $\times$ 44·6 $=$ 29%

From these figures there is a greater loss of high notes due to the H.F. Stage in the case of 1,550 metres, than 350 metres. The same is very evident in the selectivity curves. The reason is that 1,550 metres is equivalent to 193,500 cycles, and 350 metres 857,000 cycles, and

F

5,000 cycles is a much larger percentage of 193,500 than it is of 857,000, so, for the 5,000 cycle note, we are much further off resonance in the 1,550 metre case, than in the 350 metre.

RADIO FREQUENCY AMPLIFICATION OF SET.

It is found that, with these sets, a standard output of 50 milliwatts is obtained from an input of 30% modulated radio frequency to the grid of the detector valve, which is constant over the wave lengths 200 metres to 2,000 metres ; such variations which result are less than normal experimental error. For the set used to obtain sensitivity curves of Figure 39, this input is ·2 volts. We can now obtain the radio frequency amplification.

Thus at 350 metres, a full reverse reaction of ·44 millivolt gives 50 milliwatts output. Hence the R.F. amplification $= \dfrac{0 \cdot 200}{0 \cdot 00044} = 455$. Using full forward reaction, the input is 14 microvolts to give 50 milliwatts output. Hence the R.F. amplification $= \dfrac{200,000}{14} = 14,300$, which is $31\frac{1}{2}$ times the first figure. Similar figures can be obtained for other wave lengths.

Value of R.F. voltage which can be used at grid of detector valve.

There remains the question of what radio frequency signal will be used at the grid of the detector valve in practice. To find this, we start with two assumptions. First, that the B.B.C allow only a maximum modulation of 80% on the peak passages. This is their published statement. In this connection, we find in the *Bell Journal*, October, 1925, the statement by Sacia that measurements show that " for English speech, the maximum amplitude to mean, is about 5/1." Hence, a 16% mean modulation would give a maximum of 80%.

Secondly: That the user will adjust, so that his power valve commences to overload when the modulation is 80% —that is, on the peak passages.

In the A.C. Model 32 Receiver, the power valve P625, with 230 volts on the plate, will begin to overload when an 18 volt R.M.S. signal is applied to its grid. The amplification from the grid of the detector to the grid of the power valve is 120, so that the audio signal at the grid of the detector is $\frac{18}{120} = \cdot15$ volt. For this, the efficiency of rectification is 65%. Hence, radio frequency volts $\times \cdot8 \times \cdot65 = \cdot15$.

The radio frequency signal is, therefore, $\cdot29$ volt.

Referring to the section on rectification, the curves and figures given there show that there is no appreciable distortion with such a signal except, that on the peak transient passages of 80% modulation, some small percentage harmonic may be introduced.

WAVETRAPS OR REJECTORS.

On page 45 we reviewed the difficulty due to very strong incoming signals on the aerial, and showed that the only thing to be done, was to reduce the amplification of the receiver most conveniently by a series aerial condenser, to all signals, and then to increase it by reaction for the wanted signal to which it was tuned. There is a limit to the efficacy of this procedure, but in any case, the use of much reaction spoils the music by high note loss.

There is another method of reducing the interference of strong stations which does not necessitate the use of so much reaction, by means of what is known as a wave trap.

It consists of a circuit of some type tuned accurately to the unwanted station and coupled to the aerial circuit to impose in this a high impedance to the unwanted signal, and a smaller impedance to the wanted one (there is another type of lesser importance). We must warn the reader that though a well-made wave trap can be very effective, the best will not do the miraculous things the public have been led to expect ; many of those offered are, in addition, highly inefficient and bad.
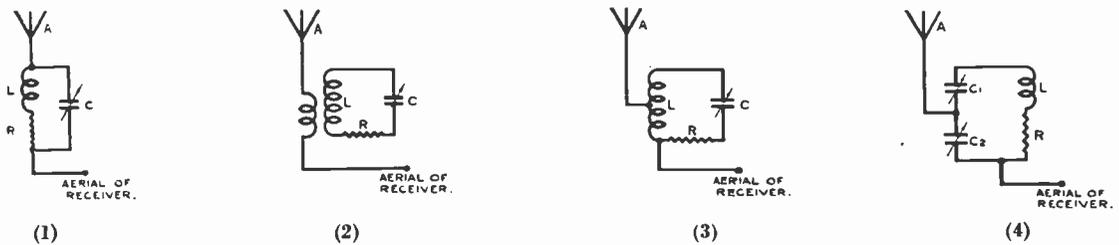
Fig. 44.

In Figure 44 (1), (2), (3), and (4) a tuned circuit (L, R, and C) is tuned to the unwanted station. The aerial is connected to some point on the circuit, and from some some other point a connection is brought to the aerial connection of the set. We have found, both by calculation and by measurement, that the simplest direct form (1) is *not less* effective than the others, and so we shall deal with it alone.

The impedance of the circuit to the station to which it is tuned is $\dfrac{L}{CR}$. If we made L large, and, therefore, C small, and R as small as possible, we should get a very great reduction of the unwanted station, 500 times or more ; but, unfortunately, the reduction extends to a lesser degree to other wave lengths that we wish to receive, and would make it difficult to receive them. Hence, we must be satisfied with a lesser reduction of the unwanted signal.

We have chosen for the 250 to 360 metres band a coil of litz wire wound on a ribbed former with an inductance of 53 microhenries. This, and its tuning condenser, have the very low resistance of $1\frac{1}{2}$ ohms at 350 metres, so that any results with it are likely to be the best that can be obtained.
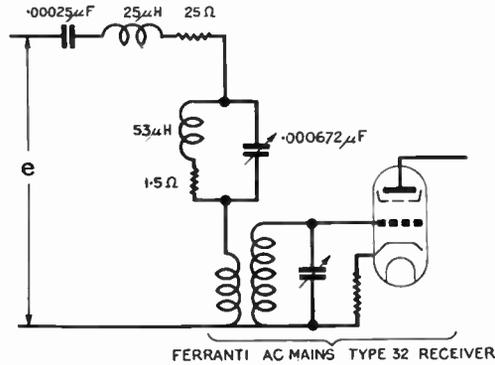
Fig. 45.

Figure 45 shows it connected into a standard aerial, and the first tuned circuit of A.C. Mains Receiver, Model 31 or 32.

TRAP TUNED TO 356 METRES.

Tuning wave length of Receiver.	Resistance introduced. Ohms.	Inductance introduced. Microhenries.
356 metres	52600	− 53
356 metres ± 10,000 cycles	2520	2120
356 metres ± 20,000 cycles	638	− 1082
396 metres	41·5	+ 279
508 metres	5·77	+ 104
324 metres	34	− 252
274 metres	3·15	− 77

The exact effect of the trap depends on the aerial circuit of the set to which it is connected, and on the properties of the aerial. You cannot be sure that the trap will behave with one type of set as with another. The table on page 85 gives the amount of inductance (negative inductance is capacity) and resistance put into the aerial by the trap tuned to 356 metres whose inductance L is 53 microhenries and resistance $R = 1\frac{1}{2}$ ohms.

Referring to Figure 45, the aerial impedance of the standard aerial at 356 metres is 465 ohms (capacitative) with the trap in, the total aerial impedance is 52,600 ohms, hence the signal of 356 metres is reduced to a first approximation in the ratio $\frac{52600}{465} = 113$ (the more correct figure allowing for other factors is 108). But if the incoming signal is $\pm$ 20,000 cycles away, the aerial impedance is still 465 ohms without the trap and is 5,740 ohms with the trap in, due to the $- 1082$ microhenries introduced (see table). Hence, the reduction of this signal due to trap is $\frac{5740}{465} = 12\cdot3$ (the more correct figure is 12·9). Thus, contrary to what the public has been led to believe, the trap in this particular case reduces the wanted station to less than 1/12th of its proper value.

On the other hand, if the trap had not been used, the unwanted station would have had to be reduced 108 to 1 by other means (the series aerial condenser in A.C. 32), which would also have reduced the wanted station 108 : 1 ; hence, using the trap, the wanted station is 8 times as strong as it would have been otherwise. This is a very great gain, but not the miraculous result often ascribed to the wave trap.
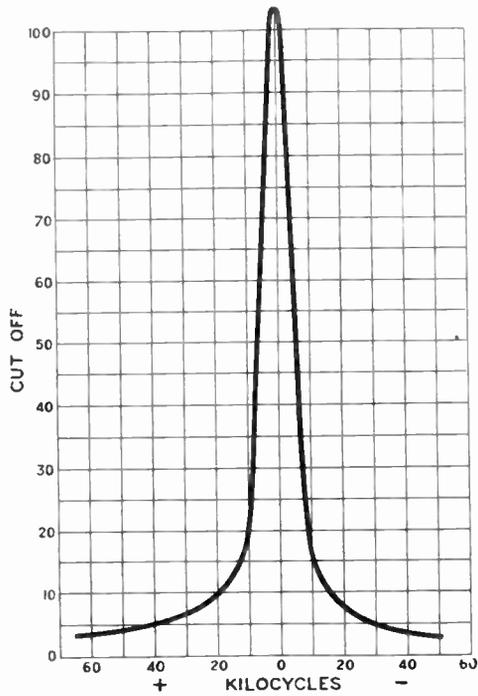
Fig. 46.

53 μh.
Coil, 27 turns of 27/42 Litz 840 Kc.
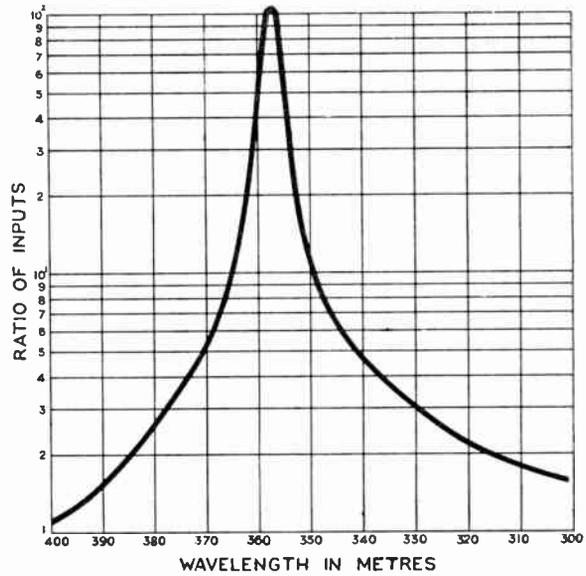
Fig. 47.

Wavetrap 53 μh. R = 1·5 ohms Tuned to 357 Metres in
Standard Aerial connected to A.C. Mains 32.

Figs. 46 and 47 show the measurements of the above trap in a standard aerial connected to the latest A.C. Mains Receiver, Model 32.

First, on a plain scale for ± 60 kilocycles each side of tuning, 356 metres.

Secondly, over a broader wave band, the vertical scale being logarithmic.
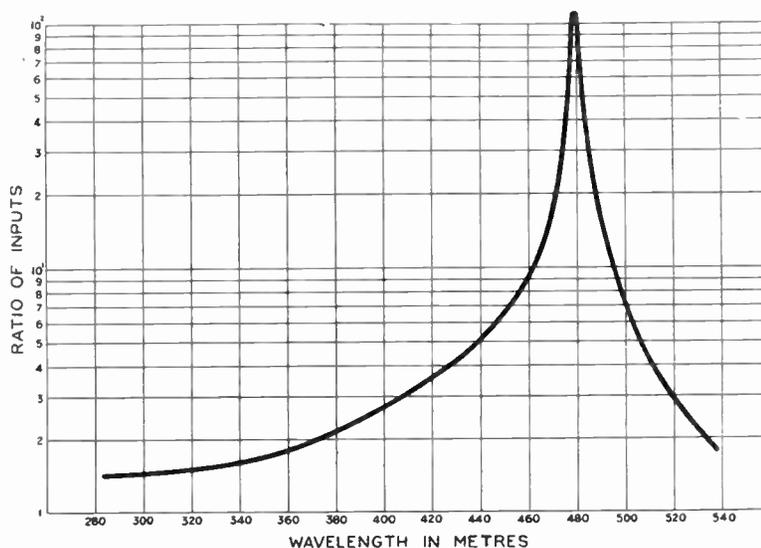
Fig. 48.
Wavetrap in A.C. 32 Mains.   For New North Regional, 479·2 Metres.
104·6 μh.   R = 3 ohms.   Measured Input for Constant Output.

Figure 48 shows similar measurements taken with a trap designed to cut out 479·2 metres, one of the wavelengths assigned to the North Regional Stations.   The coil here is made of ordinary wire on a ribbed former and has an inductance of 104·6 microhenries and a resistance at 479 metres of 3 ohms.

These curves will enable the reader to get an idea of the effect on wanted stations of various wavelengths, when the trap is tuned to an unwanted one.

The above deals with the effect of a single wave trap.   But listeners near high powered Regional Stations will receive exceedingly strong signals on two wavelengths from the station, and to be able to receive other stations, two wave traps in series may be necessary, one tuned to one unwanted station and the other to the second.   One must know, therefore, the combined effect of the two on reception.
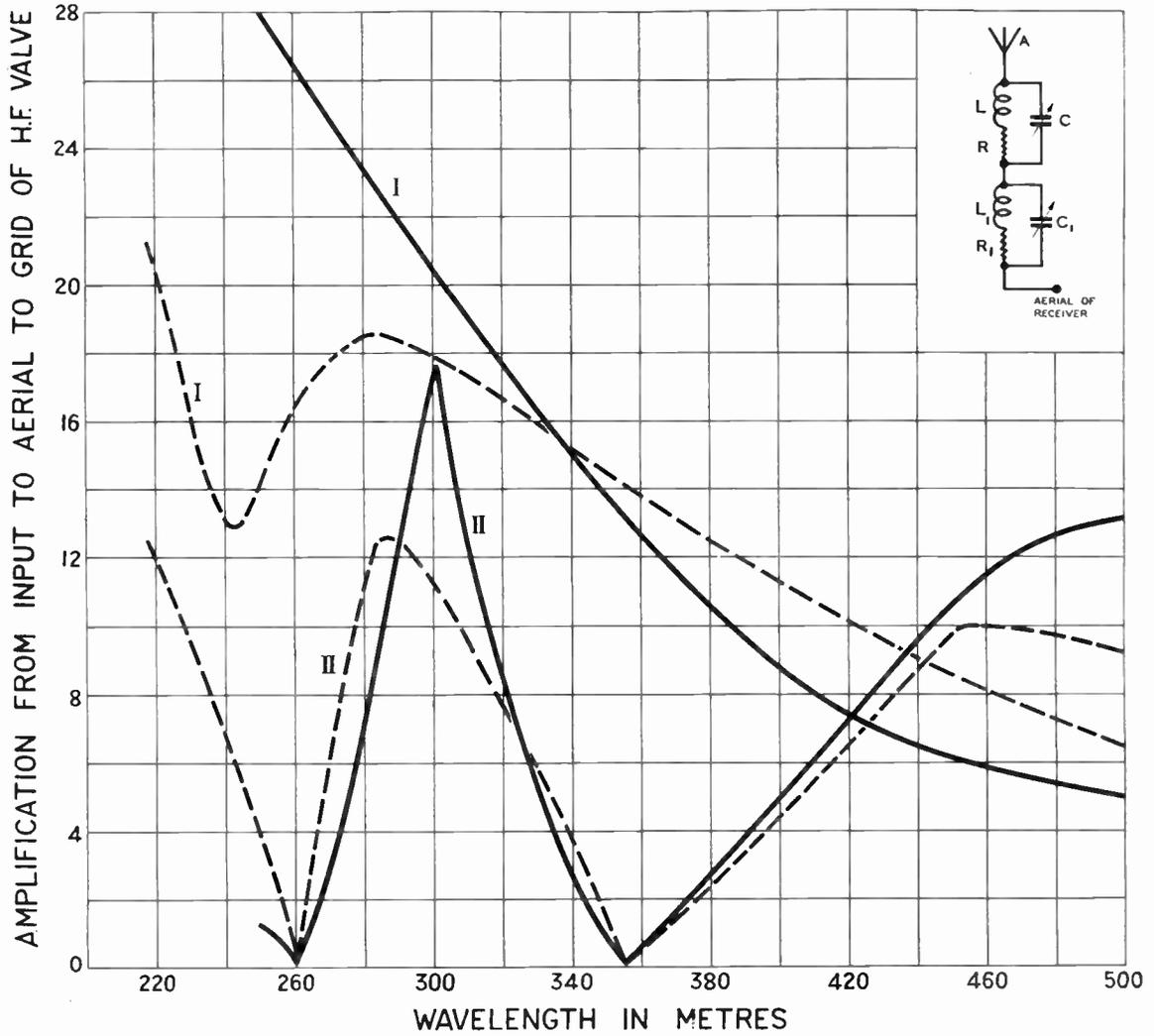
Fig. 49.

The two traps to be considered consisted of two litz wire coils each of 53 microhenries, and resistances at 356 metres of $1\frac{1}{2}$ ohms, and at 261 metres, 2 ohms, and were tuned respectively to these wavelengths. By a long series of laborious calculations one can calculate the effect, taking probable values of the resistance of the first tuned circuit at different frequencies in the set. Curves I and II (full lines), Figure 49, show the result of such calculations for the wave traps in series with a standard aerial and A.C. Mains Receiver, Model 31. Curve I is the amplification from the input to standard aerial, to the grid of the H.F. screened grid valve, for the receiver without the wave traps. Curve II is the equivalent for the receiver with the two traps in the aerial. The dotted curves give the results of measurements made with this set with traps in and out.

From these curves, one can see that there are quite large portions of the wave band close to where the traps are cutting out the unwanted stations, over which the wanted signal is greatly reduced. At the higher end of the wave band, the traps actually increase the amplification of the combination. This is due to the inductance thrown into the aerial by the traps, bringing the whole aerial circuit nearer to tuning to the higher wavelengths.

DETECTION, RECTIFICATION, OR DEMODULATION.

The Second Operation.

As explained on page 20, the radio frequency signal as received direct in the aerial circuit, or amplified by the means described previously, has to undergo the process of " detection " or " rectification " ; or, to give it its proper name, " demodulation," before it can be utilised to operate headphones or loud speakers. We will now explain the processes by which this can be done—that is, how the audio frequency components corresponding to the music in the studio can be separated so that they can be amplified to operate a loud speaker.

Any conductor which is " non-linear," to use a technical term—that is, when an alternating current is applied to it, lets more current through when the voltage is positive than when the voltage is negative, or vice versa—will act as a " rectifier." The nonlinearity in our case must be measured when the alternating current is of radio frequency.

Crystal Rectification.   Such a conductor is the usual crystal of silicon or galena, used in receivers of the simpler type. Figure 50 shows such a receiver circuit connected to headphones. Below the circuit diagram is the current, signal voltage curve. Its shape
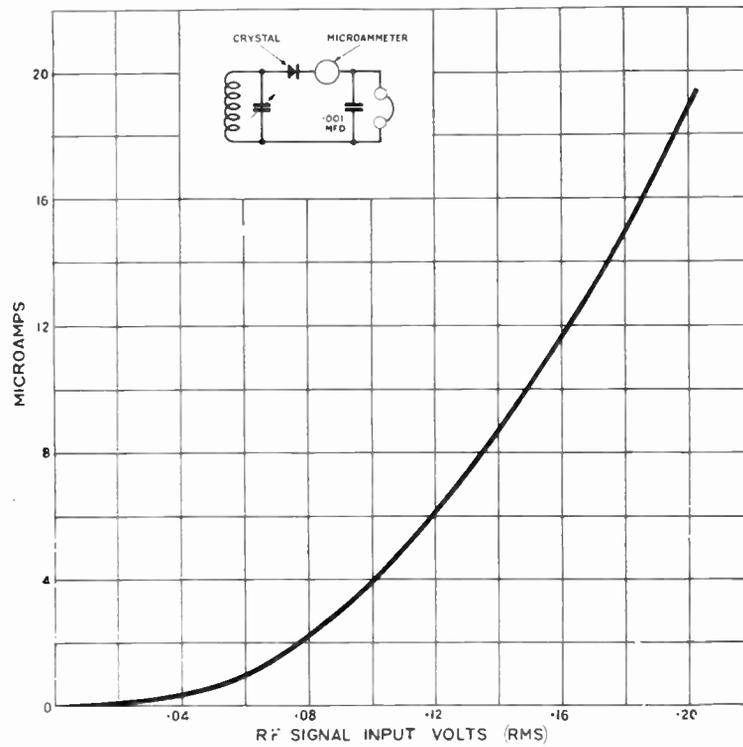
Fig. 50.

is almost identical with the plate current curve in Figure 52, illustrating anode bend rectification, and the explanation given there applies to the crystal if one reads instead of plate current, signal current, and for grid volts, signal volts.

In view of the fact that recent improvements have made it possible to obtain at least as good, or better, results with valve rectification, we will not deal further with the crystal.

Valve Rectification.   The valve is used to give rectification, commonly in one of two ways, namely, " Anode Bend Rectification " and " Grid Leak Rectification." The first depends on the non linearity of the plate current, grid volts curve, and the second on the grid current, grid volts curve.
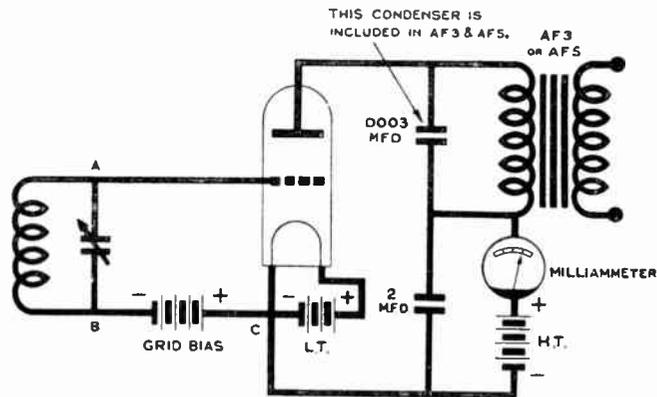
Fig. 51.

Anode Bend Rectification.

Anode Bend Rectification.   Figure 51 shows the connections of a valve to its circuits for this purpose. To successfully adjust the valve, one must decide the plate volts (120 volts as a minimum), and know something of the plate current—grid volts curve for the valve. Then the grid bias must be made sufficiently negative to bring the working point on to the curved portion of the plate current curve.
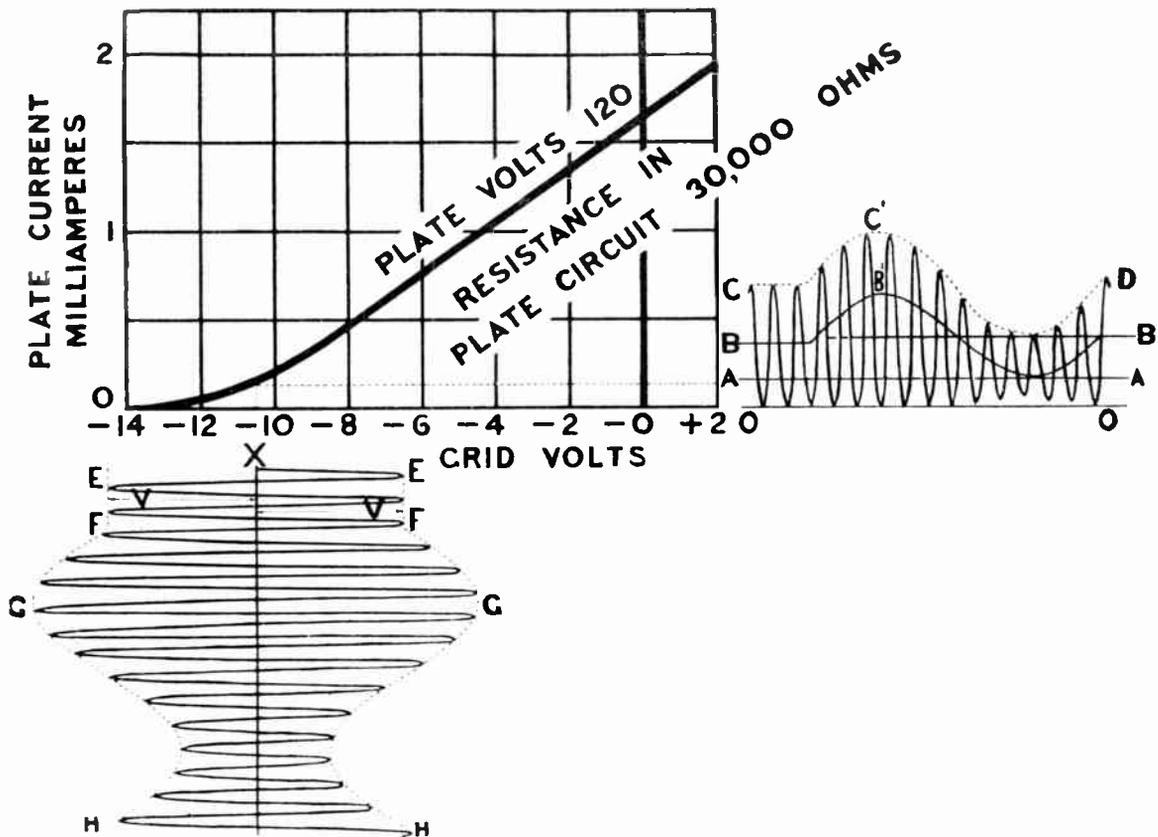
Fig. 52.

Figure 52 shows the characteristic curve of a particular valve : the horizontal distances representing the voltage of the grid A relative to that of the filament C, and the vertical lines the corresponding plate currents of the valve. Anode bend rectification is obtained with such a valve by adjusting the grid bias battery so that the voltage of the grid before a signal comes in is that of the point X, for this particular valve – $10\frac{1}{2}$ volts. A milliammeter in the plate circuit would then show a reading of plate current equal to OA. These values will vary for different valves, and tables given later

show this point for typical valves. As valves of the same type and make vary considerably, some additional adjustments may be necessary to get the best points.

When the radio signal is received, the voltage of the point A rises and falls with the frequency of the carrier wave. This is indicated by the waves from E to F. If the microphone is receiving music, then the carrier wave varies in strength correspondingly, as shown by F G H. Looking at the diagram, it will be seen that if the voltage of the grid rises above the mean value represented by point X, then the plate current rises also ; but when the voltage, due to the other, the negative half of the carrier wave, falls below the point X, then the plate current changes or falls by only a small amount. In the diagram, the current for the positive half rises from the value A to the value C, and for the negative half falls from A to O. When the microphone is modulating the carrier wave, by the note represented by F G H, the peaks of the plate current, due to positive halves of carrier wave, form the line C C¹ D, while those due to the negative halves keep near the line O O. In the diagram of connections, a bypass condenser C¹ is shown in the plate circuit across the transformer primary. Its function is to smooth out the very rapid pulses of the plate current corresponding to the radio frequency signal. In effect, it gives a mean current represented by the line B B¹ B and so acts that when the current is in excess of that mean value it receives the excess energy, and when the current as shown in the diagram falls below the mean value it gives out the energy and maintains the current to the mean value B B¹ B.

Another way of looking at this action is to say that C¹ by-passes the very rapid pulses of current corresponding to the carrier wave, leaving behind the comparatively slow pulses of current indicated by line B B₁ B, *which represents the sound in the studio.* This latter, or audio frequency current, passes through any other apparatus in the plate circuit, such as the transformers AF3 or AF5. The milliammeter,

however, cannot respond to the pulses of current represented by the line B B¹ B, and only shows the mean current represented by the straight line B B.

This description applies to all anode rectification, but if the signal received is small, so that the length X E is also very small, the action occurs wholly in the part of the valve characteristic curve which is curved. In such a case, the form of the rectified current B B¹ B does not exactly represent the music ; some additional vibrations are added.

Mathematical calculations show that in addition to the original fundamental note, there is an octave or second harmonic, whose amplitude is $\frac{m}{4}\%$ of that of the fundamental, where m is the percentage of modulation occurring in the broadcasting station. Thus, if the modulation is 20%, then the second harmonic will be 5% of the fundamental ; with higher degrees of modulation the amount of harmonic will be so great as to produce bad music. There is a further disadvantage in working with a weak signal, namely, that the valve impedance corresponding to points on the bend of the characteristic curve is high, and this affects the question of further amplification.

On the other hand, if by suitable high frequency amplification the variations of the voltage of the grid of the detector valve are large, then the current corresponding to the point B can be well within the straight portion of the valve curve.

If this is high enough, then the Station can modulate strongly without the variations of the carrier wave which represent the music, entering the bend, and so avoiding the introduction of undesirable harmonics. Further, under these conditions, the impedance of the valve is at its lowest, and further amplification is easier. This is proved later by measured figures, and curves of a good example of anode bend rectification.

It is possible, however, for the wireless signal, if great enough, to make the grid positive during part of the time, thus causing grid current to flow. This will introduce serious damping into the tuned grid circuit. In the tables at the end of this book we give rectified current values which must not be exceeded for typical valves, if this distortion is to be avoided.

Radio Meter. The use of a Radio Meter reading 0 to 2 milliamps, in the plate circuit of the detector, is very desirable if anode bend rectification is employed. It enables adjustment of grid bias to be made, to ensure correct rectification, and it also gives the indication of the relative signal strength, thus enabling the set to be so adjusted that the signal on the detector is always sufficient to give good rectification and the best reproduction under all conditions.
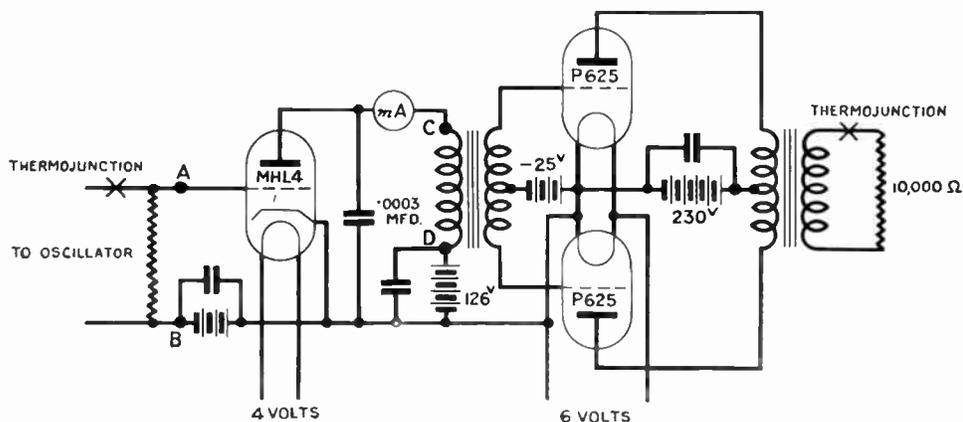
Fig. 53.
Set used for Anode Bend Measurements.

Measurements on Anode Bend Rectification. Figure 53 shows the detector and single stage amplifier used. The detector valve is an MHL valve, 126 volts on the plate, connected by an AF5c Intervalve Transformer to two P625 valves in push pull, 230 volts on the plate, coupled by an OP3c Output Transformer (1 : 1), to a 10,000 ohms resistance in series with a thermojunction. All volts input and output were measured by resistances and thermojunctions.
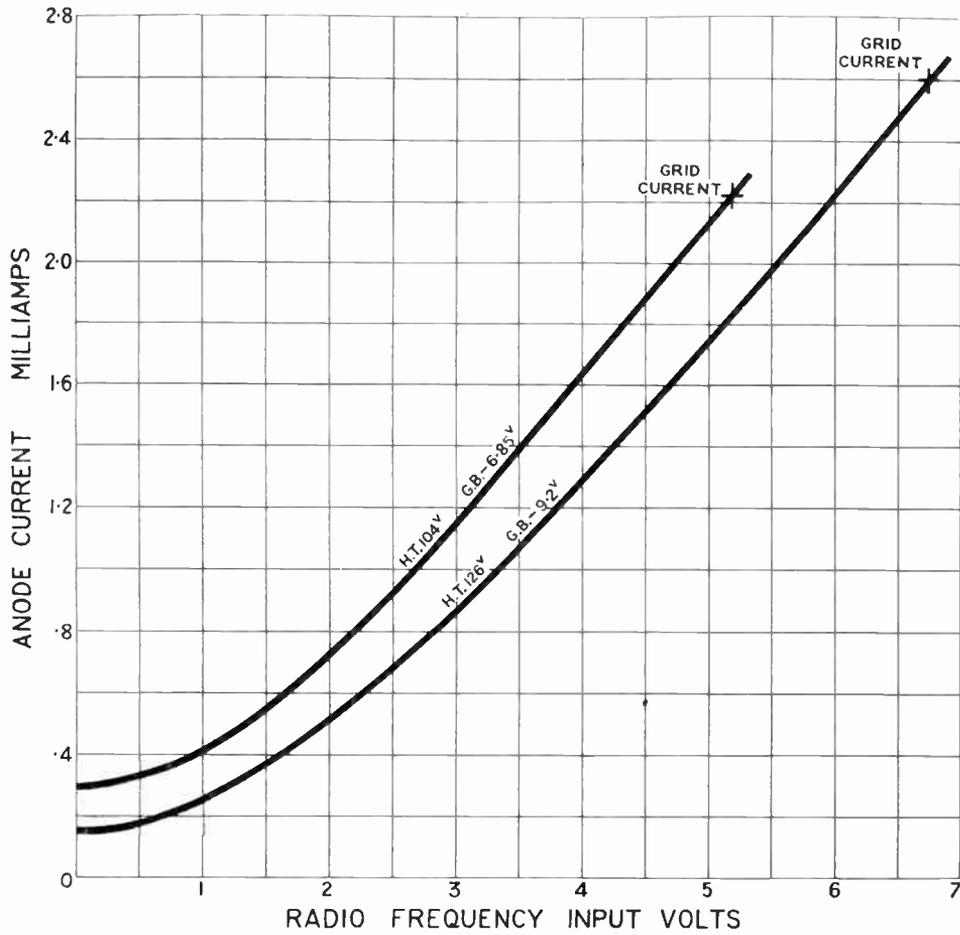
Fig. 54.
Anode Bend 104 Volts H.T.   G.B. – 6·85 Volts.
Input 3·16 Volts 30% Modulated 400 Cycles (1·17 ma).   Output 123 Volts.
Input 1·3 Volts 30% Modulated 400 Cycles (·49 ma).   Output 51 Volts.
Audio Amplification Transformer Primary to Output, 12·4.

Curves I and II, Figure 54, show the plate milliamps of
the MHL4 valve for various inputs of radio frequency

(870 metres) unmodulated, to the grid (A B) of the valve,

    I.   For plate voltage 104 volts; grid-bias −6·85 volts.

    II. For plate voltage 126 volts; grid-bias −9·2 volts.

On each curve the voltage at which grid current begins is marked " X."

The first thing evident (taking curve II) is its straightness, once the input exceeds 1 volt. It is clear that if the carrier wave input were 3·68 volts, causing the plate current to change from ·15 to 1·15 milliamps—an increase of 1 milliamp—that the changes of amplitude due to, say, 70% modulation, viz.:—1·1 volts to 6·26 volts, would occur along a part of the curve so nearly straight that little harmonic is introduced. On the other hand, if the carrier wave were only $\frac{1}{2}$ volt the changes would occur on a part of the curve which is parobolic (square law rectifier), and one would then get $\frac{70}{4} = 17\frac{1}{2}\%$ of second harmonic introduced.

The other interesting fact is that for the ·52 volt signal the plate current change is ·18 − ·15 = ·03 milliamps. But, except for the very low notes, the audio output, for the same modulation, will be proportional to the milliamp change. Hence, we find for the signal ratio $\frac{·52}{3·68} = \frac{1}{7·07}$ that the audio output ratio will be $\frac{·03}{1·0} = \frac{1}{33·3}$. Hence, the audible output will only be $\frac{7·07}{33·3} = \frac{1}{4·7}$ of what it would have been if the rectifier had been perfect and linear. If the rectifier had been one with a square law (parabolic), the audio output for signal ratio $\frac{1}{7·07}$ would have been $\frac{1}{50}$.

When obtaining curve I (plate voltage 104), the following measurements were taken. The input to the detector valve was modulated with a 400 cycle note 30%, voltage of 3·16

(modulated) gave an audio output of 123 volts; the reduced input voltage of 1·25 (modulated) gave an audio output of 51 volts, almost accurately in proportion.

The following more complete measurements and curves were taken with conditions as for curve II. The modulated input was kept constant to give a constant plate current of 1·15 milliamps (3·68 volts unmodulated), 3·56 volts modulated), the modulation being 30½% with notes from 50 cycles to 8,000 cycles. The measured output volts are given in Table I.

OUTPUT VOLTS. 30·5% MODULATION RF INPUT TO GRID 3·56 VOLTS CONSTANT

MHL4 126 PLATE VOLTS
− 9·2 GRID VOLTS
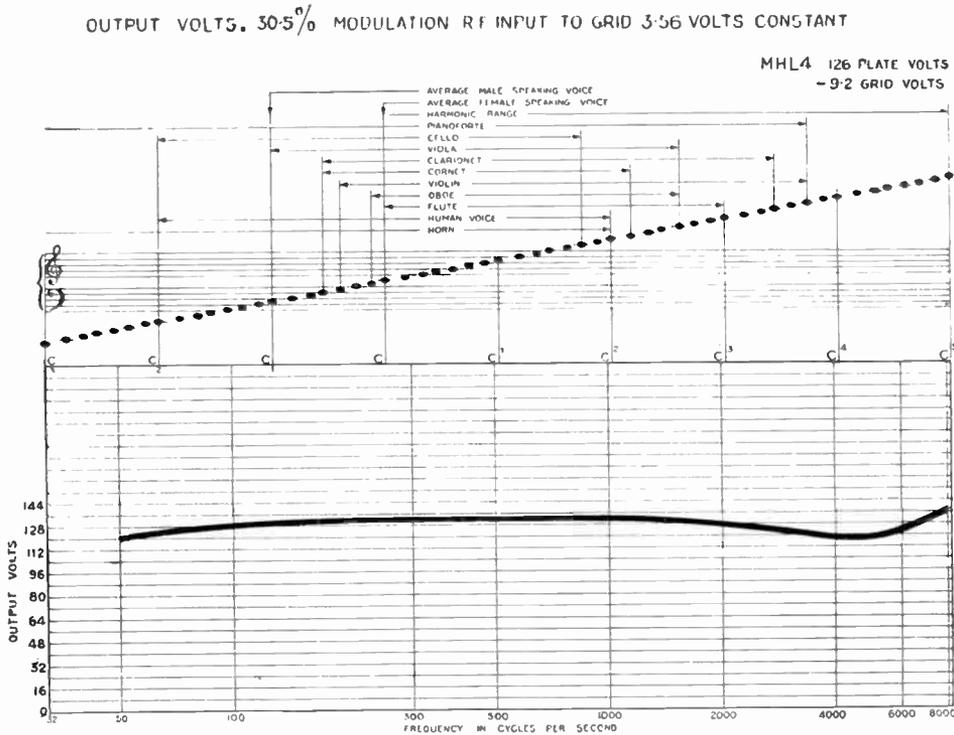
Fig. 55.
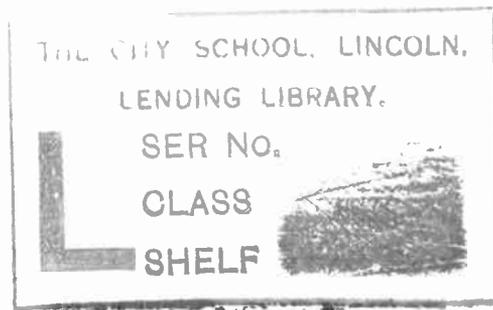
TABLE I.

Frequency in cycles ...	50	100	300	500	1,000	2,000	4,000	6,000	8,000
Output volts	121	129	132	132	132	126·5	117·5	125	136·8

The curve, Figure 55, shows these results. It is a very interesting curve. To be ideal it should be a straight line. The very small departure from ideal includes losses caused by rectification, the transformer AF5c, and the output transformer. The small reduction of the 50 cycle note proves the earlier statement that with large signals the anode bend rectifier is used under low *impedance* conditions for the valve in question.

The figures of the two following Tables were taken at the same time.

Firstly, the bias of the MHL valve was reduced to 3 volts, giving 6 milliamps through the AF5c primary and an audio frequency signal applied to its grid and adjusted to give constant output voltages 135·4 volts.

TABLE II.

Frequency in cycles	100	500	1,000	4,000	6,000	8,000
Input to give 135·4 volts output	·838	·78	·775	·671	·558	·515
Amplification	162	174	174	201	243	263

Figure 56 shows these figures.
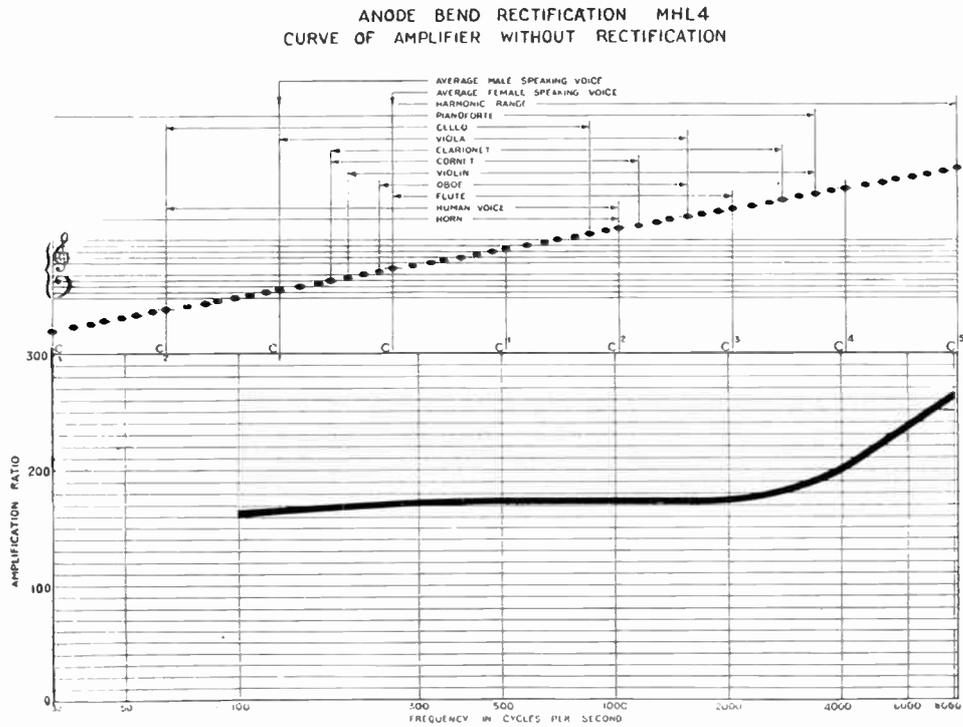
Fig. 56.

An audio frequency signal was then applied to CD, the primary of the AF5c, and measured to give a constant output voltage of 135·4.

TABLE III.

Frequency in cycles	50	100	300	500	1,000	2,000	4,000	6,000	8,000
Input to AF5c to give 135·4 output ...	12·3	12·2	11·9	11·8	11·6	11·46	9·82	5·4	3·25
Amplification	11·01	11·1	11·38	11·46	11·68	11·8	13·79	25·1	41·7

This clearly shows the second resonance effect dealt with in a later section, when the damping of the valve resistance is removed.

A comparison of the 500 cycle figure of Table III to the equivalent figure of Table II shows that the valve is amplifying 15·2 times, and probably has an amplification factor of 15·5.

Using the 500 cycle figure of Tables I and II, we can get an idea of the " efficiency " of rectification. Thus, 3·56 volts of $30\frac{1}{2}\%$ modulated radio frequency gives 132 volts output. But ·78 volts audio signal on the same grid gives 124·5 volts output, or ·765 audio signal will give 132 volts output. Hence, 3·56 × 30·5 × efficiency = ·765, whence the efficiency is seen to be 70·5%.

Strong signals. *Anode Bend and Grid Leak Rectification.* It will be shown later that the efficiency of grid leak rectification for large signals is also about 70%. Hence, we can say that for these large signals, *anode bend rectification gives as large an output as grid leak rectification.*

But it is otherwise with weak signals. Above, it was shown that with a signal of ·52 volts modulated 30·5% gives $\frac{1}{33\cdot3}$ of the output of the 13·56 volts signal, or $\frac{132}{33\cdot3} = $ 3·96 volts.

Anticipating the figures of the next section, a ·52 volt signal modulated 30·5% would give
·52 × 30·5 × ·65 × 174 (amplification of amplifier)
= 18 volts output with the same amplifier.

Weak signals. Hence, *for this weaker signal (·52 volts) grid leak rectification is $4\frac{1}{2}$ times as efficient as anode bend rectification.*

The figures illustrate another important point. The output of 135 volts is just the overloading point of the two P625 valves in push pull. With this very desirable signal, 3·56 volts, we just overload with $30\frac{1}{2}\%$ modulation, therefore, on the peak passages (often 80% modulation) we should overload seriously. In fact, with this modulation we should get 31 volts (R.M.S.) across the primary of the transformer and should only have a ratio of 1·3 to 1 instead of $3\frac{1}{2}$ to 1. The question of adjusting the audio-frequency amplification to suit the best radio frequency signal will be dealt with later.

Grid Leak Rectification. It has been shown how the plate current curve of a valve can be used for *anode bend rectification*. But a valve has a grid current curve which is also nonlinear, and in some respects better suited for our purpose. Using this curve, we have the well-known "*grid or grid leak rectification*," which is very popular owing to its greater sensitiveness to weak signals, and its simplicity.

The greater sensitiveness is due to the fact that the rectification takes place wholly in the grid circuit, leaving the audio frequency voltage corresponding to the modulation across the grid to filament. This then is amplified by the valve just as any other audio signal would be, giving us the full amplification of the valve. The valve thus serves a *double purpose,* and was described by Hull, in 1920, as the "*amplifying rectifier*."

Figures 57, 58, and 59 show the apparatus and connections for this. The return point B of the oscillatory circuit must be made so that grid current flows. In Figure 59, for an indirectly heated valve, B is connected direct to the cathode, since grid current begins to flow when the grid is – 0·8 volts.
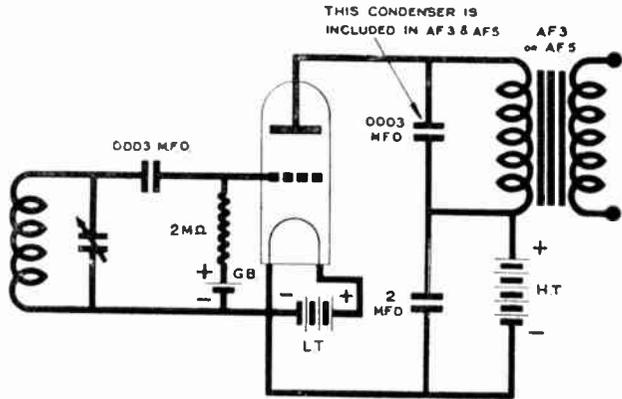
Fig. 57.
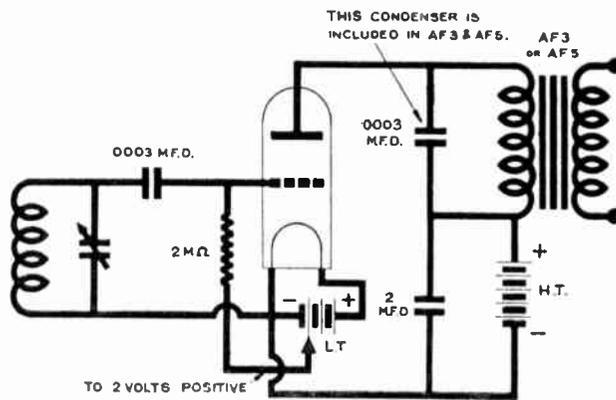
Fig. 58.
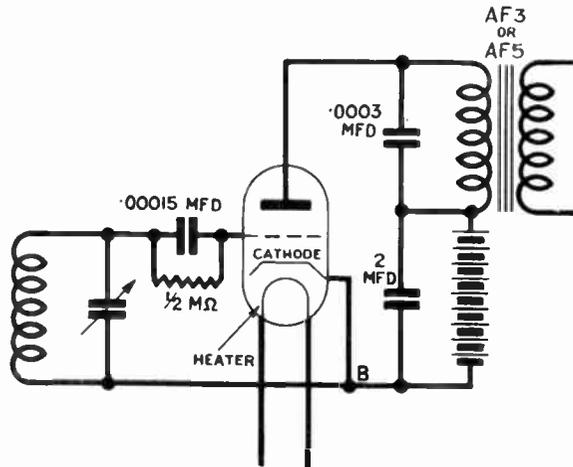
Fig. 59.

The other figures are for battery valves, and some positive bias is necessary. In Figure 57 it is obtained from a $1\frac{1}{2}$ volt bias battery, the negative end of which goes to the negative filament and positive to the grid leak. In Figure 58 the return is taken to 2 volts positive for 4 or 6 volt valves, or direct to the positive of the filament for 2 volt valves.

We will give in simple language an explanation of grid leak rectification. We do not know of any published easily understandable explanation, though the mathematical results of the process have been thoroughly worked out.

When unmodulated radio frequency oscillations are applied between the grid and filament, positive halves produce a greater increase of grid current than the decrease produced by the negative halves, leaving a mean increase, on account of the shape of the grid current curve. This increase of current flowing through the grid leak causes a drop of voltage at the grid, in addition to the drop already caused by the steady grid current flow through the leak when no signal is coming in. When modulation of the carrier wave occurs, this drop of voltage varies in accordance with the frequency and amplitude of the modulation. The blocking condenser C serves to smooth out the radio frequency pulses of grid current in a manner similar to the byepass condenser described in anode bend rectification. We are then left with an audio frequency voltage variation across the grid to filament ; the rectification is complete, and the valve acts as an amplifier to the voltage thus obtained.

Referring to Figure 60, A B C is the grid current curve of an MH4 valve, the return of the circuit being to the cathode or negative. The grid leak is $\frac{1}{2}$ a megohm, and through this the grid current has to flow, making the grid more negative, so that we get F as the steady potential of the grid before any signal comes in ; and for this case B F in
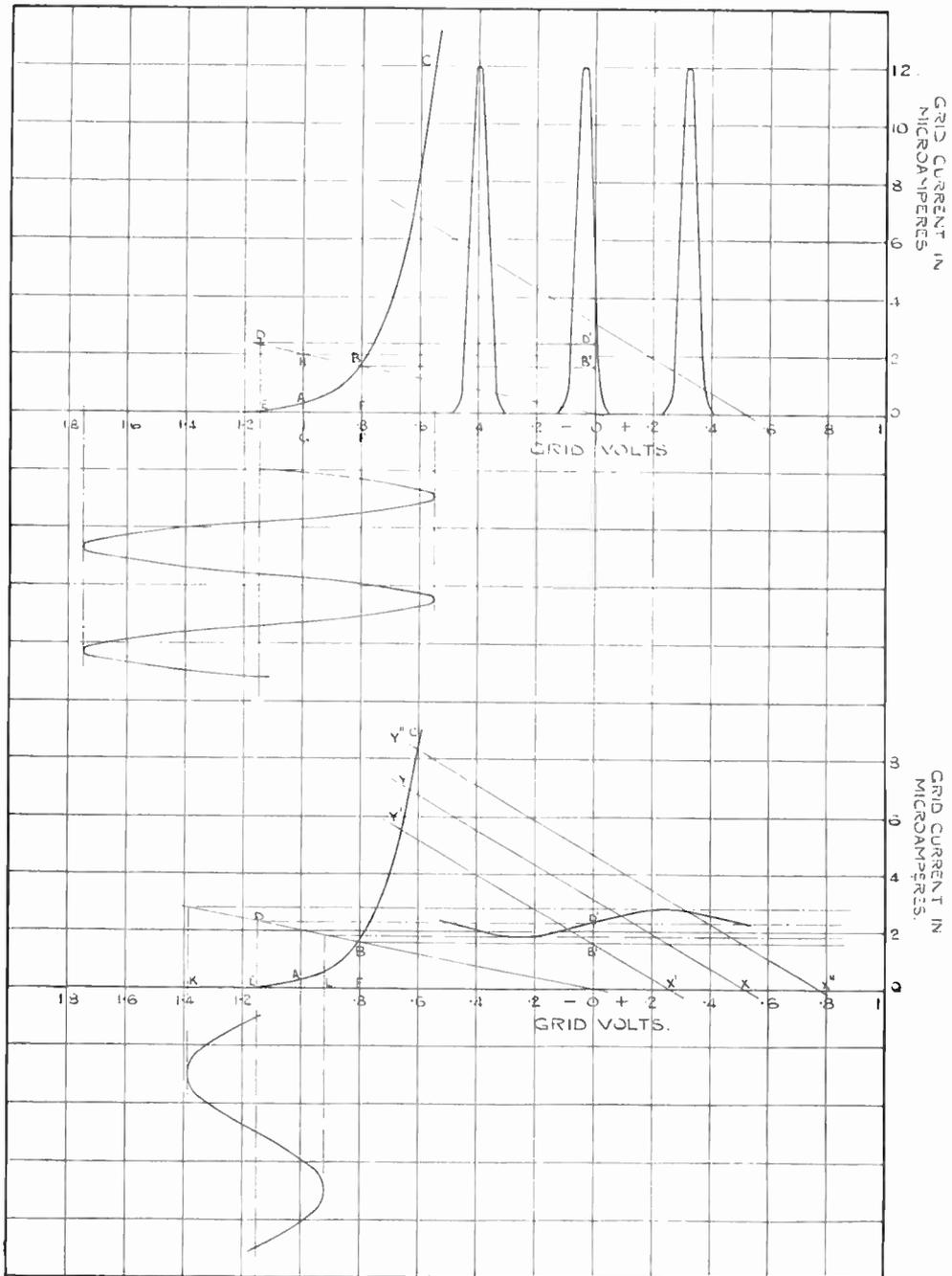
Fig. 60.

microamps $= \dfrac{\text{OF volts}}{\frac{1}{2} \times 10^6}$ ohms. The easiest graphical way to obtain B is to make O G $= 1$ volt, set up G H in microamps $= \dfrac{1}{\text{grid leak in ohms}}$. Join O H, and where it cuts the curve A B C we get B, and hence F.

When an unmodulated carrier wave comes in, since the condenser C offers no appreciable impedance to radio frequency, the voltage of the grid rises and falls and, as explained, the excess of grid current during the rise over the decrease during the fall, flowing through the grid leak, makes the grid still more negative, in the diagram by F E, so that E gives the working potential of the grid while the carrier wave is steady. There is no simple way of finding this point E—a mathematical process is necessary—but having found it and drawn the carrier wave oscillations, we are enabled better to visualise the action. The mean grid current is increased from O B₁ to O D₁ by the carrier wave. The method used to find the point E is the really beautiful graphical method evolved by W. A. Barclay, described in *Experimental Wireless,* August and September, 1927.

In Figure 60 the radio frequency oscillation corresponding to an R.M.S. voltage of ·425 is drawn below A B C. Notice that when the R.F. voltage is negative (to the left of E), no grid current flows. This applies to this particular case.

If the R.F. oscillation were of much less voltage, grid current would flow even when the oscillation was negative, and E would be nearer to F, but the gain would be greater during the positive halves than the loss during the negative. On the other hand, if the R.F. voltage is greater, E would move more to the right and grid current would not flow until the grid became more positive than this point, and would only flow during a portion of the positive half of the oscillation.

For the case in Figure 60 we have drawn the grid current flow or pulses to the right of A B C, and the mean of these is represented by the line D D¹ ; D must be the point where the vertical from E meets B F, a line representing the grid leak of $\frac{1}{2}$ megohm, since the current O D¹ passing through this causes O E. The smoothing is done by the condenser C (Figure 59) : this only applies while the carrier wave is unmodulated.

If the carrier wave is modulated, its amplitude increases and decreases, and the mean grid current rises and falls above O D¹ ; if the modulation were 100% it would fall to O B¹. This change occurring through the grid leak causes the voltage change F E to increase and decrease. In Figure 60 the curve of variation of E for 50%₀ modulation is shown below the grid current curve A B C, and to the right is shown the corresponding variation of mean grid current ; these curves represent the original music.

The three lines XY, X¹Y¹, X²Y², in this figure, are used to obtain graphically the limits of the 50% modulation voltage curve. They are equidistant from each other, and the amount of the harmonic in the wave form depends on the straightness of the portion of A B C intercepted between them ; it can be seen how straight this is.

We have stated that the condenser C, to the capacity of which should be added the capacity of the grid to filament of the valve, since this is quite appreciable, fills in the pulses of rectified grid current, and this is probably the most simple way of regarding its somewhat complicated action. But, unfortunately, this filling in action varies with the audio frequency, and also with the relative values of C to the resistance of grid leak used. The result is that the higher notes *may* not be reproduced properly. This subject will be dealt with in detail further on.

To visualise the rest of the action of the valve in grid leak rectification, place a transformer in the plate circuit of the valve, in series with this a D.C. milliammeter, and across the output of the transformer a cathode ray oscillograph.

When there is no incoming signal, the milliammeter will read approximately 2½ milliamps; this corresponds to the grid voltage O F, and the oscillograph will show nothing.

When the incoming unmodulated carrier wave is of suitable strength, the current will be reduced to approximately 2·4 milliamps, corresponding to the grid voltage O E, and the oscillograph will still show nothing. If the carrier wave now be modulated, the voltage, length F E, will increase and decrease with the modulation frequency, as explained.

The milliammeter will show no appreciable change, for its needle cannot follow the modulation variations, and it will still read 2·4 milliamps; the oscillograph, however, will show the variations of F E, magnified by the amplification of the valve, and will represent the music which is passed on to the later states. Though the rectification is complete in the grid circuit, and the condenser C smooths out the radio frequency pulses, this smoothing is not perfect, and the modulation note in the plate circuit still contains some radio frequency pulse. This is often used for reaction purposes, but must be prevented from passing on to the later stages, hence it is usual to connect across the audio frequency transformer, or other impedance in the plate circuit, a small condenser, generally of ·0003 mfd, to give the additional pulse smoothing; or, in other words, to by-pass the radio frequency current.

It has been pointed out, in dealing with anode bend rectification, that the stronger the radio frequency input, the more like the original is the audio-frequency component obtained.

Before discussing the same question for grid leak rectification, we will obtain some idea of the strength of signals necessary.

Three years ago, two stages of low frequency amplification were almost universal, and with the best power valve, the DF5A class, then in common use, and the smaller modulations possible in the studios, the radio frequency voltage on the grid of the detector valve would be about ·024 volt R.M.S., overload at 40% modulation and 1,800 amplification.

At the present time, with such a three valve set as the A.C. Mains Receiver, Model 32, the equivalent figure is ·29 volts R.M.S., overload at 80% modulation and 120 audio amplification, or about 12 times as great as formerly.

In larger sets using valves of the LS5A class in push pull, the signal may be as much as 2 volts R.M.S.

With weak signals, up to ·05 volts R.M.S., such as the first, it can be shown mathematically that, as in anode bend rectification, using small signals, a second harmonic of $\frac{m}{4}\%$ is introduced, so that with peaks of modulation of 80%, a 20% harmonic is there—quite a serious distortion.

Mention has been made of a possible frequency distortion due to the grid leak condenser combination. If the resistance of the leak is large in relation to the capacity of the condensers, there is a large loss of the higher audio notes, the impedance of the combination is much lower at 6,000 cycles than at 500 cycles. This effect is very pronounced with the use of the 2 megohm grid leak, and ·0003 mfd condenser, which is so common.

The above two facts account for the bad reputation that grid leak rectification has had in the past. This reputation is quite undeserved if moderately strong signals are used

and suitable values given to the components. Mathematics is only partly helpful for these larger signals—experiments and measurements have to be made. The following curves were taken to find out what rectification defects existed in the Receiver Model 32.

In this set, an MH4 valve was used as detector, 70 volts on the plate, with the grid return to the cathode. Grid leak, $\frac{1}{2}$ megohm, and the grid condenser ·00015 mfds. The valve is followed by an AF5 Transformer, which feeds a P625 valve in whose plate circuit is an output transformer of 1 : 1 ratio, with 5,000 ohms across the output.
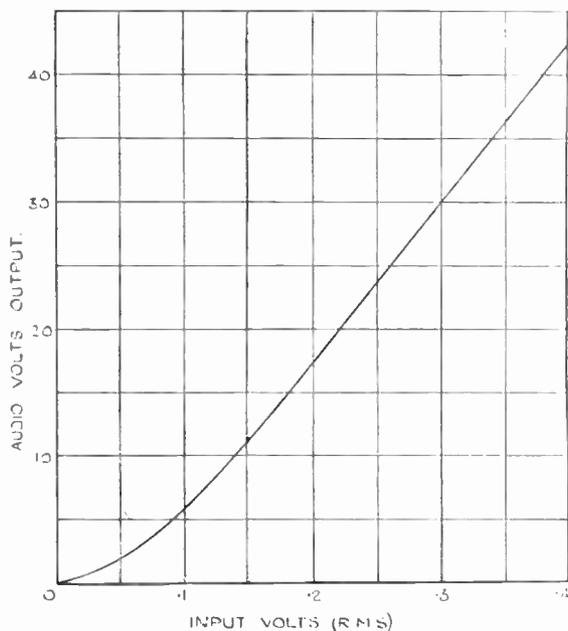
Fig. 61.
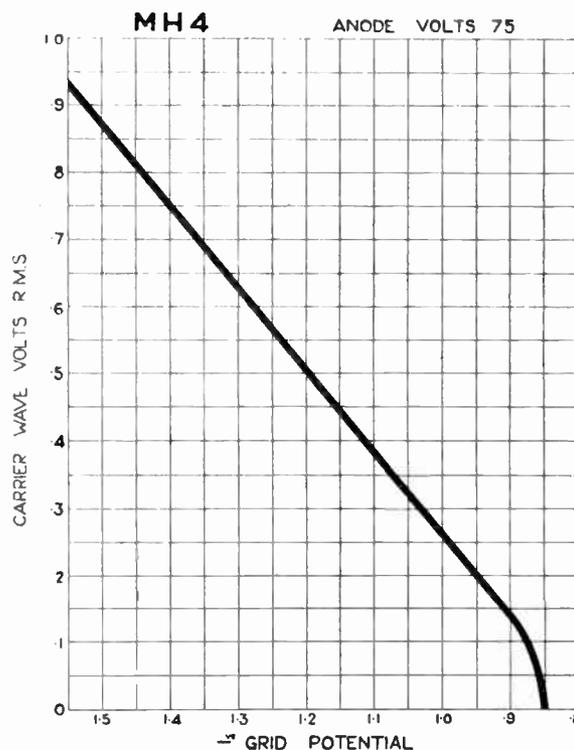
Fig. 62.

The curve, Figure 61, was taken with an input to the *detector valve grid* of a wave length of 250 metres modulated 32% with a 400 cycle note. The output audio frequency volts across the 5,000 ohms resistance are plotted against the input volts, R.M.S. radio frequency. The curve is straight from ·075 volts upwards. If the mean input were ·25 volts, then for a 75% modulation ranging from ·075 to ·425 volts, we should expect no distortion. Figure 62 proves this in a more convincing way. The carrier wave changes the plate current due to the increase of negative grid potential, here for various values of the carrier wave the measured grid potentials to cause the same plate current change have been plotted. Note that from signal ·9 volts to ·09 volts this curve is straight, so that as long as modulation keeps the carrier wave above ·09 volts there is no distortion.
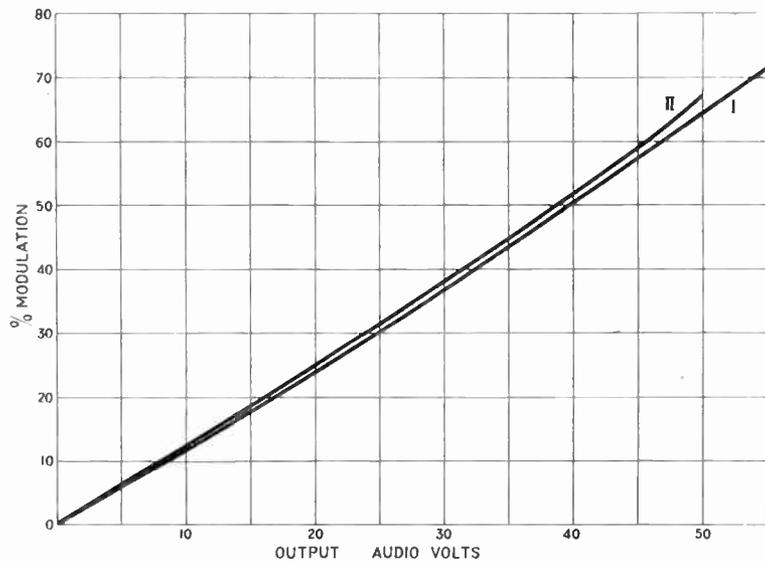
Fig. 63.

MH4 Valve.

C = ·00015. Grid Leak ½ megohm. Plate H.T. 70 Volts before reception, after reception 80 volts, due probably to reduction of plate current.

Curve I. Wavelength 250 metres, unmodulated input to detector ·21 volts (R.M.S.).
Curve II. Wavelength 1,500 metres, unmodulated input to detector ·23 volts (R.M.S.).

Curves (1) and (2) of Figure 63 show that this is practically obtained. Here we have a constant radio frequency signal modulated various percentages from 7% up to 70%, and the audio frequency output volts plotted against the percentage modulation.

(1) is for   250 metres and ·21 volts R.M.S. input.
(2) is for 1,500 metres and ·23 volts R.M.S. input.

The straightness of the curves show that there is no appreciable distortion. An examination of the output wave form with 70% modulation in a cathode ray oscillograph confirmed the want of appreciable distortion.

Fig. 63a.

H
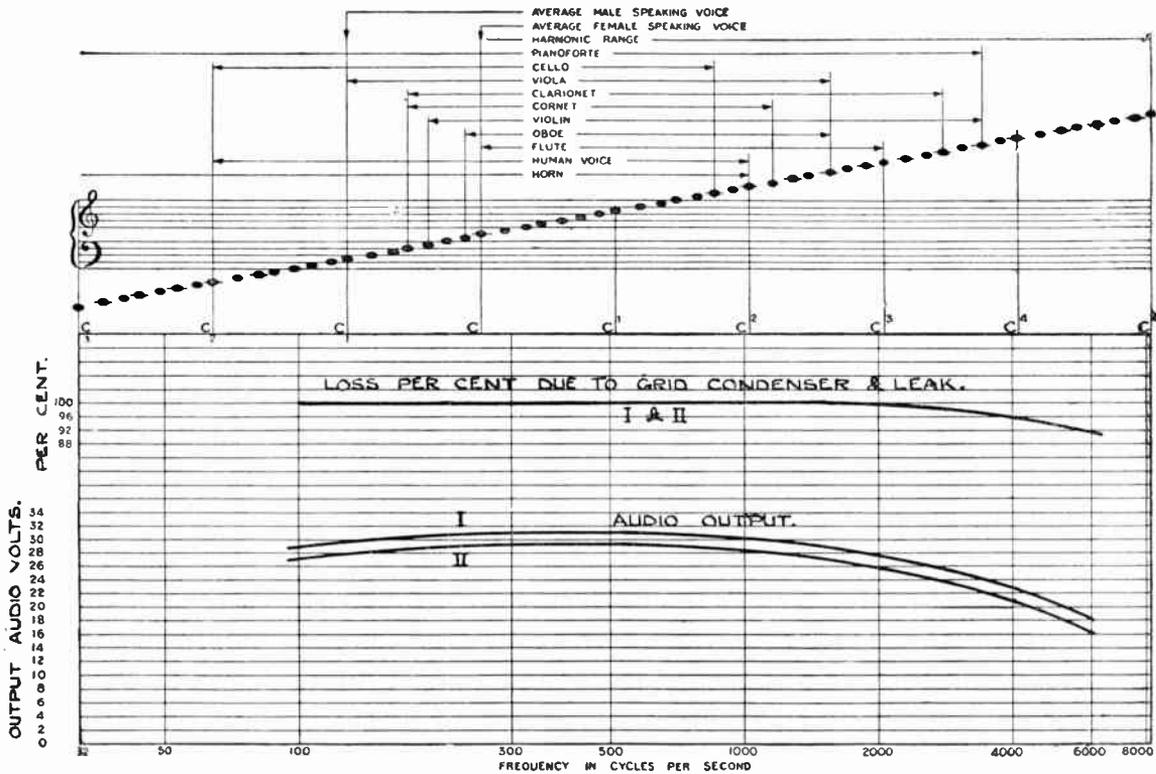
In Figure 63a the two lower curves (1) and (2) show the audio output volts of the set for (1) an input signal of ·23 volts radio frequency of 250 metres modulated $39\frac{1}{2}\%$ with notes from 100 cycles to 6,000 cycles and (2) an input signal of ·23 volts radio frequency of 1,500 metres modulated $36·2\%$ with notes from 100 cycles to 6,000 cycles. These curves include all losses, those due to grid leak combination, the audio frequency transformer, and the output transformer. Above these is the curve showing the percentage loss due to grid leak combination. It is almost the same for the two wavelengths, and at 6,000 cycles amounts to about 8%. These curves show that for the usual signal voltages to the grid of the detector required to give normal sound output, viz.: ·25 volts to ·3 volts, no appreciable distortion is introduced by the use of grid leak rectification.

MODULATION NOTE FREQUENCY.
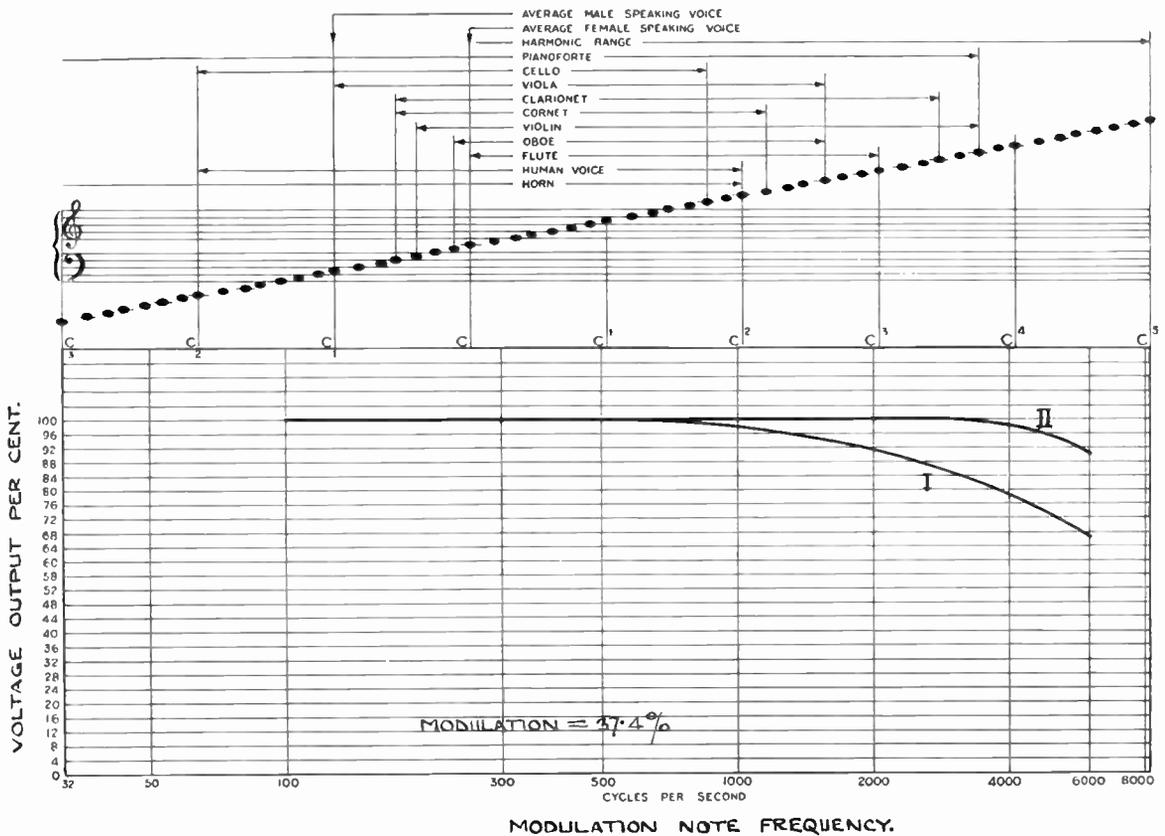
Fig. 64.

Figure 64 shows two curves demonstrating the loss incurred by the use of higher resistance grid leaks. The valve is an MHL4 with a grid condenser of ·0003 mfd. The input signal was ·23 volts of 250 metres modulated 37·4%. Curve (2) shows the loss for a grid leak of $\frac{1}{2}$ megohm, and (1) that for a grid leak of 1 megohm, this being very much greater than in (2). No doubt if 2 megohms had been used the falling off of the higher notes would have been much greater.

Power Grid Rectification.

We have dealt with this question fully, as grid leak rectification has recently come into favour again, using large signals—2 volts and above at the grid of the detector valve—under the name of "*power grid rectification,*" and the public have been led to understand that grid leak rectification for any smaller signal was bad. We have proved conclusively that for input signals as low as 0·23 volts this is not true with suitable valves and components; no doubt, as the signal decreases below this value—especially if too large a condenser and leak is used—the defects will creep in and grow larger, as the input grows less.
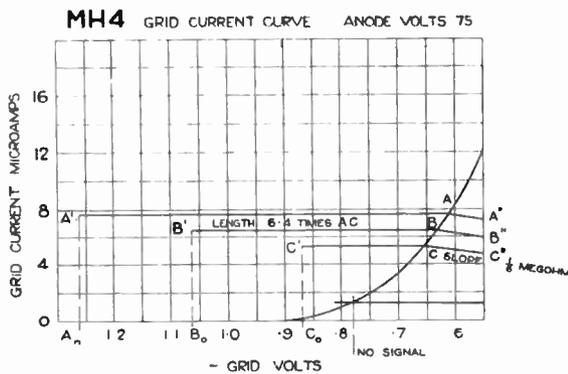
Fig. 65.
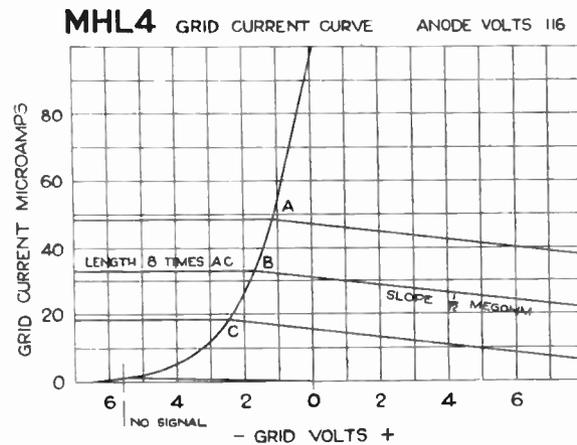
Fig. 66.

To get a graphical estimate of the distortion or harmonics introduced, consider the two Figures 65 and 66.

Figure 65 represents the grid current curve of an MH4 valve with 75 volts on the plate. Barclay's graphical construction is used to obtain points B_0, B, and B^1, corresponding to an unmodulated radio frequency signal of ·353 volts R.M.S., or 0·5 volts peak. If this signal is modulated 50%—quite a large modulation—the same construction gives us points A_0, A, A^1, and C_0, C, C_1, corresponding to the extreme values of the carrier current during modulation. The result of the rectification is that B_0, C_0; and B_0, A_0 represent the peak values of the audio frequency voltage obtained. Now, if the intercept A B C of the grid current curve were straight, the audio frequency signal obtained would be a *duplicate* of the original modulation. The departure of A B C from its chord as a percentage of the length A_0 C_0, gives an approximate measure of the harmonic introduced; clearly in this case it is small, being 2% or less.

Incidentally, from this curve the efficiency of rectification was calculated to be 77%. The grid swings between −·87 and −1·26 volts for the audio signal. For 80% modulation it would swing between −·75 and −1·38. These figures are useful in deciding how the valve will work as an amplifier. The efficiency depends on the steepness of A B C, the intercept on the grid current curve.

Figure 66 shows a similar construction for an MHL4 valve, with 116 volts on the plate for a carrier wave of 2·12 volts R.M.S., 3·0 volts peak; clearly a case of the so-called power grid rectification, the grid leak being $\frac{1}{4}$ megohm. This is modulated 50% as before. In this case, we calculate the harmonic introduced by the curvature of A B C to be 1·8%.

The grid swings between −1·55 and −4·01 volts. For a modulation of 80% the grid swings between −·84 volts and −4·77 volts. The efficiency of rectification for 50% modulated signal is $80\frac{1}{2}$%. These efficiencies, given by Barclay's

construction, are higher than the measured values. With an MH4 valve and a signal of ·5 volt, we only measure 80% of the calculated figure. Some of the discrepancy is due to the fact that the total resistance of grid leak and grid to filament resistance is lower under working conditions than the measured D.C. resistance of the grid leak. The other part is due to incomplete filling in between the pulses by the grid condenser. A large condenser, not usable because of frequency distortion, fills in better and gives efficiencies much nearer Barclay's.

It is clear from these curves that with a signal as small as ·35 volts and an MH4 valve there is no more harmonic introduced than with 2·12 volts and an MHL4 valve, and in either case it is negligible up to modulations of 50% and not very appreciable with modulations of 80%. This is only a graphical confirmation of the experimental results.

It should be pointed out, that the radio frequency signal at the grid of the detector valve is not usually chosen, as it depends on the grid swing without overloading, available at the power stage, and then on the amount of audio frequency amplification incorporated. These are generally chosen and the grid input is then settled. Typical cases are worked out in a later section, dealing with typical receivers.

This is not generally recognised in specifications for hospital installations which the writer has seen.

So far, we have only dealt with grid leak rectification on the grid side of the detector valve. This has to act as an amplifier to the signal obtained. This, with the necessary diagrams, are dealt with on page 139, in the section on audio frequency amplification.

Power Grid Rectification. A reason is usually given for its good qualities, which is to a great extent fallacious. If the grid current, grid volts curve of Figure 60 be extended to more positive values of

the grid, especially for the higher plate voltages used, it will be found that the curve becomes straighter; and it is said, arguing by analogy from anode bend rectification, that with the much stronger signals, we are using the straight portion of the curve during modulation. But in the Figure mentioned, it was shown that when the carrier wave is applied, the mean grid potential is moved to the left to E, and the stronger the signal the further it is moved. With such signals as $2\frac{1}{2}$ volts R.M.S. it is moved so far to the left, that even during modulation, the oscillation peaks do not extend far into the straight zone; but more curious still, the point is so far to the left that the radio frequency oscillation has to increase the potential of the grid considerably above its mean, before any grid current flows, so that it only flows during *the peaks* of the positive half.

There is a physical reason for the perfection of the rectification when using only the peaks of the oscillations, but it does not lie in the above argument. The reason appears in Figure 66 in the straightness and the shortness of A B C, the intercept on the grid current curve by Barclay's graphical construction. The construction merely integrates graphically the grid currents caused by the oscillations, allowing for the back E.M.F. caused by the rectified current flowing through the grid leak.

AUDIO FREQUENCY AMPLIFICATION.

As explained in the last section, the current obtained, after the detector valve has functioned, is an Audio-Frequency current, or, as it is often called, a low-frequency current, and represents very closely the sound in the Studio. It is usually strong enough to work a pair of telephones, which are sensitive instruments, but is incapable of working a Speaker, as the sound energy required from this is enormously greater than that required from a pair of Telephones. It is necessary, therefore, to magnify or

amplify this audio-frequency current. In doing so, *to avoid distortion, the amplification must be carried out in such a manner that the increased signal differs only from the original in that it is larger.* To appreciate what this means, one must recall that sound is caused by a rapid vibration of air particles and the pitch of a note is fixed by the frequency, or number of vibrations, per second. Instrumental notes have pitch frequencies from 25 to 3,500 per second, but no note is a simple vibration of one frequency but a combination of vibrations at frequencies which have a simple relation to one another. The lowest, called the fundamental, gives the pitch, whilst the others, called harmonics, give the tone or quality, and cause the sound of one instument to differ from that of another, although the fundamentals in each case have the same pitch. The frequency of audible harmonics may be as high as 10,000 per second, and thus the musical vibrations have a frequency range of from 25 to 10,000 cycles per second.

Therefore, the requirements to avoid distortion are :—

(1) That all electrical currents corresponding to notes from 25 to 10,000 cycles shall be amplified equally ;

(2) That the apparatus used shall not introduce notes not in the original Broadcast.

Audio frequency amplification is obtained by applying the weak signal across the grid and filament of a valve in whose plate circuit is an " Impedance " which must be able to pass the steady plate current of the valve. The *changes* of current caused by the varying grid voltage pass from the filament to the plate, through the " Impedance," back to the filament ; they thus cause a voltage variation across the " Impedance " which is usually much greater than, but otherwise similar to that applied to the grid. By suitable arrangements we can use this greater or amplified voltage and so obtain audio frequency amplification.

But there is also a voltage variation between the filament and the plate of the valve, and it is said to have " resistance," usually known as the anode or plate resistance. We cannot, therefore, obtain the full conceivable amplification. For the benefit of those who prefer results in the form of symbols, the amplification to be obtained with a given impedance Z is

$$\frac{\mu Z}{(Rp + R)^2 + X^2}$$

where μ is the amplification factor of the valve, Rp is its plate resistance, R is the resistance of the impedance, and X that part of its impedance depending on its inductance and capacity, or its reactance.

Clearly, then, we must strive to make Z as large as possible consistent with any other limiting conditions, and where signals of varying frequencies, as in music and speech, are applied to the grid, we must allow Z to vary as little as possible with frequency; or, if it must vary, make it so large compared with Rp that its variations have a negligible effect.

Figures 67, 68, 69, 70, and 71 show various methods of connecting valves in order to obtain amplification.
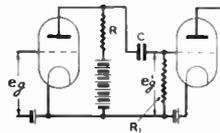
Fig. 67.   Fig. 68.   Fig. 69.

Fig. 70.   Fig. 71.

In circuits 67, 70, and 71, the impedance directly in the plate circuit is the resistance R. They differ from each other in the way in which the voltage across R is passed on to the next valve. In 68 and 69, an iron cored inductance is the impedance in the plate circuit.

Resistance capacity coupling.

The method of connection in Fig. 67 is known as Resistance capacity coupling. Where the amount of amplification required is not great, and care is taken in the design and value of components and the selection of valves, this can give good results.

The voltage across R is passed on to the grid of the next valve through the condenser C, the grid leak R_1 being necessary to keep the grid of this valve at its proper potential. C is necessary in order that the high potential at the plate of the first valve shall not affect the grid potential (settled by its bias battery) of the second. The great advantage of this method of coupling is that the amplification of even the very low notes can be just as great as that of the higher notes, and if precautions are taken, this constancy of amplification can be maintained even for the highest notes.

Campbell filter.

Figure 72 gives values and arrangements with an MHL4 indirectly heated valve used as detector. A Campbell filter is inserted to filter out the radio frequency component and

Fig. 72.

yet not lose the high notes. The first condenser of ·001 mfd. should include the plate capacity of the valve. The resistance R is 8,000 ohms. The coupling condenser of ·1 mf. should have a very high insulation resistance. The grid leak is 500,000 ohms. Such an arrangement should give a practically constant amplification of $7\frac{1}{2}$ from 25 cycles to 7,000 cycles.

The disadvantages of this method of coupling valves are :—

(1) Small amplification ;

(2) If large amplification is attempted the value of R must be large, consequently the capacity of the valve plate to earth and any additional capacity required to filter radio frequency shunting R_1, unless a proper Campbell filter is used, reduces the amplification of the higher notes of the musical scale.

(3) If R is large, then R_1, the grid leak, and C, must be large, and because of the possible leakage of H.T. potential to the grid of the next valve, it must be of mica, and therefore costly.

(4) Any radio frequency component not perfectly filtered in the detector plate circuit, can pass direct to the grid of the next valve and cause trouble. A resistance of 100,000 ohms, or a radio frequency choke in the grid lead close to the valve, reduces this effect.

(5) Since there is always a steady D.C. current flowing through R, the full battery voltage cannot appear at the plate of the 1st valve. With large values of R necessary for great amplification this can be serious.

(6) Connected with this difficulty is another. It is necessary in all receivers to use a decoupling system (see Section on Anode Feed System). If the value of R is great, the amount of decoupling resistance which can be used is small, hence a greater chance of " self oscillation " or " motor-boating."

In the typical sets described later, for the reasons given, the audio frequency amplification must be greater than that obtainable with one transformer coupled stage, but *very much* less than two. Hence, a resistance capacity stage is used after the detector where the signal voltage is not great.

Since the amplification required from this stage varies between 4 and 10, and the H.T. required is not great, all the above difficulties are avoided, and we have the ideal conditions for its use.

Transformer Coupling.

In the early days of Radio, when transformers were bad, there was an excuse for the use of resistance-capacity coupling.

But Ferranti have brought the design and manufacture of intervalve transformers to such a degree of perfection that for years such an excuse has had no status.

Primary Winding.

A transformer consists of two windings on an iron core. One, the *primary winding,* is, as in Figure 68, connected at one end, to the plate of the valve, and at the other end to the H.T. supply of the valve. This primary winding is the impedance Z previously mentioned. There is, therefore, a voltage variation across it similar to, but larger than, that applied to the grid : this is passed on to the next valve in the following manner. The voltage variation creates a similarly varying magnetic field in the iron core, which in its turn must of necessity cause a similar voltage variation

Secondary Winding.

in the other winding, called the *secondary winding.* This latter is connected across the grid to grid bias as shown, and so the voltage variation is passed on to this valve.

But another factor now enters. If the secondary winding has twice as many turns of wire as the primary winding, the voltage across it will be twice as great as that across the primary. So we have obtained a second voltage amplification, the first being that due to the valve. This is one of the great advantages of transformer coupling, though there are limitations to this gain.

A disadvantage is that the impedance of the primary winding varies with the frequency of the applied signal, and at the lower frequencies it is directly proportional to this frequency; thus, for a transformer with a primary of 120 henries inductance the impedance to the 50 cycle signal is

$$120 \times 6 \cdot 28 \times 50 = 37,700 \text{ ohms,}$$

and to the 100 cycle note it is

$$120 \times 6 \cdot 28 \times 100 = 75,400 \text{ ohms,}$$

or twice as much. At the higher frequencies the law of variation is not quite so simple. We cannot, therefore, obtain such a large percentage of the possible amplification of the valve in the case of the low frequency signal as in the case of the higher. 37,700 ohms is, however, relatively so large to the resistance of the valve—in this case 10,000 ohms—that the percentage of lost amplification at the worst is not great.

A summary is given of what is necessary in a good transformer and the way in which these are effected in Ferranti Transformers.

Self Induction. (1) That the inductance of the primary winding should be as great as possible, for the reasons given above.

Ferranti Transformers have a core of large cross section—in the AF5 Transformer it is 1·07 square inches. The stamping has no appreciable air gap and the number of primary turns of wire is large—in the AF3 and the AF5, 7,600 turns.

Magnetic leakage. (2) That the magnetic leakage of the transformer should be as small as possible. The magnetic field created by the primary is not completely passed on to the secondary : this is magnetic leakage.

In the AF3, AF5, and AF7 Transformers this is attained by interleaving the secondary between two half primary windings. This is quite important if the high notes are to be reproduced properly.

Self capacity. (3) That windings unfortunately possess another property besides " self induction," that is " self capacity," due to the close proximity of turns with voltage differences acting as condenser plates.

To reduce this self capacity and yet retain the large ratio of $3\frac{1}{2}$, we use a very fine enamel wire of ·002 inches diameter for the secondary winding and make the bobbins of skeleton form, in order to get as much air space as possible. The windings are also sub-divided. The photograph, facing page 128, shows the bobbins.
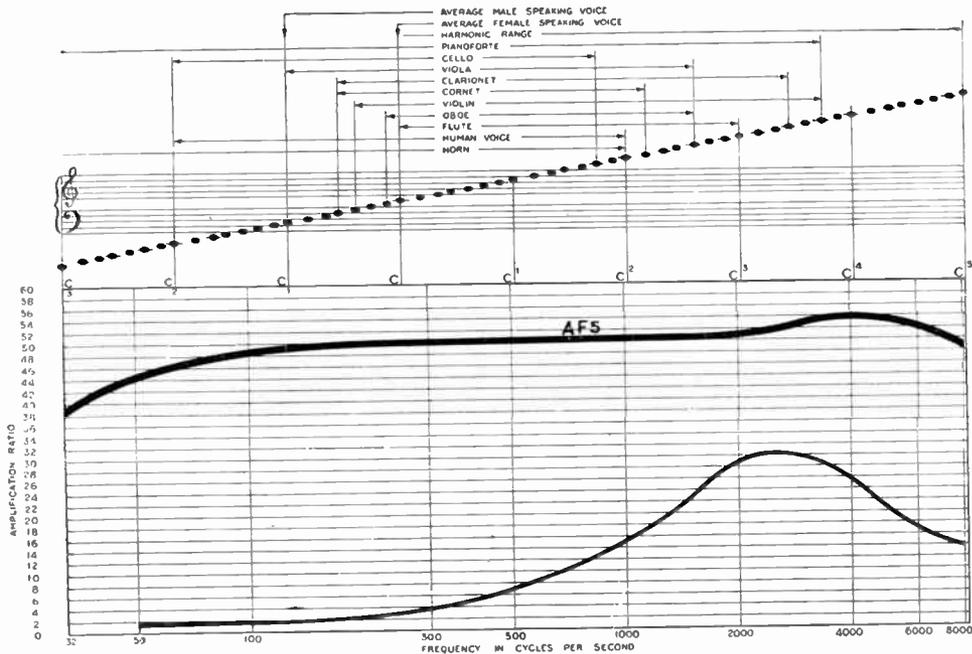
Fig. 73.

Figure 73 shows the difference between the results obtained with an AF5 Transformer and a cheap popular make.

It will be noticed that the amplification of the AF5 is very nearly the same for all frequencies, the greatest difference being the rise of about 18% at 4,000 cycles. Thus it approaches our ideal of equal amplification over the whole musical scale.

It is quite different with the cheap transformer. Its amplification at 2,500 cycles is 15 times greater than at 50 cycles, and it falls off badly at 8,000 cycles.
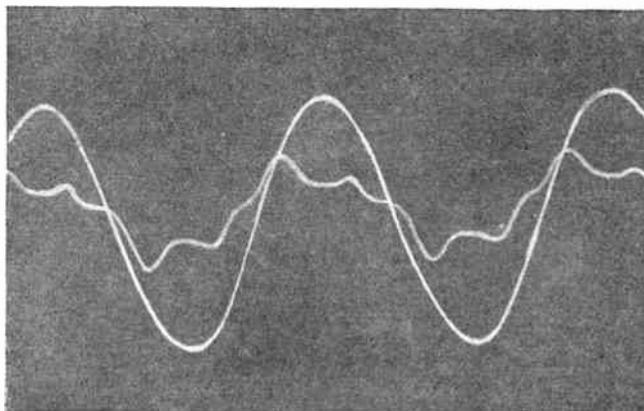
Fig. 74.

The oscillograph Figure 74 shows an even worse fault: the smooth curve is almost a pure wave form obtained from an amplifier containing two AF5's and an output transformer. It obviously had not been affected appreciably by the use of the iron cores. The other curve is from an amplifier with two cheap transformers in place of the AF5's. Clearly the pure musical note is badly distorted, or has a lot of added harmonics, making it sound harsh. The defect disappears at high frequencies, for which an explanation is given later.

In connecting a transformer secondary to the following valve it is important that the voltages of the energy supplies to this valve be adjusted correctly. In the tables giving valve data, these are specified for all suitable cases, and should enable the reader to do this.

There must be sufficient bias on this valve to prevent grid current flowing, even with the strongest signal impressed on it. In the case of battery valves, the point of grid current flow is when the grid is just at zero voltage, but with indirectly heated A.C. valves, grid current begins to flow at $-\cdot 8$ volts, so that the signal should never make the grid more positive than this.

If grid current flows, the inductance of the primary winding of the transformer will be reduced during a *portion* of one cycle of the musical note with a consequent distortion of the wave form.

The other requirement is that when the negative half of the signal makes the grid more negative than its mean, the greatest negative voltage reached must correspond to a plate current still on the straight part of the anode current, grid volts curve, so that the output is proportional to the input.

For some qualifications of this, see page 139. After adjusting the voltages correctly, one must see that they do not change from these values, to any great extent. This is a very common fault, since batteries fall in voltage sufficiently slowly to change the quality of the reproduction so gradually that it is not noticed till the change is serious.

For those more technically minded we will deal with transformer coupling in further detail.

If a voltage of 1 volt is applied across the ends of the primary winding of an AF5 Transformer, and the voltage

Fig. 75.

RATIO 1/1·75

TYPE AF2

FERRANTI
PATENT

PRIMARY

RATIO 1/7

SOLD UNDER LIMITED LICENCE

TYPE AF6

FERRANTI
PATENT

PRIMARY

TYPE AF5

FERRANTI
PATENT

PRIMARY

SOLD UNDER LIMITED LICENCE

TYPE AF 3

FERRANTI
PATENT

SOLD UNDER LIMITED LICENCE

TYPE AF4

FERRANTI
PATENT

Transformers.
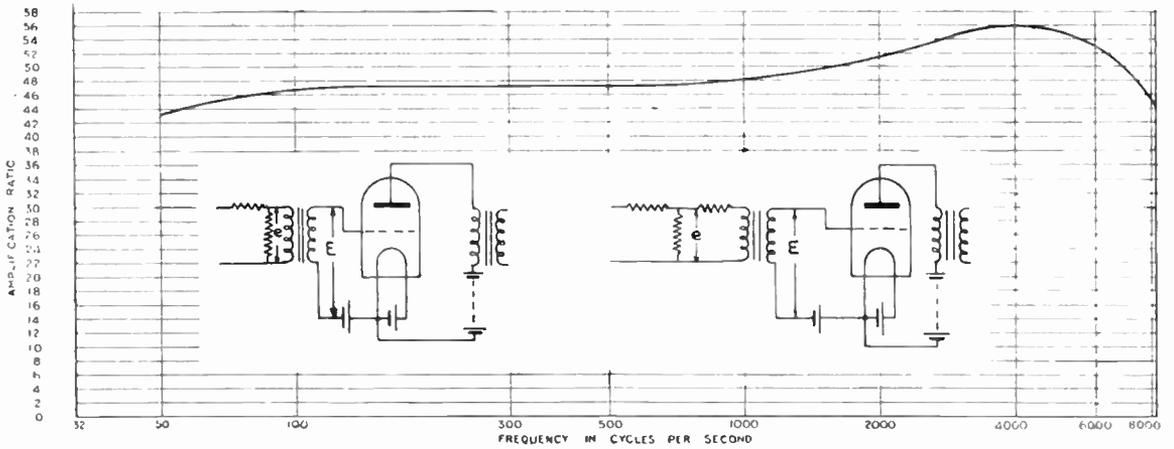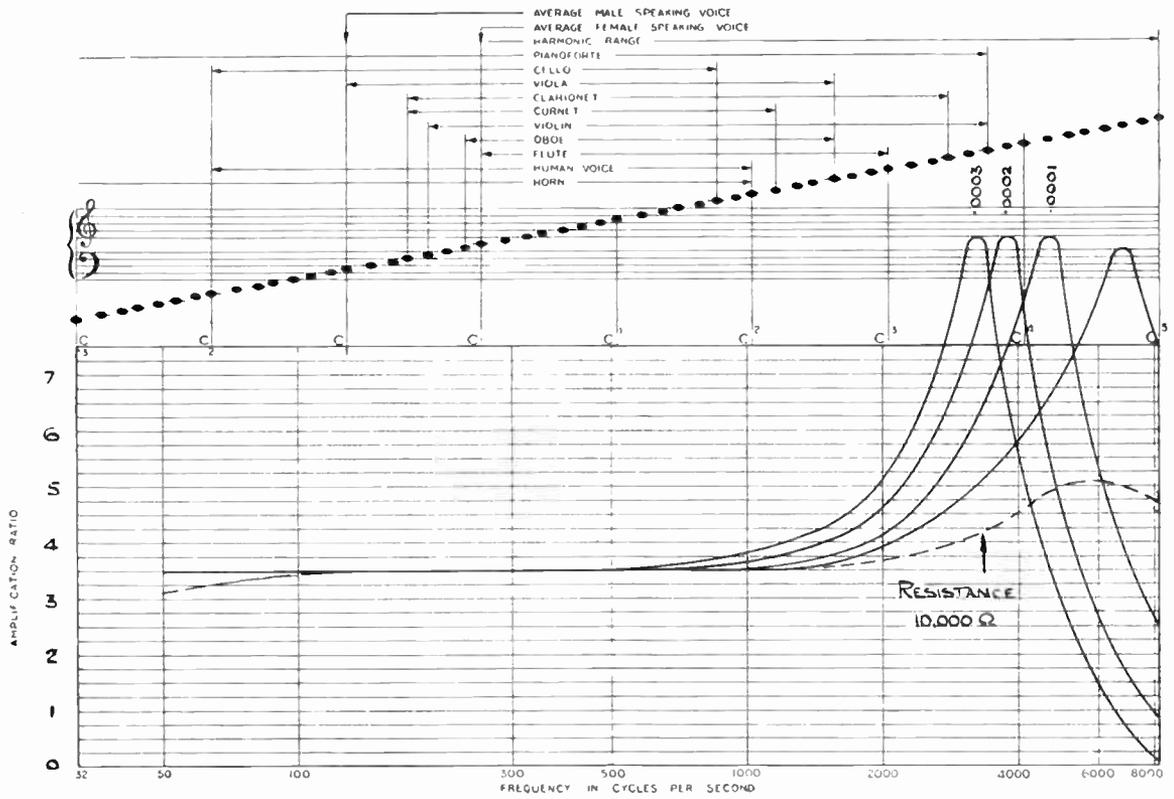
obtained across the secondary measured, we find that as the frequency of the signal increases, $E = 3\frac{1}{2}$ volts until at about 1,000 cycles, when E gradually increases till at 7,200 cycles it reaches the large value of $9\frac{1}{2}$, and then falls off again. See Figure 75.

A small condenser connected across the secondary merely moved the maximum point to a lower frequency, but a resistance of $\frac{1}{4}$ megohm connected across the secondary greatly reduced the peak.

If, however, a resistance of 10,000 ohms were connected in series with the primary, as in the circuit on the right, one obtains the dotted curve. It seems clear that the peaks obtained are due to a resonance phenomenon, and that the resistance of 10,000 ohms greatly damps it. Below is shown the usual amplification curve of an AF5 in the plate circuit of a valve. Its resemblance to the dotted curve indicates that the apparent resonance effect is of help in this amplification curve.

The historical analysis of intervalve transformers published by Mr. Dye, in *Experimental Wireless* for September, October, and November, of 1924, made the matter clear.

He showed that a transformer primary resolves itself into the circuit shown in Figure 76, and that it is made up of a leakage inductance L^1, in series with the primary inductance L, a shunting capacity C made up of all the internal capacities, and the capacity of the second valve stepped up into the primary winding, the whole shunted by a resistance S representing principally iron losses—that is, eddy current loss—and approximately independent of frequency. He gives complicated mathematical formulæ to obtain the results.

The actual values of L^1 and C under working conditions are not very accurately known, but for an AF5, L^1 is probably ·8 henry, C about ·00062 mfds., L 100 henries, and S 500,000 ohms. The ratio of the windings being $3\frac{1}{2}/1$.

If a voltage of a frequency of 50 cycles is applied to the grid of the valve, the effects of L^1, C, and S are negligible and the impedance in the plate circuit is $6 \cdot 28 \times 50 \times 100 = 31,400$ ohms inductive. The valve has an amplification factor of 14 and a resistance of 13,000 ohms.

Hence the total amplification is $\dfrac{\mu \times Z \times 3\frac{1}{2}}{R_p{}^2 + Z^2}$ since Z is purely inductive and the transformer ratio $= 3\frac{1}{2}$ the amplification $= \dfrac{14 \times 31,400}{\sqrt{13,000^2 + 31,400^2}} \times 3\frac{1}{2} = 45$ and the voltage at B would be nearly $90°$ out of phase with the applied signal at A.

If the frequency be increased gradually, the impedance of the plate circuit increases, becomes less inductive and more resistive, until we arrive at a frequency—in this case of 640 cycles—at which the circuit L C is resonant. The total impedance is now a maximum, purely resistive, and equal to S.

Now we get an amplification $= \dfrac{14 \times 3\frac{1}{2} \times 500,000}{\sqrt{(13,000 + 500,000)^2}} = 48$ and the voltage at B would be in phase with that at A.

A further increase of frequency makes the impedance of the plate circuit capacitative, and but for another factor it would get smaller and the amplification curve would fall.

This other factor is that L^1 and C form a series resonant circuit at 7,200 cycles. But for the resistance of the valve, the current through L^1 and C would be very large, causing the great rises shown in Figure 75. Though limited, the effect is clear on the amplification curve. At these higher frequencies the voltage at B is out of phase in the opposite direction to that at 50 cycles.

AVERAGE MALE SPEAKING VOICE
AVERAGE FEMALE SPEAKING VOICE
HARMONIC RANGE
PIANOFORTE
CELLO
VIOLA
CLARIONET
CORNET
VIOLIN
OBOE
FLUTE
HUMAN VOICE
HORN

AMPLIFICATION RATIO

60
58
56
54
52
50
48
46
44
42
40
38
36
34
32
30
28
26
24
22
20
18
16
14
12
10
8
6
4
2
0

I

III

32   50   100   300   500   1000   2000   4000 6000 8000

FREQUENCY IN CYCLES PER SECOND
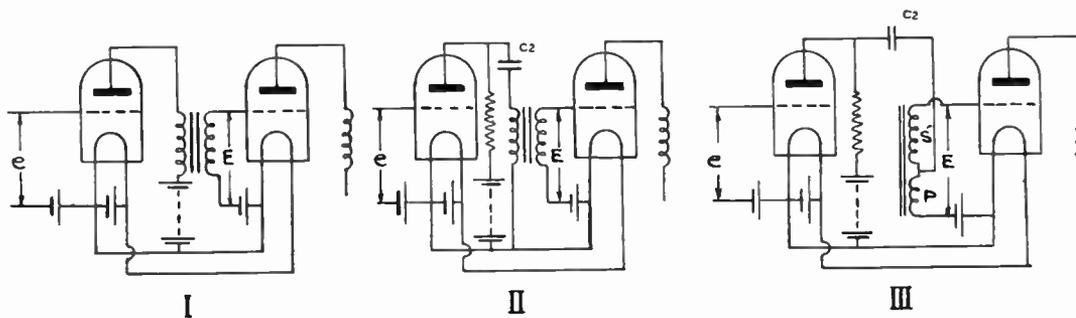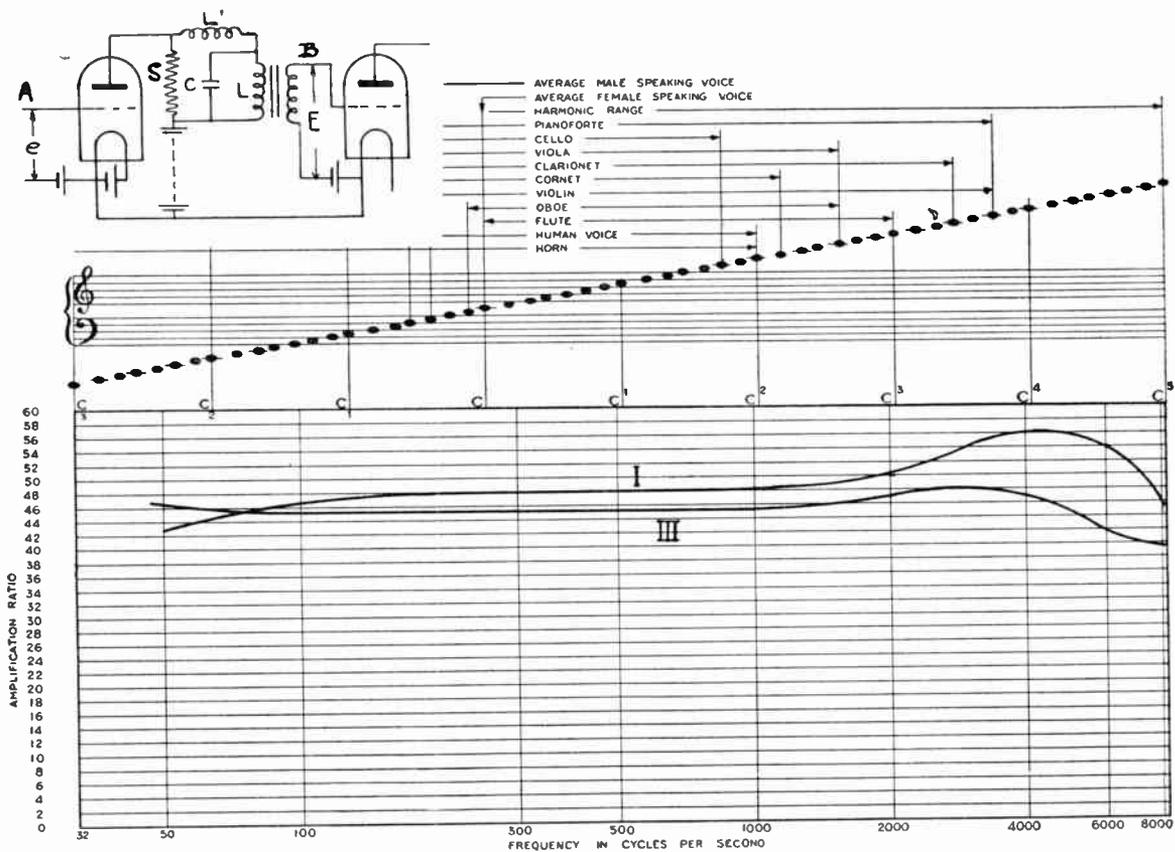
I   II   III

Fig. 76.

Combined Resistance and Transformer Coupling.

II and III in Figure 76 shows two forms of this.

A resistance in the plate circuit is coupled by a large condenser C2 to the transformer primary in II and to the transformer as an auto-transformer in III. The main primary induction forms a series resonant circuit with C2 of low frequency. When C2 = 1 mfd. this is 16 cycles. This lifts up the amplification at this end, as is evident in Curve III.

This form of circuit seems useful, but owing to the rather high value of resistance used to keep up the amplification, decoupling cannot be of the full amount, and trouble is often experienced with self-oscillation.

The effect of a Resistance across the primary winding of a Transformer whilst coupling Valves, and its use as an Audio Frequency Volume Control.

For some purposes, one stage of audio frequency amplification is insufficient and, on the other hand, two stages will always be too much in these days of powerful stations and radio frequency amplification. Further, because of the great difference of modulation by different stations, an audio frequency volume control is desirable as well as a radio frequency one.

The circuit diagram Figure 78 shows a variable resistance across the primary of the first AF5. Above are the amplification curves of the transformer, with various values of this resistance.
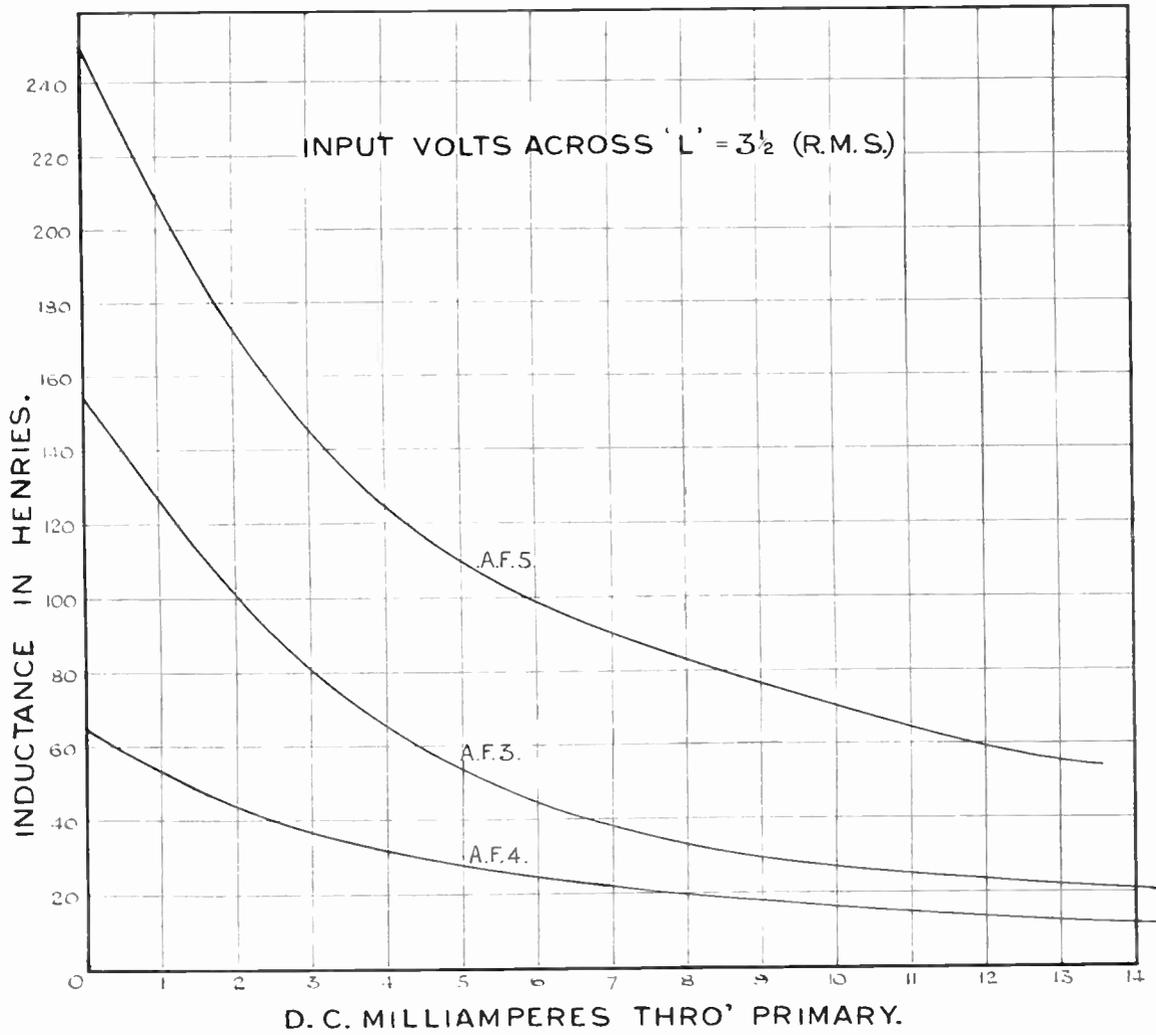
Fig. 77.

Inductance curves of AF3, 4 and 5 Transformers.

A.F.5. WITH D.E.L 610.

AVERAGE MALE SPEAKING VOICE
AVERAGE FEMALE SPEAKING VOICE
HARMONIC RANGE
PIANOFORTE
CELLO
VIOLA
CLARIONET
CORNET
VIOLIN
OBOE
FLUTE
HUMAN VOICE
HORN

SHUNT RESISTANCE 20,000 Ω

5,000 Ω

2,000 Ω

AMPLIFICATION RATIO

FREQUENCY IN CYCLES PER SECOND

THESE RESISTANCES (VALUES APPROX
100,000 Ω) PREVENT SELF OSCILLATION
IF THE VALVES ARE SLIGHTLY DEFECTIVE
OR SOFT.

Fig. 78.

When its value is 20,000 ohms the amplification is 33, instead of 48 when with no resistance, and the 50 cycle point is relatively higher.

With 5,000 ohms the amplification is 15, with 2,000 ohms only 7, and is absolutely flat until the rise above 2,000 cycles is reached.

These curves illustrate the effect of such a volume control.

The curves given for AF5 and AF3 Transformers with specified valves, are of the valves when used under purely amplifying conditions. The question arises as to whether the use of the transformer after the detector valve would produce the same result.

In the section dealing with grid leak rectification, it was shown that, properly carried out, this resulted in an audio frequency signal at the grid of the valve, which then acted relative to this, purely as an amplifier. However, in curve Fig. 63a in this section of measurements on the A.C. Mains Receiver, Model 32, we have the audio output volts for constant modulation of a fixed radio frequency signal with notes from 100 cycles to 6,000 cycles. This includes any losses due to rectification, transformer coupling, and the output transformer. It will be seen that it does not depart radically from our usual amplification curves.

In the section dealing with anode bend rectification, Figure 55 shows an amplifier fed by an anode bend rectifier and the audio output volts are given for a constant RF input modulated with notes from 50 cycles to 8,000 cycles.

Figure 56 gives the curve of the figures in Table II in that section, of the same receiver used under amplifying conditions.

A comparison of the two figures shows that the only effect of the rectification is to remove the rise at 4,000 cycles to 8,000 cycles, due to the second resonance point mentioned previously.

The amplification curves published with our transformers give an accurate indication of their performance under all conditions.

We will complete this section by setting out on the *measured* characteristic curves of two valves the instantaneous variations of plate current and plate voltage for large signal inputs to see if any appreciable distortion is produced for a pure sine wave of 100 cycles input to grid.

Figure 79 shows the *measured curves* of an MHL4 valve, and is in the well-known form of grid volts by plate current for various plate voltages. Figure 80 is in the more useful form, plate current, plate volts, for various grid voltages.

In both figures we have an AF5 or AF5c with a primary inductance of 84 henries, in its plate circuit, and the grid swing is from –1 to –7 volts, corresponding to 2·12 volts R.M.S. The curve obtained is an ellipse whose axis is inclined $2\frac{3}{4}°$ to the horizontal. It will be clearly seen that the ellipse for this 100 cycle note, cuts the grid bias lines wholly within the zone where they are parallel and equidistant.

Hence there is no distortion. The mean plate current is 7·8 milliamps, the mean bias –4 volts, and the mean H.T. voltage is 135. From the valve curves we calculate the valve resistance to be 8,000 ohms and amplification factor of 16·1 for this mean condition. Since the primary inductance is over 84 henries, the frequency-amplification curve will also be nearly ideal. An inspection of the curves shows that the grid swing could be from –1 to –8 volts, a mean of $-4\frac{1}{2}$ volts without any distortion. The plate volts would be 142 and current 7·8 milliamps. Such a swing as this would overload two L.S. 6a valves in push pull with 400 volts on the plate.
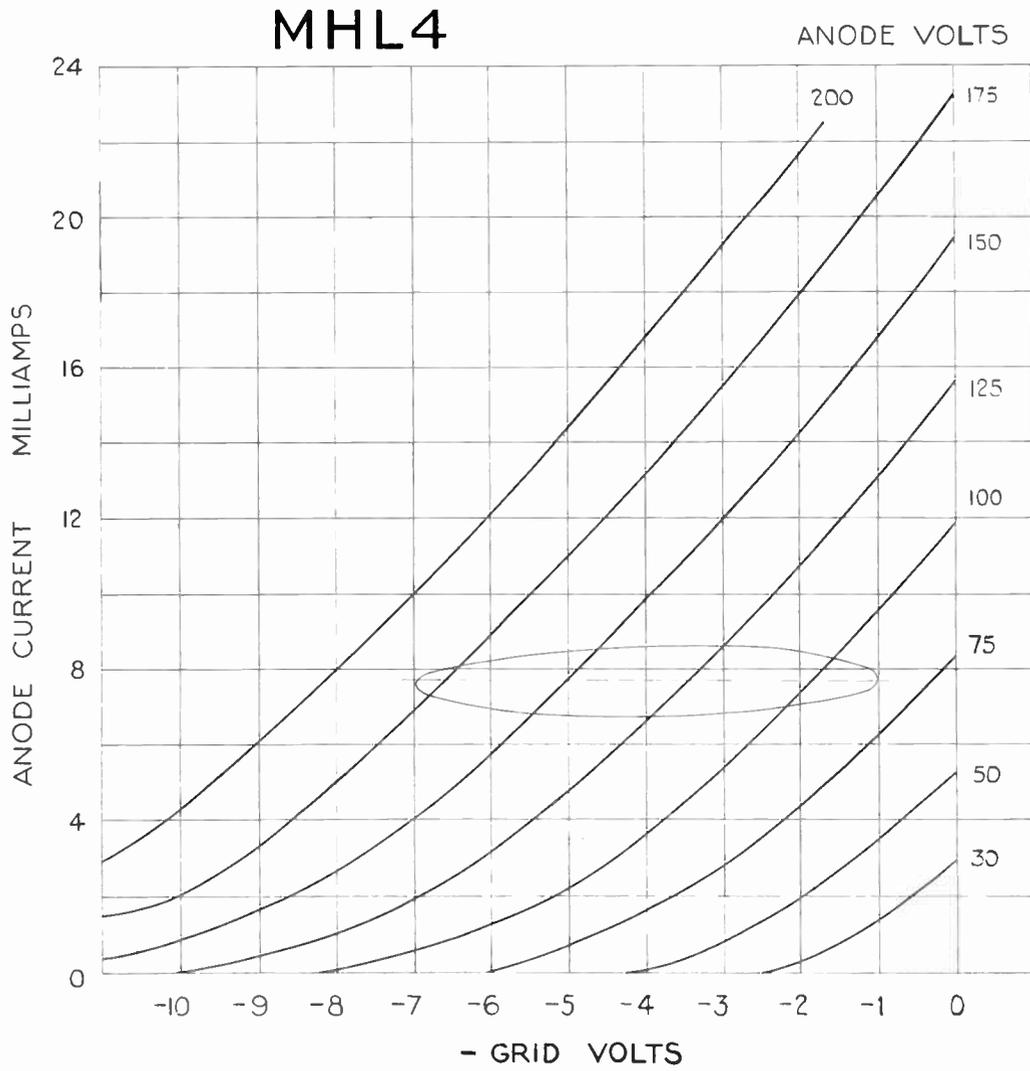
Fig. 79.
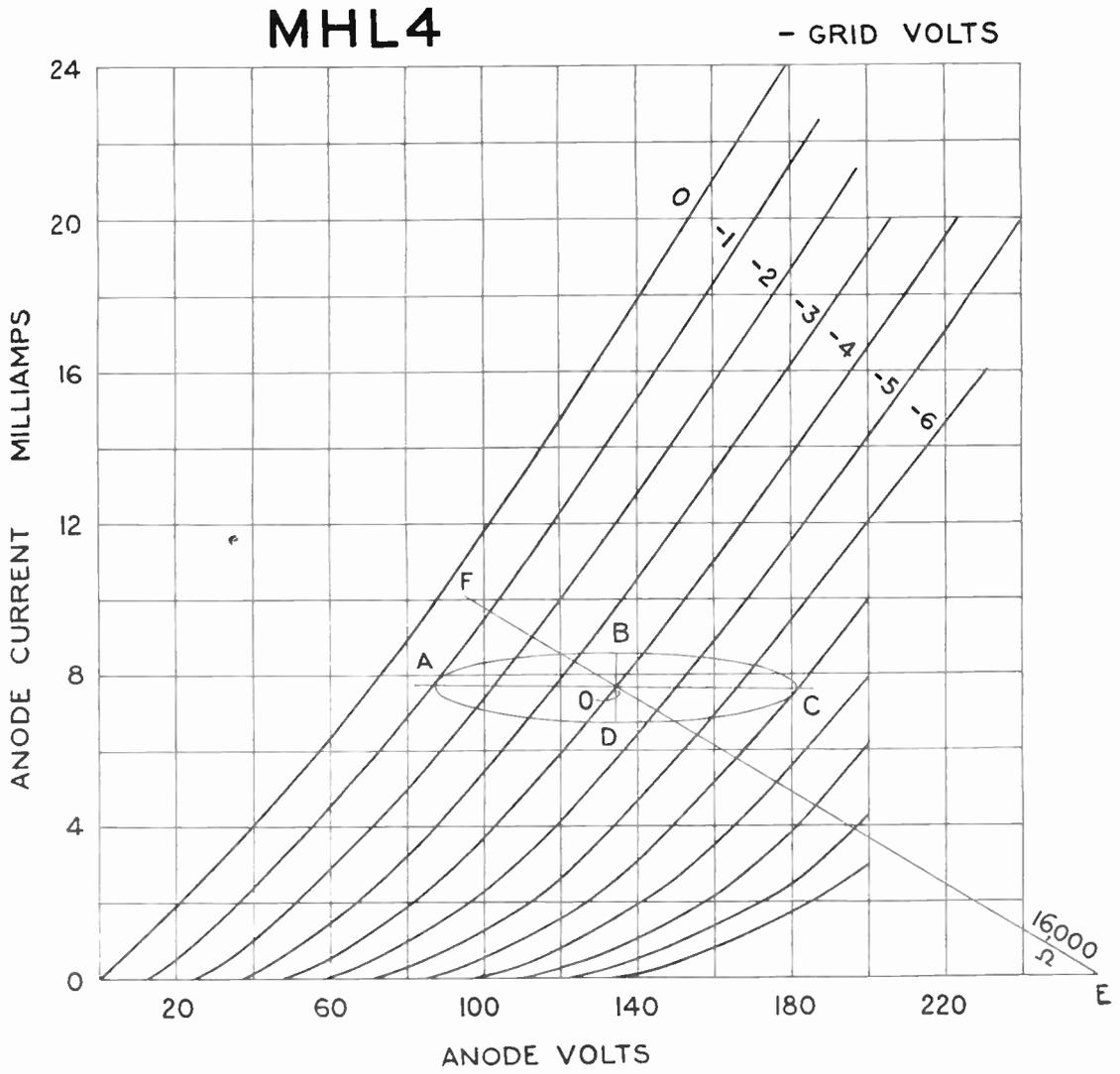
Fig. 80.

Application to Power Grid Rectification.

If this MHL4 valve was used for rectification with a grid leak of $\frac{1}{2}$ megohm and a condenser of ·00015 mfds, then from the section dealing with this we find an R.F. signal of $3\frac{1}{2}$ volts modulated 80% causes a grid swing from −1 to −7 volts and, as just explained, there would be no distortion.

There is one thing to be particularly noted : that in this case, before the signal came in, the mean grid potential would be −·52 volts, and when receiving the signal would be −4 volts. If the H.T. supply were obtained direct from a battery of 135 volts, then before reception the plate current would be 15·7 milliamps, and after would be 7·8 milliamps, not a very desirable result, since the 15·7 milliamps is rather a large current for the winding of the AF5. But usually a higher voltage is used and a decoupling resistance with the necessary by-pass condenser fitted. This greatly reduces the variation. Thus, in a Ferranti amplifier, the constant potential is 260 volts and the decoupling resistance is 16,000 ohms, line EOF. Hence, before reception, the current is 10 milliamps, point F, and after, 7·8 milliamps, point O.

Figure 81 gives similarly measured curves for an MH4 valve with an AF5 or AF5c Transformer in the plate. The large ellipse is for a 100 cycle note with a grid swing of −·75 to −3·59 volts. A small distortion is just beginning to appear since the distance between −3 and −4 is not quite equal to that between −2 and −3. Hence, 1 volt R.M.S., equivalent to the swing −·75 to −3·59, mean −2·17 volts, is the largest signal that can be applied to the grid. This signal, however, would just overload two L.S. 6a valves in push pull.

If the valve be used as a detector with a grid leak of $\frac{1}{2}$ megohm and a condenser of ·00015 mfds, this corresponds to a signal of 1·68 volts R.M.S. at 80% modulation. If, therefore, one had enough radio frequency amplification to ensure an RF signal of 1·68 volts at the grid of the detector valve, one could use an MH4 valve followed by an AF5c connected to two LS6a valves. Here again, because of the
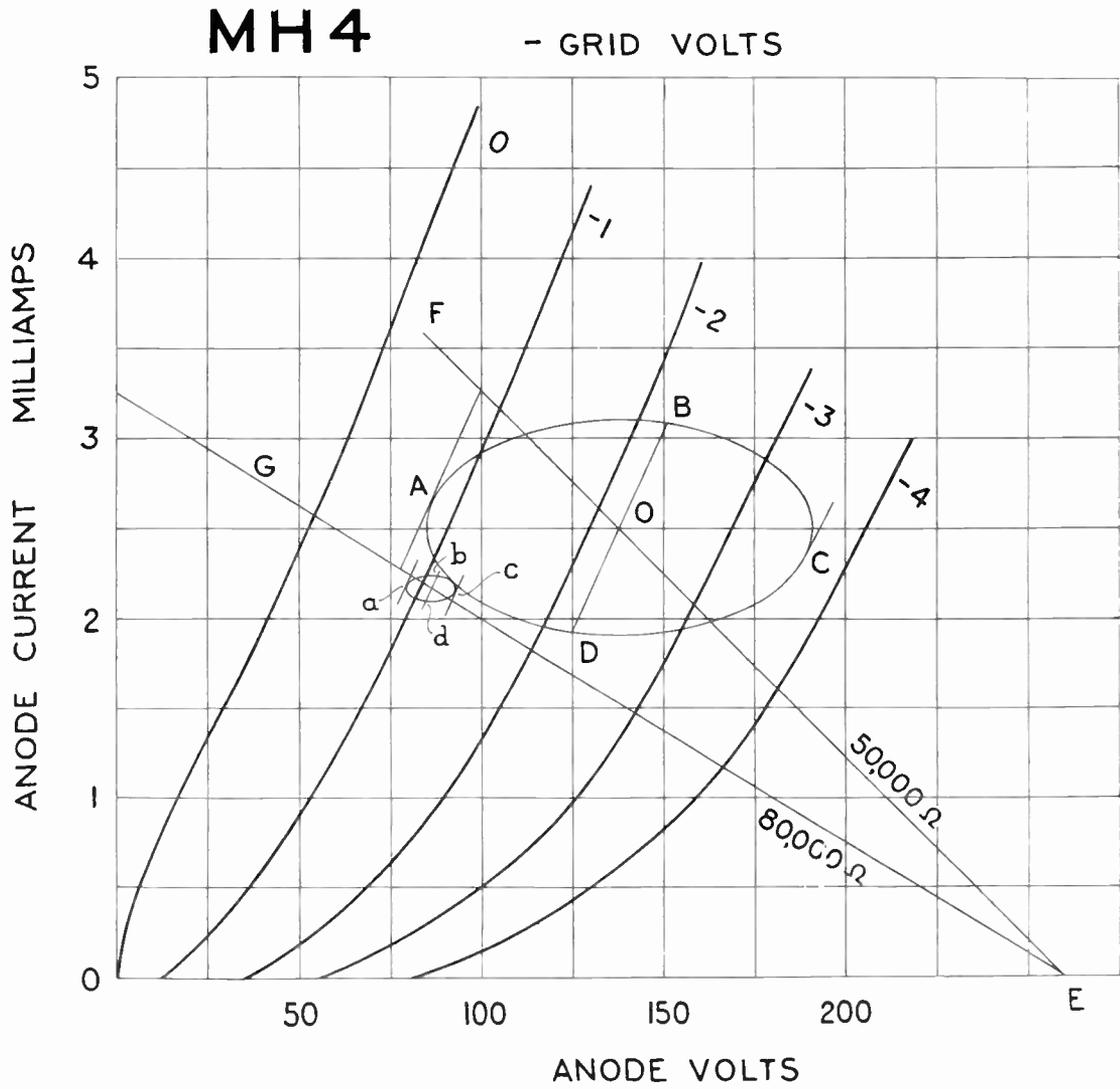
Fig. 81.

high inductance of the AF5c primary for the plate current of 2·5 milliamps, namely, 160 henries in relation to the valve resistance, 23,600 ohms, the frequency amplification curve is almost ideal. With a decoupling resistance of 50,000 ohms from a constant voltage of 260 the working plate voltage is 137·5, the plate current 2·5 milliamps, and the mean grid voltage −2·17. Before receiving the signal, the grid voltage is −·75, the plate voltage 98 volts, and the plate current 3·25 milliamps.

The small ellipse represents the plate volts—plate current, variations for an R.F. signal input to the grid of the MH4 as detector equal to ·325 volts, 80% modulated with a 100 cycle note. Before reception, the grid is at −·75 volts, with plate volts 77, plate current 2·25 milliamps, and decoupling resistance of 80,000 ohms. This is the condition which results at the detector valve of the A.C. Mains Receiver, Model 32. When the signal of ·325 volts which will normally result in this set, comes in, the plate volts increase to 86 volts, the mean grid volts to −1·05, and the mean plate current becomes 2·15 milliamps. This small ellipse is well within the zone where the grid bias lines are straight, parallel, and equally spaced, and hence there is no distortion due to the amplification.

Mathematical Section.   Calculations to get the plate current—plate volts, variations when the grid input is sinusoidal.
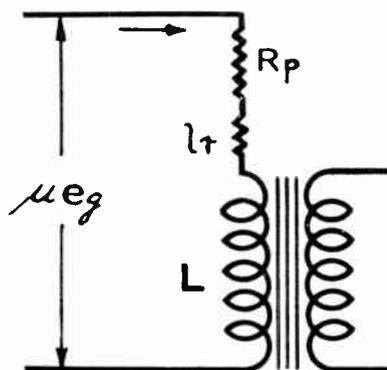
Fig. 82.

The valve and transformer become equivalent to the above figure.

Let $Rp + R = p$

For instantaneous currents.

$$\mu E \sin \omega t = ip + L\frac{di}{dt} \quad \dots \quad \dots \quad \dots \quad \dots \quad (1)$$

Solution is : $i = A \sin (\omega t + \propto) \quad \dots \quad \dots \quad (2)$

where $\tan \propto = -\dfrac{\omega L}{p}$ and $A = \dfrac{\mu E}{\sqrt{\omega^2 L^2 + p^2}} \quad \dots \quad (3)$

$$\therefore i = \frac{\mu E}{\sqrt{p^2 + \omega^2 L^2}} \sin \left[\omega t - \tan^{-1} \frac{\omega L}{p} \right]$$

The voltage across the inductance

$$p = Ri + L\frac{di}{dt}$$

$$= \frac{\mu E R}{\sqrt{\omega^2 L^2 + p^2}} \sin \left[\omega t - \tan^{-1} \frac{\omega L}{p} \right] + \frac{\omega L \mu E}{\sqrt{\omega^2 L^2 + p^2}} \cos \left[\omega t - \tan^{-1} \frac{\omega L}{p} \right]$$

The relation between i and p is clearly an ellipse. The slope of the major axis is $\dfrac{R}{\omega L}$ radians.

Correction for the loss of higher notes due to selectivity in the Audio Frequency side of a Receiver.

The science of boosting up attenuated frequencies in telephone systems and the like, is an old and important one, and would require much space to deal with it. Forming as it does a most important part of the B.B.C. land line transmission. Without these " correctors " the quality of the transmissions would be intolerable and every microphone amplifier contains these.

In view of the unjustified claims made in recent patent specifications and applications, the history of this science as applied to radio receivers will be briefly related. The radio Press towards the end of 1925 dealt with some of the devices then known for boosting up the high notes at will in the audio side of a receiver. In November, 1925, at a demonstration at the Institute of Electrical Engineers,

Dr. McLachlan showed the effect of boosting up or reducing the higher notes by one of these devices on the first English moving coil speaker.

In December, 1926, the writer received from Ralph S. Hayes, an engineer of the Bell Telephone Company, a copy of a memorandum which was distributed to thirty Engineers of that Company, as a matter of common interest, dealing with an equalising audio stage for Radio Receivers to correct for :

(a) Loss of higher tones due to Radio frequency tuning.—

(b) Loss of lower tones due to audio frequency attenuation.—

At a later date, Hayes applied for an American patent, but the sending of a copy to England and its exhibition to a number of English Engineers before the date of the American Application would form publication and invalidate any English patent granted.
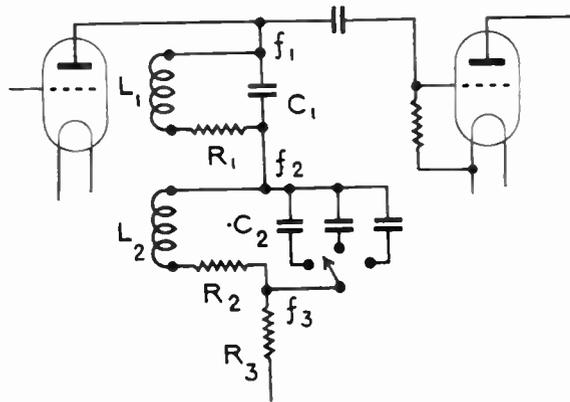
Fig. 83.

L_1 = ·25 henries.	L_2 = 82 henries.
C_1 = ·004 mfd.	C_2 = ·2 mfd. ... f_2 = 39 f_3 = 710.
R_1 = 20 ohms.	·3 mfd. ... = 32 = 580.
R_2 = 630 ohms.	·4 mfd. ... = 28 = 500.
R_3 = 2,100 ohms.	f_1 = 5,000.
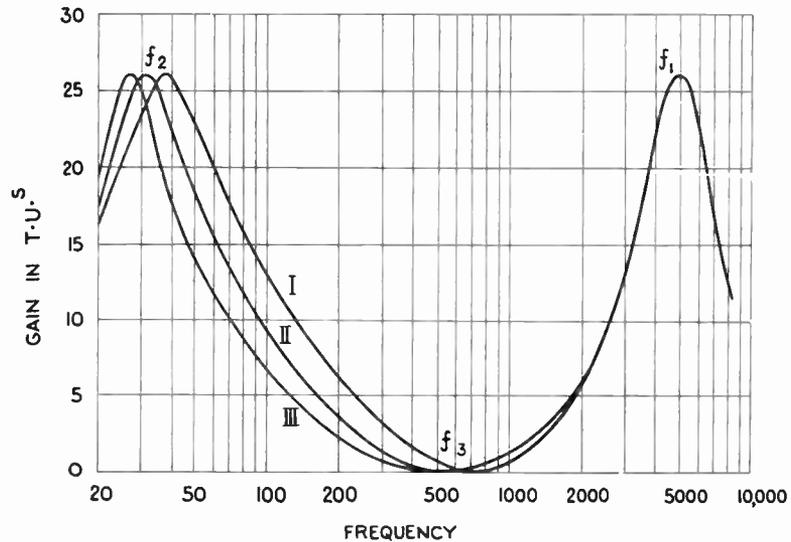
Fig. 84.

Curve I. $C_2 = \cdot 2$ mfd.
Curve II. $C_2 = \cdot 3$ mfd.
Curve III. $C_2 = \cdot 4$ mfd.

Figure 83 shows one of his arrangements, where two tuned circuits, one resonant to a very low frequency and the other to about 5,000 cycles, are connected in series with a resistance R3 in the plate circuit of a valve, coupled by capacity to another. The figure gives the values of the components. Figure 84 gives the curve of amplification. If one leaves out the low frequency series circuit one obtains the corrector for the high frequency loss only. By varying the value of the series resistance R3 in relation to the resistance of the tuned circuit at resonance, we can vary the amount of correction.

Since the summer of 1927 we have applied a similar corrector system to transformer coupled audio frequency amplifiers and fitted these to receivers, to correct for the loss of the high notes.

High Tension Unit.
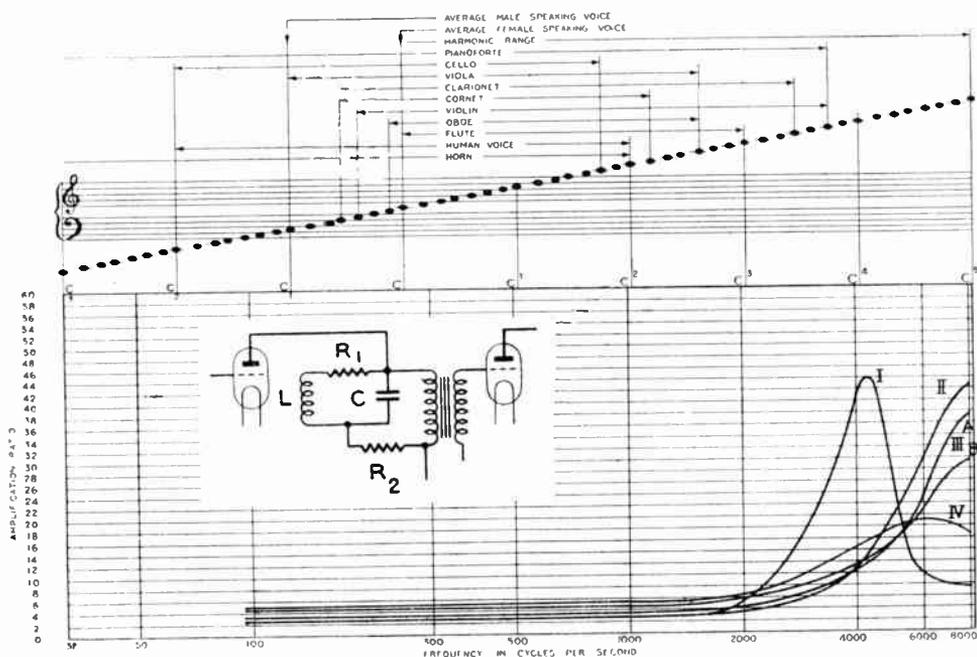
Fig. 85.

			Ohms. R₁		Ohms. R₂		Mfd. C		Microhenries. L
I	...	...		...	1,012	...	·022	...	51,000
II	...	...	--	...	1,012	...	·007	...	51,000
IIIA	...	...	100	...	1,012	...	·004	...	51,000
IIIB	...	...	700	...	1,012	...	·004	...	51,000
IV	...	...	1,000	...	1,012	...	·010	...	51,000

Here the tuned circuit in series with a variable resistance R2 is connected across the primary of an AF5. Usually a switch is fitted across the tuned circuit so that it can be shorted out when the corrector is not needed, leaving only the variable resistance. The values of R, L, C, and R2 can be adjusted to suit the circumstances.

The curves of Figure 85 shows the measurements obtained when the combination follows a D.E.610 valve, taken in February, 1928. The inserted figures corresponding to the

K

curves give the values of L, C, R, and R2. The increase of the higher notes relative to the lower is very great in these cases.

It will be noticed that the absolute amplification of the lower portion of the musical scale is small, so that this method of correction can only be carried out when there are at least two stages of audio frequency amplification.

THE POWER STAGE.

Having amplified the signal to the desired amount, we have a voltage representing fairly accurately the original music in the studio. But if a loud speaker is to produce sound we must convert the *voltage* into *electric power,* and this is the function of the last stage.

Power and Super Power Valves.

For this purpose we use a different valve to that used for amplification, called, according to its size, a *Power Valve* or a *Super Power Valve,* though the larger sizes have no such simple name. We shall deal here with only three electrode valves. Such a valve amplifies but little, has a low resistance, but a large plate current, the rise and fall of which, with the variations of grid voltage, through the impedance of the speaker, is the power which is converted into sound.

It is unfortunate that the conversion of electric power into sound in a speaker is an extremely inefficient process. In the best of speakers, only $\frac{1}{50}$ of the energy input is converted as sound. Special speakers have been made for which an efficiency of 25% has been claimed, but such are too cumbersome and expensive for general use. Listening is also a very wasteful process, for, of the sound energy scattered through space from the speaker, only the smallest fraction is caught by the two square inches of ear orifice.

Hence, to obtain good reproduction from a speaker, the power valve must be large and the high tension supply capable of giving sufficient current. Power valves similar to other apparatus having a rating fixed by the manufacturers which gives a good indication of what one can obtain from them. The maximum D.C. watts which can be dissipated from the plate of the valve without endangering it, is specified.

The tables at the end give full details, but here we will indicate the classes into which they fall.

(A) Small power valves made for a maximum plate voltage of 150 and taking 8 milliamps mean plate current, a 1·2 watt valve $\left[150 \times \dfrac{8}{1,000} = 1 \cdot 2 \right]$ Type P215.

(B) Medium power valves made also for a maximum plate voltage of 150, but taking 16 milliamps plate current, a 2·4 watt valve. Type P240.

(C) Large power valves of 6 watts rating, of which there are two types. The first represented by a P625A valve with a maximum plate voltage of 200, and the plate milliamps = 30. The second by a P625 valve. Maximum plate voltage 250, plate current : 24 milliamps.

(D) 10 watt power valve ; of the PX4 type. 200 plate volts 50 milliamps.

(E) 15 watt class, represented by L.S.5A valves.

(F) 30 watt class, represented by L.S.6A valves.

Where the high tension can only be obtained from dry batteries, one is practically limited to Class A, unless the batteries can be renewed very frequently, in which case Class B is possible. Since only $\frac{1}{8}$ of the rated D.C. Power can be obtained as A.C. power (i.e. music or signal power)

without serious distortion, the use of Class A limits us to $\frac{1\cdot2}{8}$ watts or 150 milliwatts. This is not sufficient, and listeners often complain of a rattle in the speaker. This is caused by the harmonics introduced by overloading the valve in an attempt to get more volume than is warranted by the available power of 150 milliwatts. Until a loud speaker is made with five times the efficiency of the present type, users of Class A valves cannot be helped.

Class C and above are, because of the large power they take, practically limited to those who have electric mains. Such valves, properly used with suitable speakers, provide music which is similar to the original at a reasonable mean volume level, without " blasting " or " rattling " on loud peak passages. There is a doubt amongst the general public as to which power valve to use ; many people having electric mains purchase mains apparatus of so small capacity that they still have to use Class A, or at the best Class B, power valves ; they have the means to obtain real music, but do not utilise it. It is imagined that the larger power valves are only for those who want a large amount of noise, " noise fiends " who must make their speakers " bellow." These " noise fiends " misuse the larger power valves just as much as the dry battery man is unwittingly compelled to misuse his valve of Class A. The general sound level of music or speech does not require much energy ; the loud passages, however, may represent fifty times or more the energy of the mean. The use of a large power valve of Class C does not give a greater mean sound level, but a sufficient reserve the reproduce the loud passages without overloading.

Figure 86 shows the simplest method of connecting a loud speaker to a power valve. One terminal is connected to the plate of the valve and the other to the positive of the high tension battery. This method has been in very common use in the past, but must only be used where
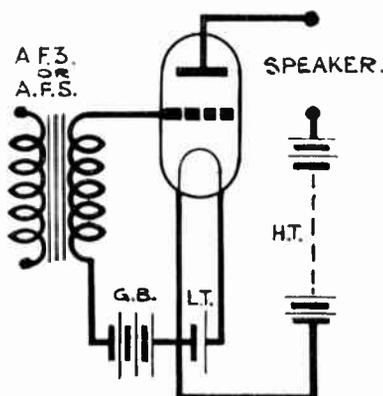
Fig. 86.

the H.T. battery is of small power and of a voltage not exceeding 150. It must not be used where higher voltages are utilised, or where the set is connected in any way to the electric light mains, unless the speaker is combined with the receiver in a cabinet so that it cannot be touched, on account of the danger should contact be made with either speaker terminals. The alternative methods will be shown later.

It is necessary, if we are to obtain the maximum power with the minimum distortion, that the impedance of the speaker should be twice the resistance of the valve. A proof of this is given in a Paper by Green in *Experimental Wireless*, July and August, 1926. The type of small power valve usually used for such a direct connection has a resistance of 5,000 ohms, hence the impedance of the speaker should be about 10,000 ohms. It is clear, therefore, that only *high resistance speakers* should be so connected.

There is a further disadvantage of such a direct connection. The speaker generally has a D.C. resistance of 1,000 to 2,000 ohms, and the steady plate current through this prevents the full battery voltage from being available at the plate with some loss in power of the valve.
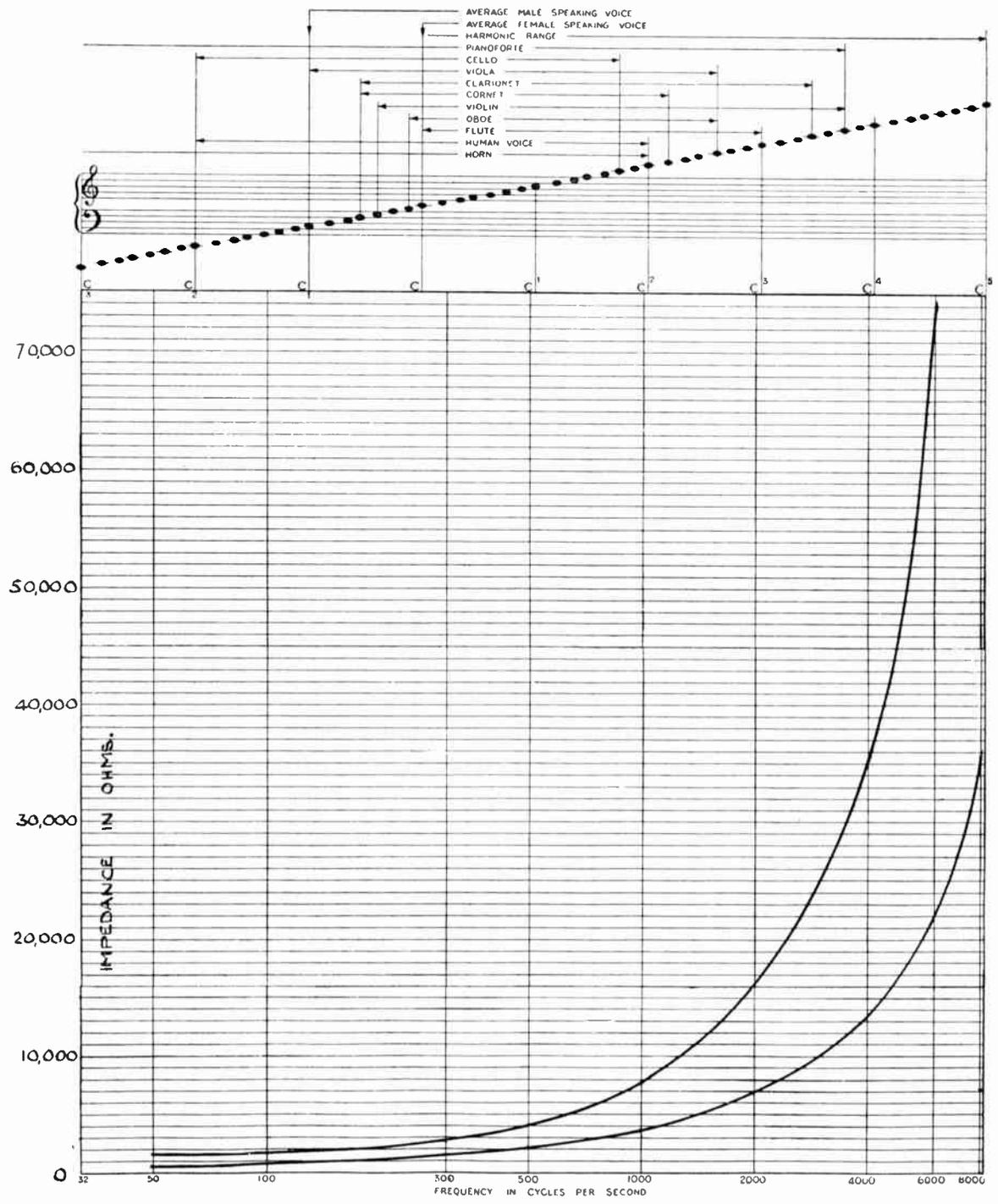
Fig. 87.

Impedance Curves of two Speakers.

Figure 87 gives the curves connecting the impedance of two good high resistance speakers with the frequency of the applied signal. It will be seen that the impedance for a good portion of the musical scale differs very greatly from the most desirable value given above, so that its use is only a rough compromise.
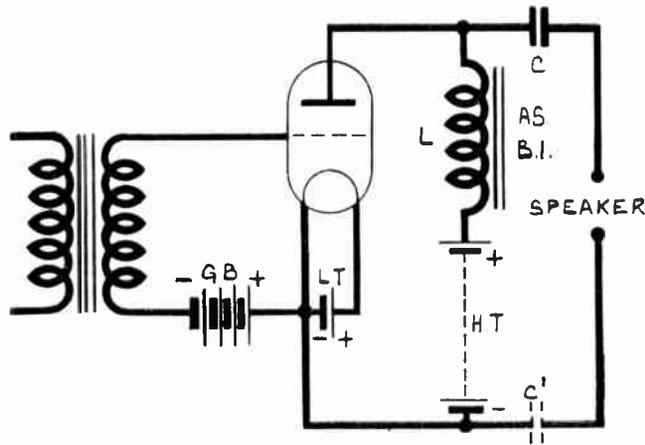
Fig. 88.

Choke capacity.

Figure 88 shows another method of connecting speakers which avoids to a great extent the dangers of shock when contact is made with the speaker terminals ; it is, therefore, permitted when the receiver is connected to the mains. It is known as *choke capacity* connection. It requires care in handling, however, for if the speaker becomes disconnected at one end, whilst the battery or mains are still on, condenser C becomes charged to the full voltage, which may give a severe shock if one touches the lead from it. If the set is connected to D.C. mains, the regulations require a condenser C, which is shown dotted, in the other speaker lead.
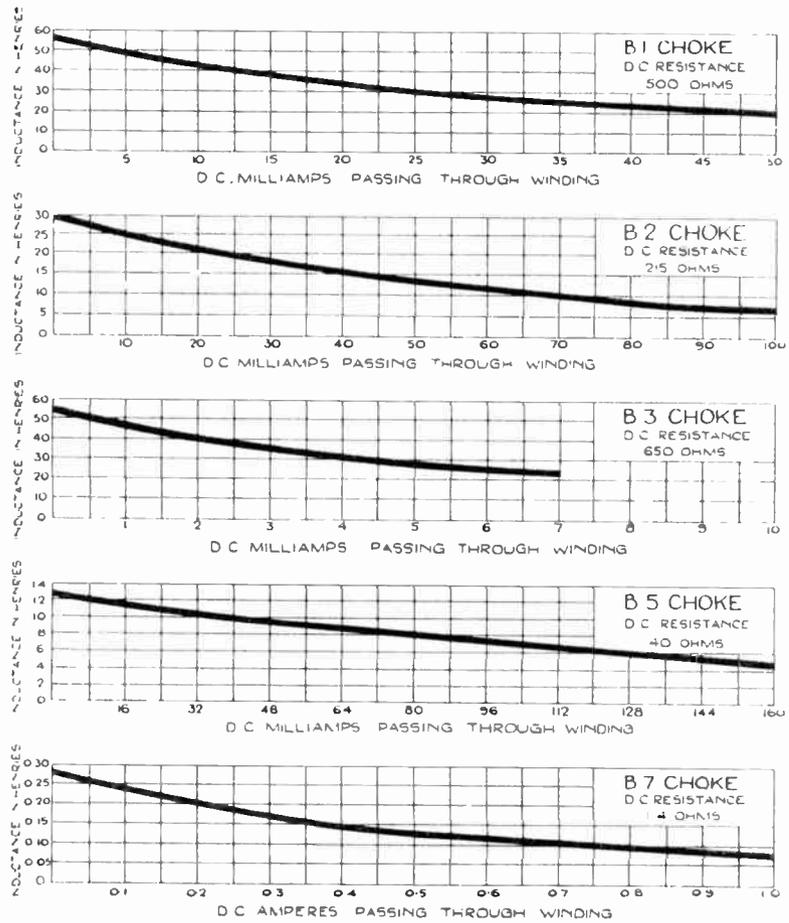
Fig. 88a.

Inductance of Ferranti Chokes.

The condenser C must be large—at least 2 or 4 microfarads. If both C and C' are used, each *must* be 4 mfd. The choke L must be of sufficient inductance when passing the full D.C. plate current ; B_1 is 32 henries when passing 24 milliamps. Electrically, the above connection is the same as the previous

one, with the advantage that the choke L can have less D.C. resistance than the speaker winding, hence there is less loss of the battery voltage. If the inductance L is sufficiently great, only the D.C. plate current passes through it, the signal current passing through the condenser and speaker. This is the one advantage of this connection, since the signal is not passing through the battery or eliminator, there is much less chance of interaction between the stages of the receiver, and therefore less chance of self oscillation or motor-boating. This is not, however, as ideal as it appears—some portion of the signal does pass through the choke. All the difficulties due to the varying impedance of the speaker with frequency apply to this case.
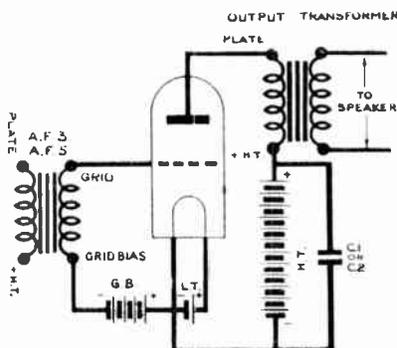
Fig. 89.

Output Transformers.

Figure 89 shows the third method of connecting a speaker to a power valve by the use of what is known as an output transformer. The primary winding is connected in the plate circuit between the plate and the H.T. supply, and the speaker across the terminals of the secondary winding. Here, the speaker is completely isolated from the H.T. supply, so this is certainly the safest connection, and in receivers connected to mains this method should always be used where possible.
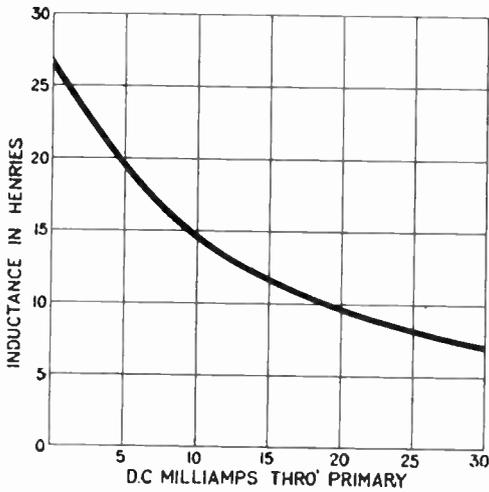
Fig. 89a.

Inductance Curve of Primary of OPM1 Output Transformer. D.C. Resistance of Primary 180 ohms.
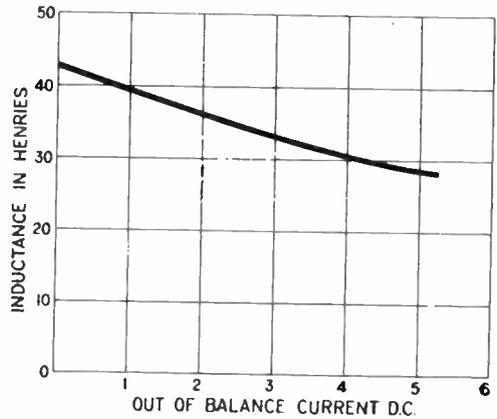
Fig. 89b.

Inductance Curve of Primary of OPM1C Output Transformer. Total D.C. Resistance of two half Primaries, 590 ohms.
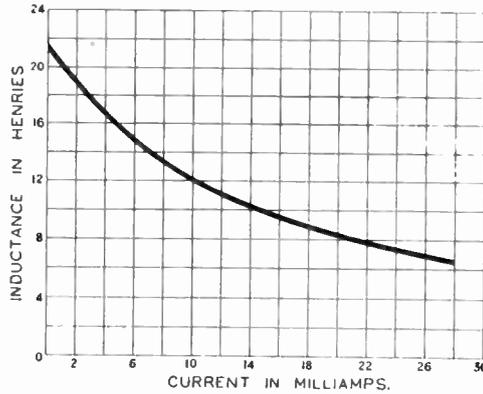
Fig. 89c.

Inductance Curve of Primary of Output Transformer, as used on A.C. 32 Receiver.

Leakage Inductance : Ratio 15 : 1 Shorted = ·0315 Henry.
Ratio 1 : 1 Shorted = ·0394 Henry.
D.C. Resistance of Primary = 384 ohms.
D.C. Resistance of Secondary = 1·3 ohms. Ratio 15 : 1.
377 ohms. Ratio 1 : 1.

When the signal voltage is applied to the grid of the valve, signal currents pass through the primary winding of the transformer. The varying magnetic field thus created in the iron core generates similar currents in the secondary winding proportional to the impedance of the speaker.

These, however, create a magnetic field in opposition to that due to the primary which cancels a very large portion of its inductance, leaving in the plate circuit an impedance equal to that of the speaker, where the ratio of the two windings is 1. We have, in effect, in this indirect method, obtained the same result electrically as if the speaker were connected directly in the plate circuit. This is only strictly true if a good transformer is used whose windings have sufficient initial inductance and small D.C. resistance, and if the speaker were all resistance; the fact that it has inductance complicates the matter, but does not affect the general principle. Formulæ and measurements will be given showing the defects that can exist in output transformers.

If the transformer has a different number of turns on the primary to those on the secondary, then we define the ratio of the transformer as

$$\frac{\text{number of primary turns}}{\text{number of secondary turns}} = N$$

Output Trans-former enables speaker to be matched to valve impedance.

In such a case the speaker impedance when transferred to the plate circuit is increased N^2 times. This gives one the chance of adapting a low impedance speaker to the high resistance of a valve.

EXAMPLE. To connect a 20 ohms impedance moving coil speaker to the plate of a P625 valve whose resistance is 2,400 ohms. Let N be the ratio of the transformer wanted.

Then

$$N^2 = \frac{\text{twice resistance of valve}}{\text{Impedance of speaker}} = \frac{4,800}{20} = 240$$

hence $N = 15\cdot5/1$. Using a transformer of this ratio, we have adapted or matched the speaker to the valve.

We have indicated three methods of connecting a speaker

to a power valve, and explained that the increase and decrease of grid voltage due to the signal, caused a similar change in the plate current of valve, which, passing through the speaker, provided the power to operate it. We have pointed out that the size of the power valve, and the plate voltage applied to it, often determines the quality of the music obtained. There still remains another variable, the grid bias applied to grid of the valve. The grid must not be allowed to become positive or too negative by even the strongest signal applied.

Should this happen, " blasting " or " rattling " results in the speaker during loud passages of music. Hence a suitable initial grid voltage or grid bias must be used. The Tables at the end give the correct bias for various power valves for several plate voltages.

Detailed explanation of actions in Power Stage.

For those interested, the matter will be dealt with in more detail with the aid of valve characteristics, more particularly with reference to the second and third method of connection.

Before the following can be properly appreciated or utilised, it must be thoroughly understood that when the signal is applied to the grid of the valve the voltage of the plate rises above and falls below that of the supply battery by an amount depending on the impedance of the speaker. The fall can immediately be understood, but it is not evident that the plate can be at a higher voltage than its battery. This arises from the change of rate of flow of electrons. The following explanations will greatly help the reader to visualise the action.

Both the circuits in Figures 90 and 91 reduce to that of Figure 92, where L (DG) is supposed to be a large inductance, such that rapid current changes cannot take place through it, but that it passes the steady D.C. current. All rapid changes of current must pass through EF, whose A.C. resistance R represents the impedance of the speaker; there is some imaginary means H, which, though offering no impedance
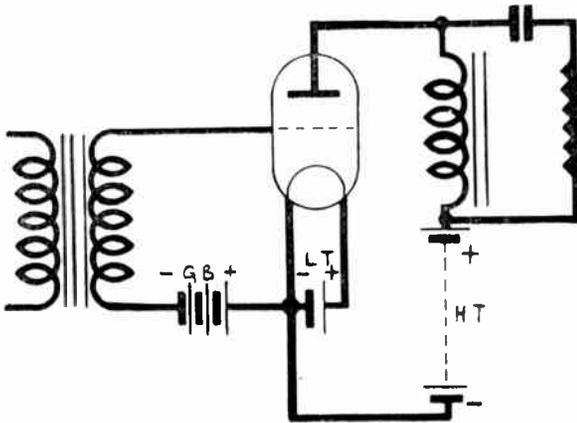
Fig. 90.
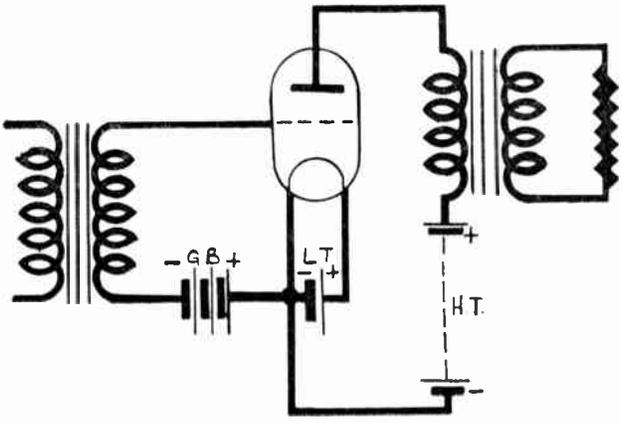
Fig. 91.
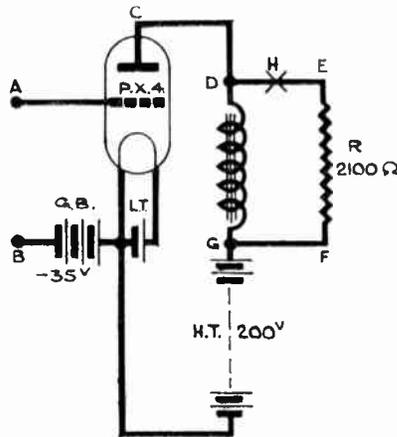
Fig. 92.

to the rapid changes, prevents the passage of steady D.C.
current.   A large condenser would do this, but it would
complicate the explanations unnecessarily because of the
change of voltage phase from side to side of the condenser.

We shall talk of the rate of flow of electrons, rather than current, and it must be continually kept in mind, *that electrons can only flow to a point of higher voltage than the point from which they come.*

Steady state.　Let us consider the steady state of the valve in Figure 92.

When the voltage of A equals the voltage of B—that is, −35 volts—then the electrons from the filament to the plate flow at a steady rate through L, represented by the current 50 milliamps. Nothing will pass through R because of H, and the voltage C will be 200 volts, the same as that of G, the positive end of the battery.

Grid suddenly made positive.　Now imagine that A be suddenly made 35 volts more than B, the rate of flow of electrons from filament to plate, and from G to the filament, must increase, and in this case be represented by 87·8 milliamps. But the rate of flow of electrons through L cannot change from 50 milliamps in the extremely short time, hence those represented by 87·8 − 50 = 37·8 milliamps, must flow through DEFG to the battery. Hence, the voltage of F must be higher than that of D, and therefore of C the plate, by the amount $\frac{37·8}{1,000} \times 2,100 = 79·4$ volts ; the plate volts at this instant must be 120·6.

Grid suddenly made negative.　Now let A be made suddenly 35 volts less than B. The rate of electron flow from filament to plate and from G to filament, must be less than 50, and in this case is represented by 17·5 milliamps. Again the rate of flow through L is 50 milliamps. The difference, 50 − 17·5 = 32·5 milliamps must therefore flow from GF through R to ED down through L to keep this at 50 milliamps ; the 50 milliamps being composed of 17·5 milliamps from the plate and 32·5 milliamps from G, the bottom end of the choke, back

through R to D. Since the electrons flow from F to E, E and the plate of the valve must be $\frac{32 \cdot 5}{1,000} \times 2,100 = 68 \cdot 3$ volts higher than the battery volts.

The action here is shown as occurring in jumps, but actually, of course, the changing process is a continuous one.

A more technical and brief explanation than the above is that if an electron flow is accelerated, the point from which it comes must be of lower potential than that to which it flows. Alternatively, if the flow of electrons is decelerated, as it must be when the grid is more negative than the mean, from a point, *this point,* the plate, must be of a higher potential than the point to which it flows, the positive of the battery. It is deceleration that causes the increase of potential.

As the grid voltage at A rises above, or falls below, that of the grid bias connection at B, by virtue of the impressed signal, the voltage at the plate falls below or rises above, that of the battery or energy supply and the consequent variation of current flow is that which actuates the speaker.

Valve curves. Figure 93 shows curves of plate current grid volts. They are for a 10 watt valve, type PX4. If the plate volts are 200 and there is no impedance in the plate circuit, the plate being connected direct to the battery, when we vary the grid volts up and down, the curve ABC gives the plate current for every voltage of the grid. Thus, if the grid is — 35 volts, then the plate current is 50 milliamps. Now, if an impedance be put into the plate circuit as explained previously, the plate voltage will be less than 200 when the grid becomes more positive, and will be more than 200 when the grid becomes more negative than the mean condition
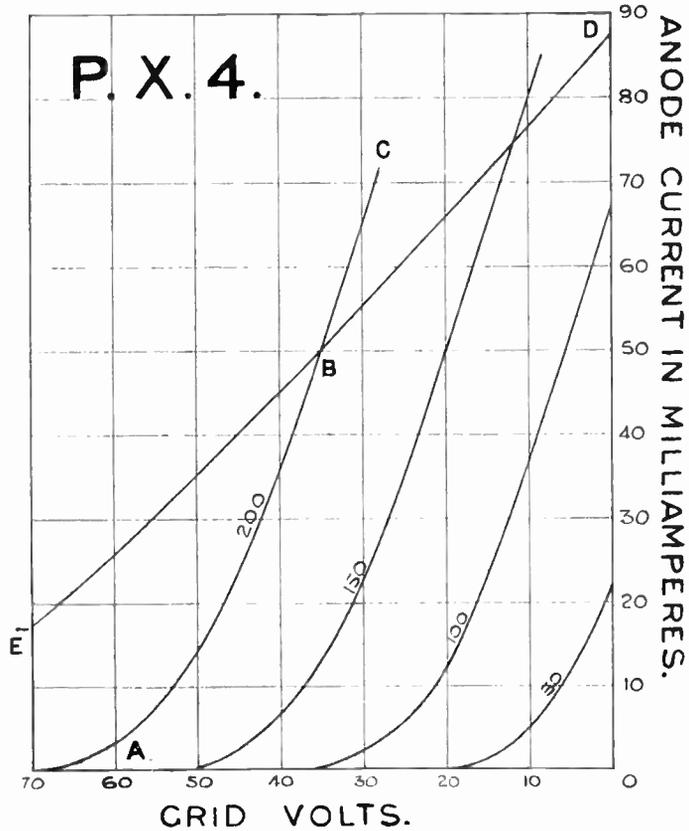
Fig. 93.

of −35 chosen. The plate current curve is therefore the line EBD. The line is drawn for an impedance or resistance of 2,100 ohms, and is almost straight, there being a small general curvature running along it. This curvature is an unfortunate defect of a valve.
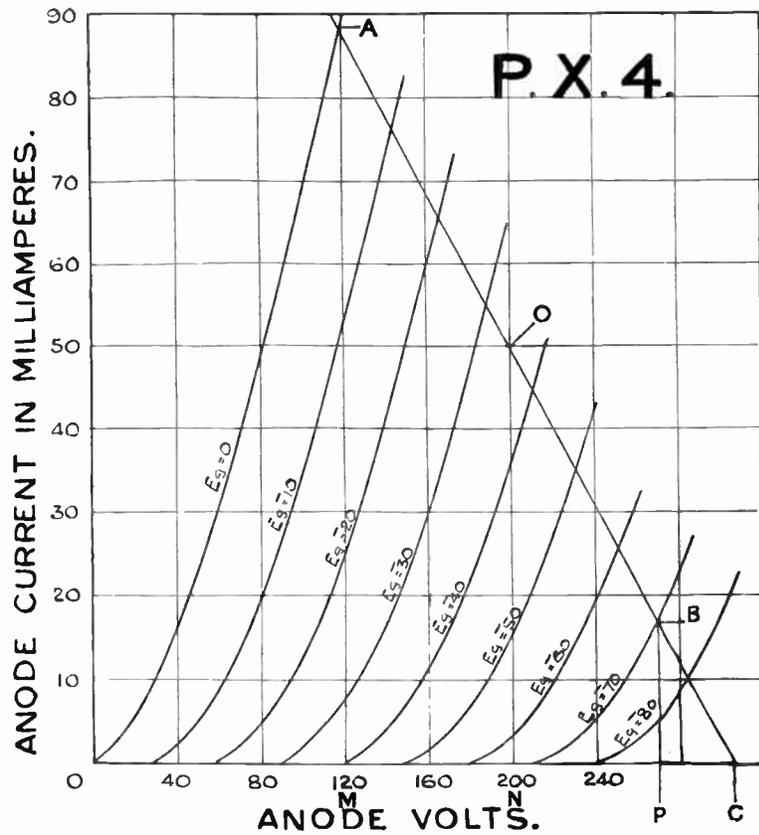
Fig. 94.

The whole matter can be better examined by changing our valve curves into another form. Figure 94 shows plate voltage, plate current curves, each for a definite grid bias, for the same PX4 valve.

Makers are beginning to issue curves for their power valves in this form. The point O, corresponding to a plate voltage of 200, plate current of 50 milliamps, and a grid bias –35 volts, represents the steady condition before any signal

L

comes on the grid. If there is no impedance in the plate circuit, a vertical line through O will give the currents for various grid voltages with no change of plate volts. If the impedance inserted is exceedingly great, a horizontal line will give the conditions, since there is then no appreciable change of current. For any intermediate impedance a sloping line will give the conditions, such as line A O B C representing a resistance of 2,100 ohms. To obtain the line, through O draw O C so that

$$\frac{\text{CN volts}}{\text{ON amps}} = 2,100, \text{ or } \text{CN} = \frac{2,100}{1,000} \times \text{ON milliamps}$$

$$\text{CN} = \frac{2,100 \times 50}{1,000} = 105 \text{ volts.}$$

Join O and C and produce.

Plate impedance should be twice valve impedance.

If the impedance had contained inductance as well as resistance, the line would have been an ellipse and more difficult to construct; 2,100 ohms is twice the resistance of the valve, and both Brown, in *Proceedings of the Physical Society,* April, 1924, and Green, in *Experimental Wireless,* July and August, 1926, showed that for ideal valves with straight characteristics, this gives the maximum power. The law is true for our valves if we work them within the range of small distortion.

The mean grid potential is –35 volts, and since the grid must not become positive, the half swing of grid voltage is 35 volts and the total is from 0 volts to –70 volts. Where the line cuts the curves we can immediately read the plate current and plate volts for corresponding grid voltage.

The following Table gives such a set of figures, together with angles or times over one complete period corresponding to an applied grid signal of –35 $(1 + \sin \omega t)$ where $\omega = 2\pi$ times the frequency of note. This enables us to set out the curves of Figure 95.

FIGURES FROM PLATE CURRENT, PLATE VOLT CURVES OF A PX4 VALVE (FIG. 94), FOR GRID SIGNAL −35 $(1 + \sin \omega t)$.

Angle or Fraction of one period.		Corresponding Grid Volts.	Corresponding Plate Volts.	Plate current for a single Valve. Milliamps.	Push Pull.	
					Signal Current.	Battery Current.
0°− 0′	0	−35	200	50	0	100
8°−12′	·0228	−40	210	45·4	4·9	100·6
16°−36′	·0462	−45	221	40·3	10·1	100·8
25°−22′	·0705	−50	230·5	35·7	15·05	101·5
34°−52′	·0969	−55	241·6	30·3	20·45	101·5
45°−35′	·1267	−60	251·0	25·8	25·25	102·1
59°− 0′	·164	−65	260	21·5	30·3	103·6
90°− 0′	·25	−70	268	17·5	35·15	105·3
121°− 0′	·336	−65	260	21·5	30·3	103·6
134°−25′	·3733	−60	251	25·8	25·25	102·1
145°− 8′	·4031	−55	241·6	30·3	20·45	101·5
154°−38′	·4295	−50	230·5	35·7	15·05	101·5
163°−24′	·4538	−45	221	40·3	10·1	100·8
171°−48′	·4772	−40	210	45·4	4·9	100·6
180°− 0′	·50	−35	200	50	0	100
188°−12′	·5228	−30	188·7	55·2	4·9	100·6
196°−36′	·5462	−25	178	60·5	10·1	100·8
205°−22′	·5705	−20	166·6	65·8	15·05	101·5
214°−52′	·5969	−15	156	71·2	20·45	101·5
225°−35′	·6267	−10	145	76·3	25·25	102·1
239°− 0′	·664	− 5	133·3	82·1	30·3	103·6
270°− 0′	·75	0	121·6	87·8	35·15	105·3
301°− 0′	·836	− 5	133·3	82·1	30·3	103·6
314°−25′	·8733	−10	145	76·3	25·25	102·1
325°− 8′	·9031	−15	156	71·2	20·45	101·5
334°−38′	·9295	−20	166·6	65·8	15·05	101·5
343°−24′	·9538	−25	178	60·5	10·1	100·8
351°−48′	·9772	−30	188·7	55·2	4·9	100·6
360°− 0′	1·000	−35	200	50	0	100
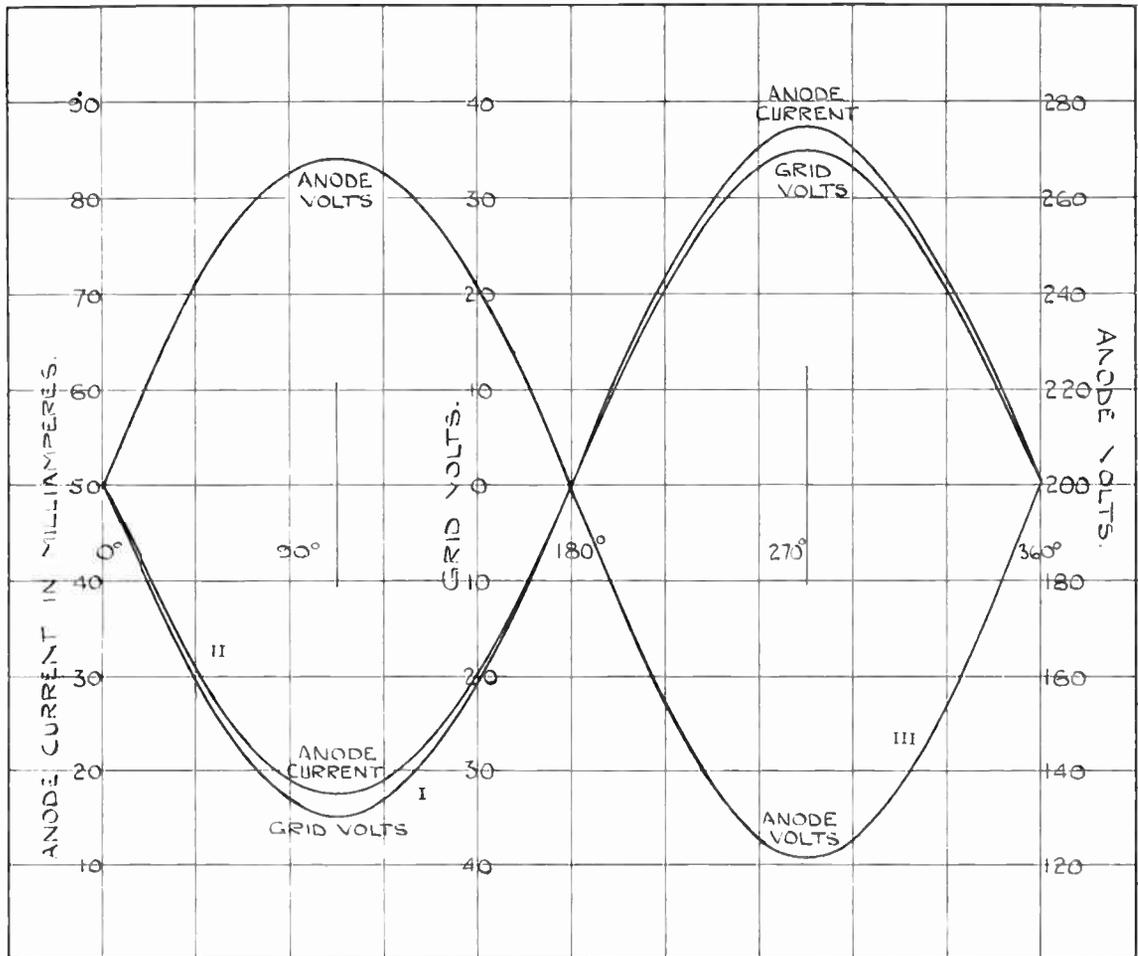
Fig. 95.

Curve I is the pure sine wave representing the grid signal. Curve II represents the plate current. In the negative half it falls inside the pure sine wave form, and in the positive half it falls outside, thus showing that there is very appreciable distortion. This arises from the curvature of the valve characteristic making the distance between curves along the resistance line A O B unequal for equal differences of grid voltage. Curve III, representing the plate voltage, is identical in form with Curve II.

Referring to Figure 94, the swing of the current is from A to B, the amplitude each side of the mean will be $\dfrac{AM-BP}{2}$

which corresponds to $\dfrac{AM-BP}{2\sqrt{2}}$ (RMS). The power is

Calculation of maximum A.C. Power output.

$(\text{current})^2 \times \text{resistance} = \left[\dfrac{AM-BP}{2\sqrt{2}}\right]^2 R = \dfrac{(AM-BP)^2\ R.}{8}$

Alternatively, since PM equals the total voltage swing, the power also equals $\dfrac{(AM-BP)}{2\sqrt{2}} \times \dfrac{PM}{2\sqrt{2}} = \dfrac{(AM-BP)\ PM,}{8}$

which is the usual form given for A.C. power output.

From our curves the maximum A.C power is

$$\frac{1}{8}\left[\frac{87\cdot8}{1,000} - \frac{17\cdot5}{1,000}\right]^2 \times 2,100 = \frac{1}{8}\left[\frac{70\cdot3}{1,000}\right]^2 \times 2,100$$

$$= 1\cdot295 \text{ watts, or } 1,295 \text{ milliwatts.}$$

The amount of distortion is generally indicated by comparing the mean of AM and BP with ON the original current before the signal comes in, and is given as percentage distortion

$$= \frac{\dfrac{AM + BP}{2} - ON}{ON} \times 100$$

In our case $\dfrac{\dfrac{87\cdot8 + 17\cdot5}{2} - 50}{50} \times 100 = \dfrac{52\cdot65 - 50}{50} \times 100 = 5\cdot3\%$

While the signal is on, a milliammeter in plate circuit will read 52·65 milliamps, instead of the 50 formerly, and this difference is an indication of the distortion ; in fact, if the needle moves, there is distortion. But if a valve is giving approximately full power there must be a movement of the needle of about 5%. If the movement exceeds this, the input signal should be reduced to prevent distortion of the music.
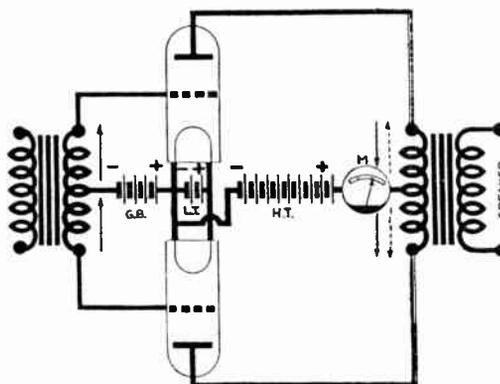
Fig. 96.

Push Pull.   Figure 96 shows a method of connecting two valves in the power stage which has many advantages.

1st : It greatly reduces the harmonics caused by curved characteristics.

2nd : It gives at least double the power of one valve ; because of the reduction of harmonics, and for a given percentage of distortion, it gives considerably more than double that of one valve.

3rd : The arrangement enables the D.C. magnetisation due to steady plate currents of the valves to cancel each other out. Hence, a smaller output transformer can be used. There is also less chance of distortion due to high D.C. magnetisation of the iron core.

4th : The signal passes from valve plate to valve plate ; practically none passes through the battery or H.T. supply. Hence, there is much less chance of feeding back to previous stages, with consequent self-oscillation or " motor-boating."

5th : Any ripples in the H.T. supply, grid bias, or filament supplies do not appear in the transformer or speaker, since they affect each half of the winding by equal amounts in opposite directions.

The method of connection is clearly shown in Figure 96. The audio frequency transformer from the previous stage has it secondary tapped. Hence the designations AF3c, AF5c. The outer ends of the secondary are connected respectively to the two grids. The centre tapping is connected to the negative end of the grid bias, the positive end of which is connected to the negative filament or cathode.

An output transformer with a centre tapped primary winding is used, its outer ends being connected to the plates of the valves and the centre tapping to the positive end of the H.T. supply.

It will be realised that for this system to work correctly, the valves should have identical properties. In practice, however, for power valves up to the size of L.S.6a's, very little trouble has been experienced due to the two not being exactly alike ; only very occasionally have we found valves which would not function together. The trouble shows itself as a violent oscillation, which greatly increases the plate current, and voltage swing, occasionally breaking down the insulation of the output transformer. The insertion of two grid resistances of 50,000 ohms each in the leads near to the grids greatly lessen the chances of this oscillation.

Valve makers have recently advertised their inability to make valves alike, by demanding that the two valves should have different biases, adjusted to bring the two plate currents equal.

For this we supply, on request, the audio frequency transformer type AF5cX, with two separate half secondary windings ; the middle end of each being taken to the
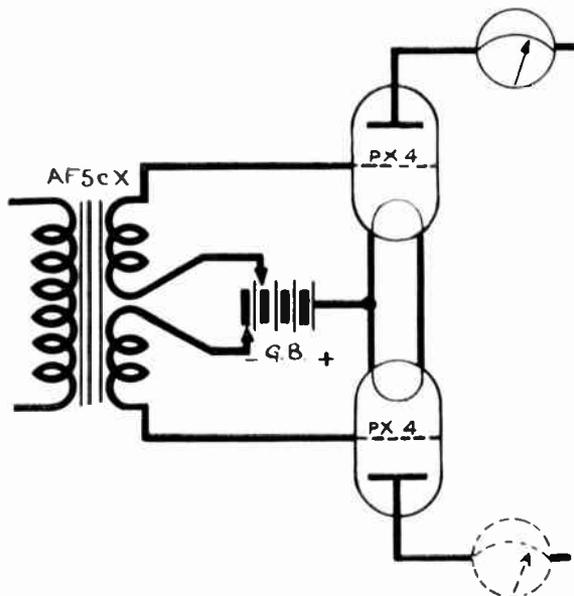
Fig. 97.

suitable bias tapping. This requires the necessary milliam-meter inserted alternately in each plate lead to enable one to adjust the bias suitably.

Adapting Speaker Impedance to Valve resistance. In "Push-Pull" Connection.

The two valves are really in series as regards the applied signal. Hence, the total valve resistance is twice that of one valve. Considering the PX4 valves, used previously, the resistance of one valve is 1,050 ohms, the resistance of the push pull combination will then be 2,100 ohms, and the speaker impedance should be brought up to $2 \times 2,100 = 4,200$ ohms. With the 20 ohms moving coil speaker we shall require a transformer ratio of

$$\sqrt{\frac{4,200}{20}} = 14\tfrac{1}{2}/1,$$

To understand the mechanism of " push-pull," consider Figure 98, where two valves are connected similarly to

Detailed explanation of " Push-Pull." Figure 96, except that in each plate circuit is an exactly similar output transformer of ratio 1/1 with a resistance of 2,100 ohms across each output. The valve is a PX4, having a resistance of 1,050 ohms.
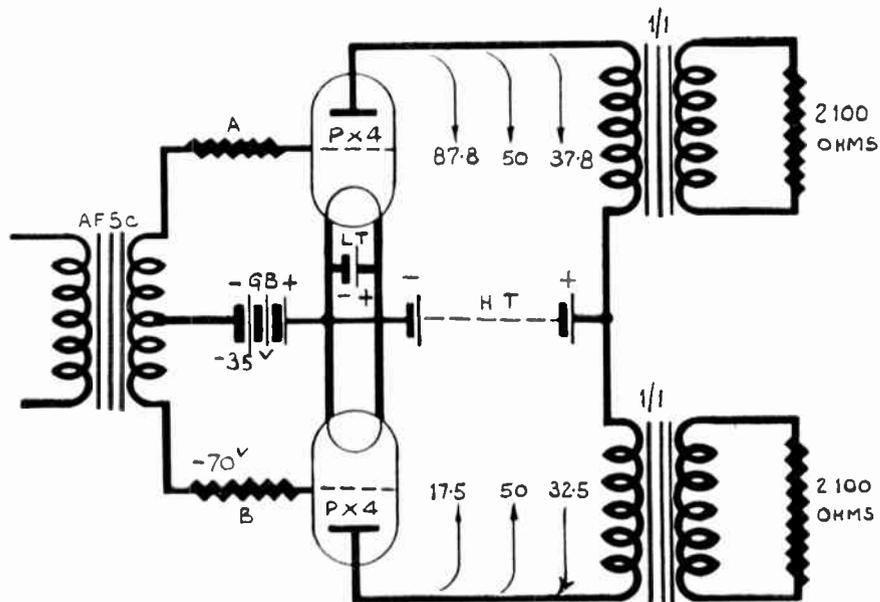
Fig. 98.

If grid A is positive relative to the centre tapping, when the signal is applied to the AF5c transformer, then at the same instant, grid B is negative by an equal amount. Hence, the upper half of the plate current curve (Figure 95) gives the current flowing through the upper transformer, but at the same time the other half gives the current flowing through the lower transformer. Considering the magnetic fields created in the transformer cores, that in the upper core can be considered as made up of a constant part due to the mean plate current of 50 milliamps, together with a variable portion due to the excess current above the mean. Similarly in the lower core, we have a constant part due

to the mean plate current, with a variable portion due to the defect below the mean.

In the figure the current in the upper transformer is shown as 87·8 milliamps, equivalent to 50 milliamps downwards constant, and 37·8 milliamps downwards variable ; in the lower transformer it is 17·5 milliamps, equivalent to 50 milliamps upwards constant, and 32·5 milliamps variable downwards. Since the constant currents of 50 milliamps flow in opposite directions, the two constant fields created are equal and opposite, and since the variable portions of the current flow in the same direction as indicated by the arrows in Figure 98, the variable fields are in the same direction and nearly equal, the difference being due to the distortion caused by the curved valve characteristics, as already explained. When grid A becomes negative relative to the centre tapping, and consequently B positive relative to this, during the negative half of the impressed signal, the constant currents of 50 m.a. keep the same directions, as shown, whilst the variable portions naturally flow in the same direction as each other, though opposite to that shown.
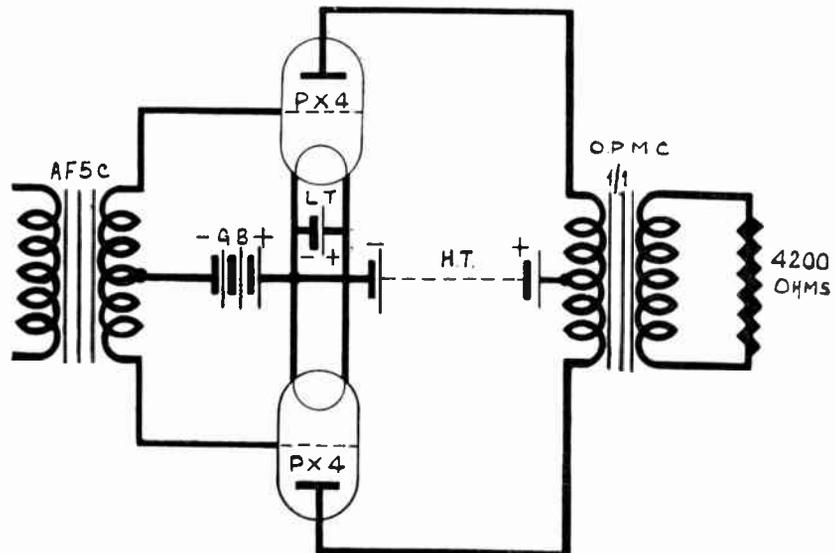
Fig. 99.

We can now take the next step and join the transformers together by one common core. Clearly, then, the equal and opposite constant fields cancel out, leaving no D.C. magnetisation. The variable portions add together; taking the above figures, 37·8 m.a. through one half of the total winding added to 32·5 m.a. through the other half winding, become equal to $\frac{1}{2}$ (37·8 + 32·5) = 35·15 m.a. through the total winding. At the same time, the two resistances on the output side, add together to give 4,200 ohms, and we have obtained the exact equivalent to Fig. 99.
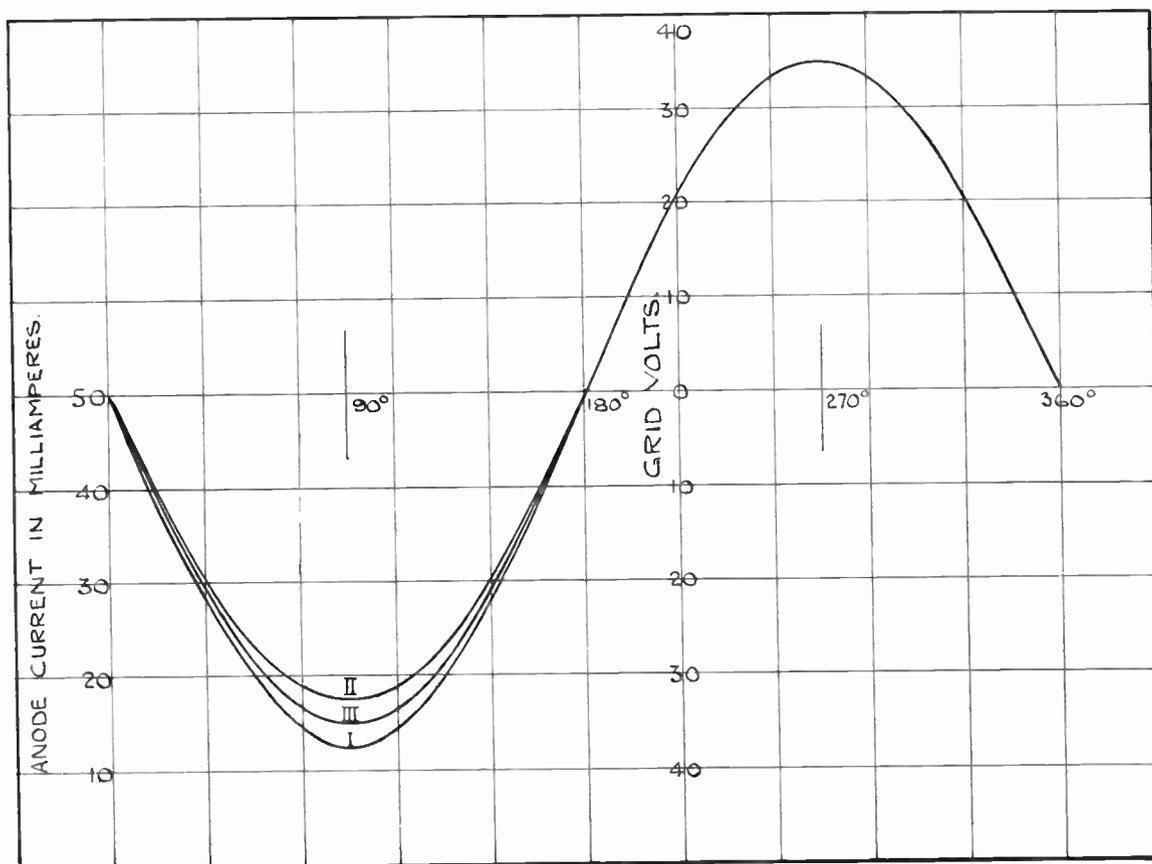
Fig. 100.

In Figure 100, we plot I, the upper half of the current curve, II, the lower half, and III, the mean. Curve III represents the current caused by the push pull valves with total speaker resistance of 4,200 ohms, and is almost a pure sine wave. By the push pull arrangement, we have practically eliminated the distortion caused by the curved characteristics of the valve, and at the same time obtained a current of 35·15 m.a. peak value through a 4,200 ohms resistance, against a current of 35·15 m.a. mean peak value through a resistance of 2,100 ohms in the case of the single valve. *That is, we have obtained double the power* 2,590 *m.w. with two valves in push pull and negligible distortion, to that obtainable with a single valve, namely* 1,295 *m.w. with a distortion of* 5·3%, *a serious amount.*

It will be seen from Figure 99 that each valve requires the same grid swing as when used alone. In the above example, the upper grid swings from 0 to –70 and the lower grid from –70 to 0. *The grid bias used for valves in push pull is the same as that used for the single valve.*

Hence we must have twice the input signal to obtain the extra power of the push pull arrangement. This is the one disadvantage of the connection.

On account of the averaging out of the magnetic field, there is a slight variation of the battery current due to small fluctuations at the mid point of the output transformer. Column 1 of the Table shows the variation. When there is no in-coming signal, the total battery current is 50 + 50 = 100 m.a. During one period of the signal, it varies between 100 and 105·3 m.a., the mean as read on a milliam-meter being 102·6 m.a. Thus it will be seen, *that even with negligible distortion there is a small change in the plate current of the valves, and such a change is not necessarily an indication of distortion.*

Experimental proof of reduction of harmonics by Push-Pull.

The curves demonstrate clearly the manner and amount by which the use of push pull reduces the harmonics introduced by the curved characteristics of the valve. The figures in the following Table give experimental proof of the fact. Two L.S.6A valves were connected in parallel, with approximately 400 volts on the plate, and 100 volts grid bias due to the plate current passing through an 800 ohms resistance. The output is choke capacity, to a step down transformer of $12\frac{1}{2}/1$, across whose secondary is the load resistance of 8 ohms. The stepped up impedance $= 12\frac{1}{2}^2 \times 8 = 1,250$ ohms. The valve resistance $= \dfrac{1,300}{2} = 650$ ohms. The same two valves were then connected in push pull, the transformer connecting the 8 ohm output resistance being 25/1. The stepped up impedance is therefore $25^2 \times 8 = 5,000$ ohms and valve resistance $= 1,300 + 1,300 = 2,600$ ohms.

Output Volts across 8 ohms.	Output Watts.	2nd Harmonic.		3rd Harmonic.	
		Valves in Parallel.	Valves in Push Pull.	Valves in Parallel.	Valves in Push Pull.
2 volts	·5	1·0%	Negligible.	1·2%	Negligible.
4 volts	2·0	3·5%	Negligible.	·5%	Negligible.
6 volts	4·5	6·0%	Negligible.	·5%	Negligible.
8 volts	8·0	9·5%	Negligible.	1·0%	Negligible.
9 volts	10·1	11·7%	Negligible.	3·8%	0·5%
10 volts	12·5	14·2%	0·8%	9·6%	2·5%

The figures speak eloquently of the gain in purity obtained by using push pull. But the ear can be even more convincing. Using a good amplifier and moving coil speaker with a change over switch from a power stage with valves in parallel to one with similar valves in push pull, the gain in beauty of the music with large volume is astounding.
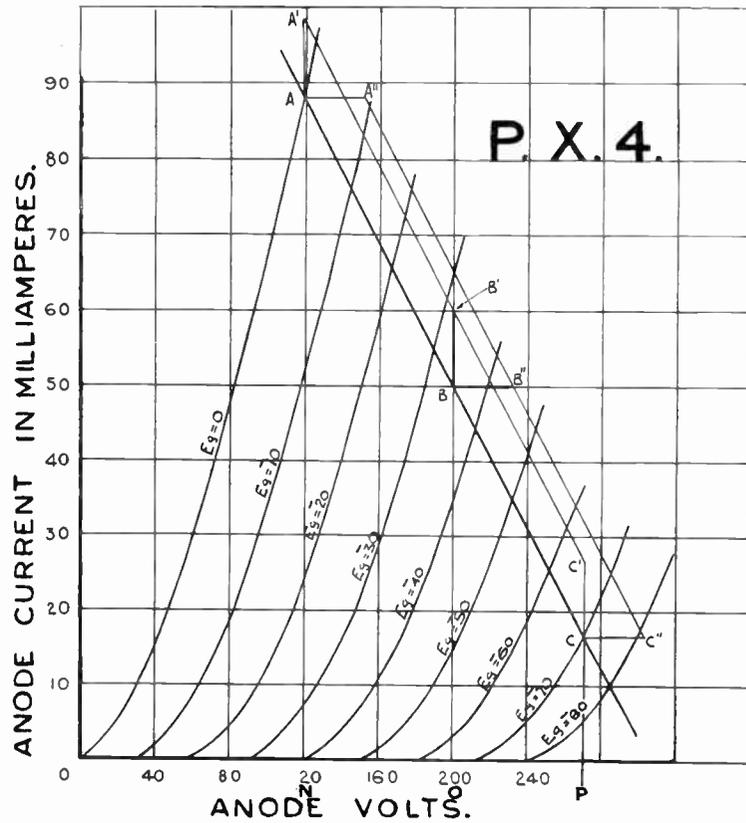
Fig. 101.

Figure 101 shows graphically why ripples or variations in plate voltage or grid voltage have practically no effect on the signal for valves in push pull. When the plate potential and grid bias are steady and represented by point B (200 volts, –35 volts) then the line ABC represents the resistance line (2,100 ohms) and AN, CP represent the currents obtainable for a certain grid swing. If there is a ripple in the grid supply the initial point will be B^1 and the instantaneous currents will be A^1N, C^1P. Since BB^1

will be of a different frequency to the signal that the final signal current will have, in the case of a single valve, or valves in parallel, this ripple superimposed on it will appear as sound in the speaker.

But if two valves are used in push pull, then as explained previously, the current due to the upper valve at some instant will be $(A^1N - B^1O)$ and that due to the lower valve at the same instant will be $(B^1O - C^1P)$. The current through the total winding will be : $\dfrac{(A^1N - B^1O) + (B^1O - C^1P)}{2}$. This equals $\dfrac{A^1N - C^1P}{2}$. But the curves are nearly straight and parallel, therefore $\dfrac{A^1N - C^1P}{2}$ equals $\dfrac{AN - CP}{2}$. That is, the signal current is just the same as if the ripple did not exist ; the effect of the latter on the one valve cancels out that due to the other, and no sign of the ripple appears in the speaker. The same applies if the ripple is B^1B^{11} in the H.T. supply and for a combination of both forms of ripple.

PROPERTIES OF OUTPUT TRANSFORMERS.

These are essentially power transformers used at alternating flux densities very much lower than the usual power station transformer.

The use and connections of these into valve circuits have been fully discussed in the previous sections. The simple formulæ to calculate the correct ratio to adjust a speaker impedance to valve impedance is given on page 155. Here we will deal principally with their faults and losses.

The windings have resistances, and because they cannot occupy the same position in space, there is a small magnetic leakage between them, so that the signal in the primary is
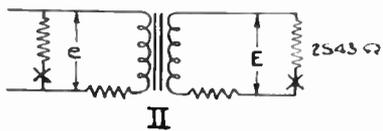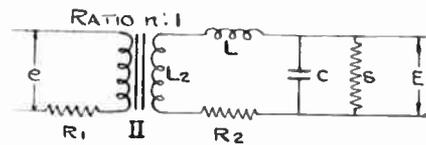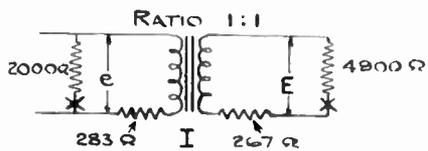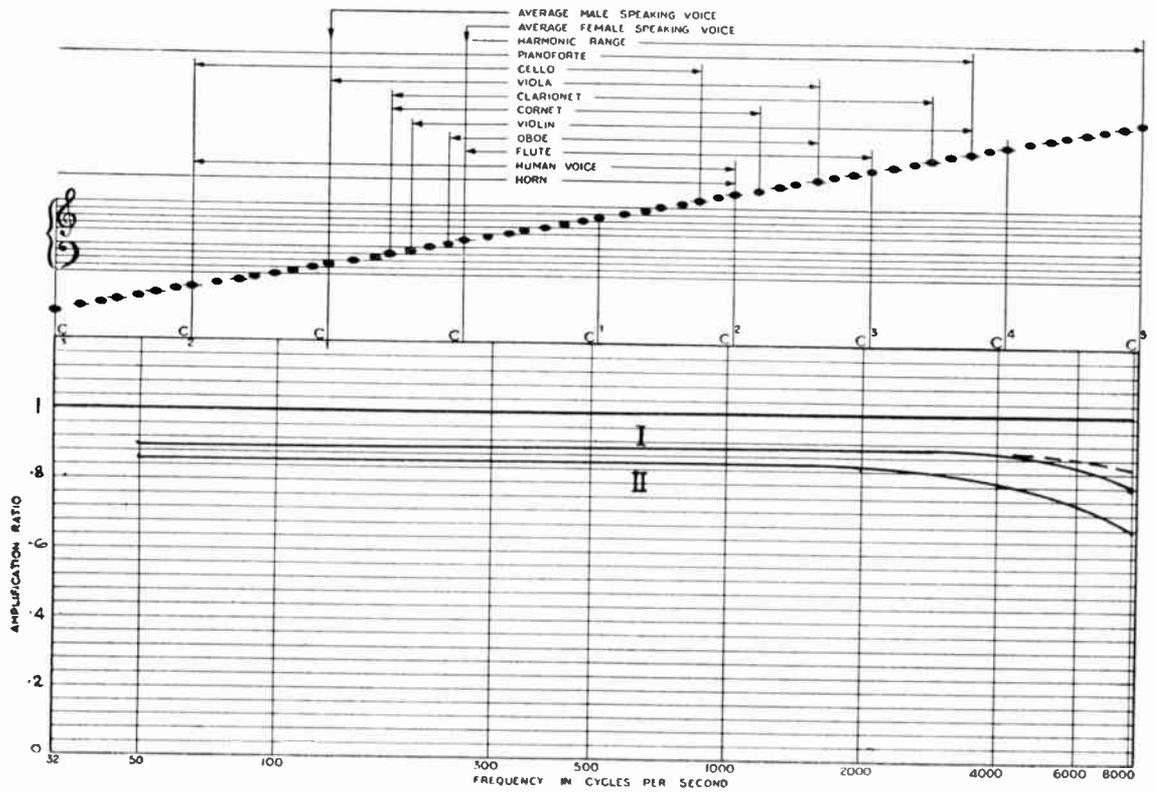
$$n^2 \left(\frac{E}{c}\right)^2 = \cfrac{1}{\left[1 + \cfrac{R_1 + R^2}{n^2} \cfrac{}{S}\right]^2 + \left(\cfrac{\omega L}{S}\right)^2}$$

$L =$ LEAKAGE INDUCTANCE $= L_2 (1 - k^2)$.
$\omega = 6 \cdot 28 \times$ FREQUENCY.

Fig. 102.
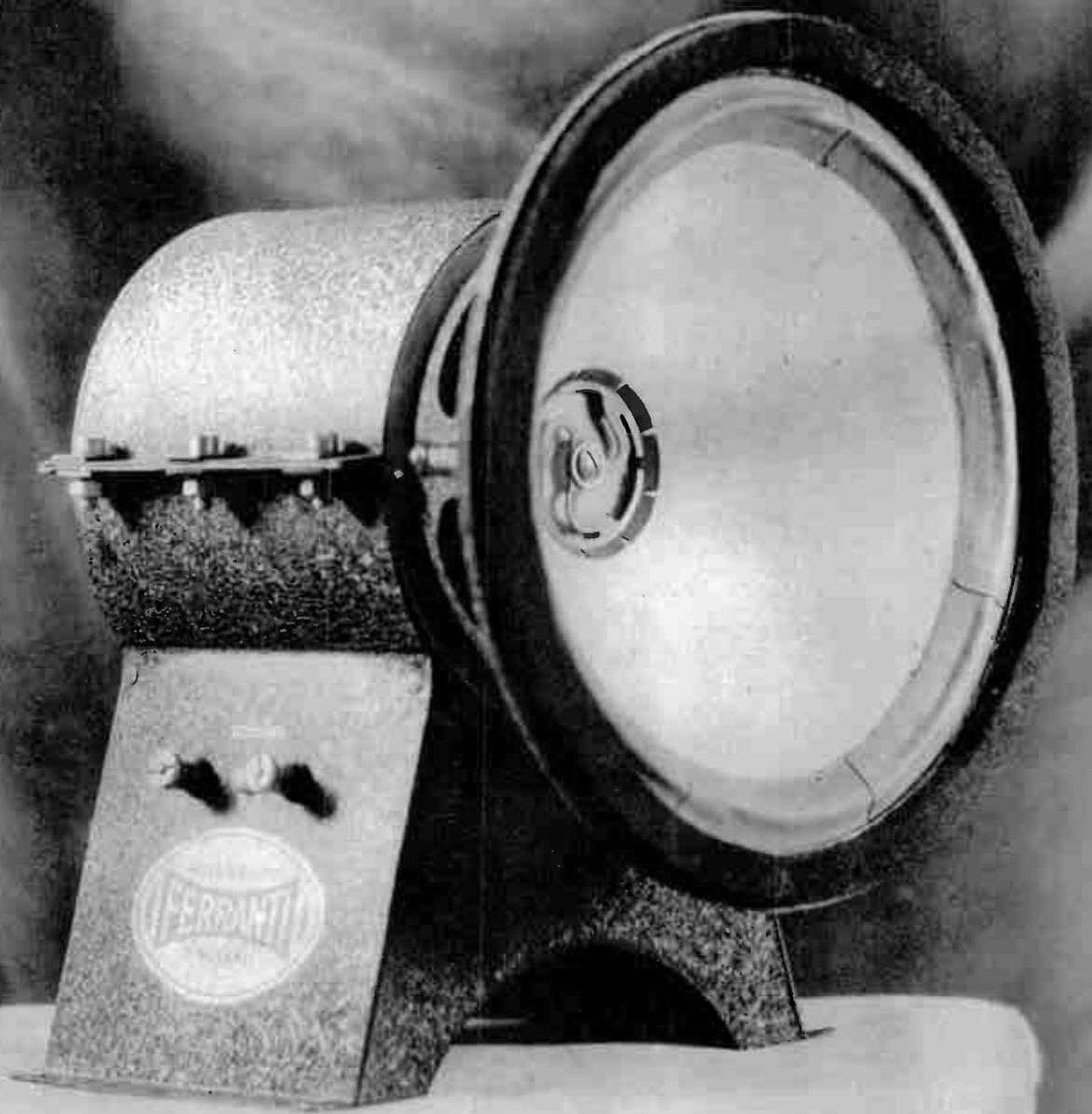
Magno Dynamic Speaker.

not exactly reproduced in amount in the secondary. There are capacity effects between the parts of the windings, though these on examination turn out to be negligible. The same applies to the eddy current losses in the core, equivalent to a resistance across the output at least a hundred times greater than the usual speaker impedance, hence its effect is negligible. Both these play an important part in intervalve transformers (see page 129).

To appreciate the effect of these deficiencies, examine Figure 102, with the curves giving measurements.

Figure 102 shows the transformer ratio 1/1 with a signal e volts applied to its primary, the output voltage E, across the resistance representing the speaker, being measured.

The straight line through 1 represents the result that should be obtained from a perfect transformer. Curve I gives the measurements when the speaker resistance is 4,900 ohms, the correct value, if the transformer is in the plate circuit of a P625 valve. Notice that there is a loss of 10% of the ideal signal from 50 cycles to 2,000 cycles. For higher frequencies there is a further falling off. Curve II gives the measurements when the speaker resistance is only 2,543 ohms. From 50 cycles to 2,000 cycles there is a greater loss—about 14%—and above 2,000 cycles the falling off is greater than before.

We may say here that the constant loss up to 2,000 cycles is due to the resistances of the windings and the further loss above 2,000 cycles is due to magnetic leakage.

Figure 103 shows the same transformer in the plate circuit of a P625 valve, with the correct speaker resistance of 4,900 ohms across the output. Curve I represents the measurements of absolute output voltage for an input to the grid of the valve of 6·5 volts. For comparison, Curve II gives the output voltage for the same input with choke capacity coupling, the primary of the output transformer
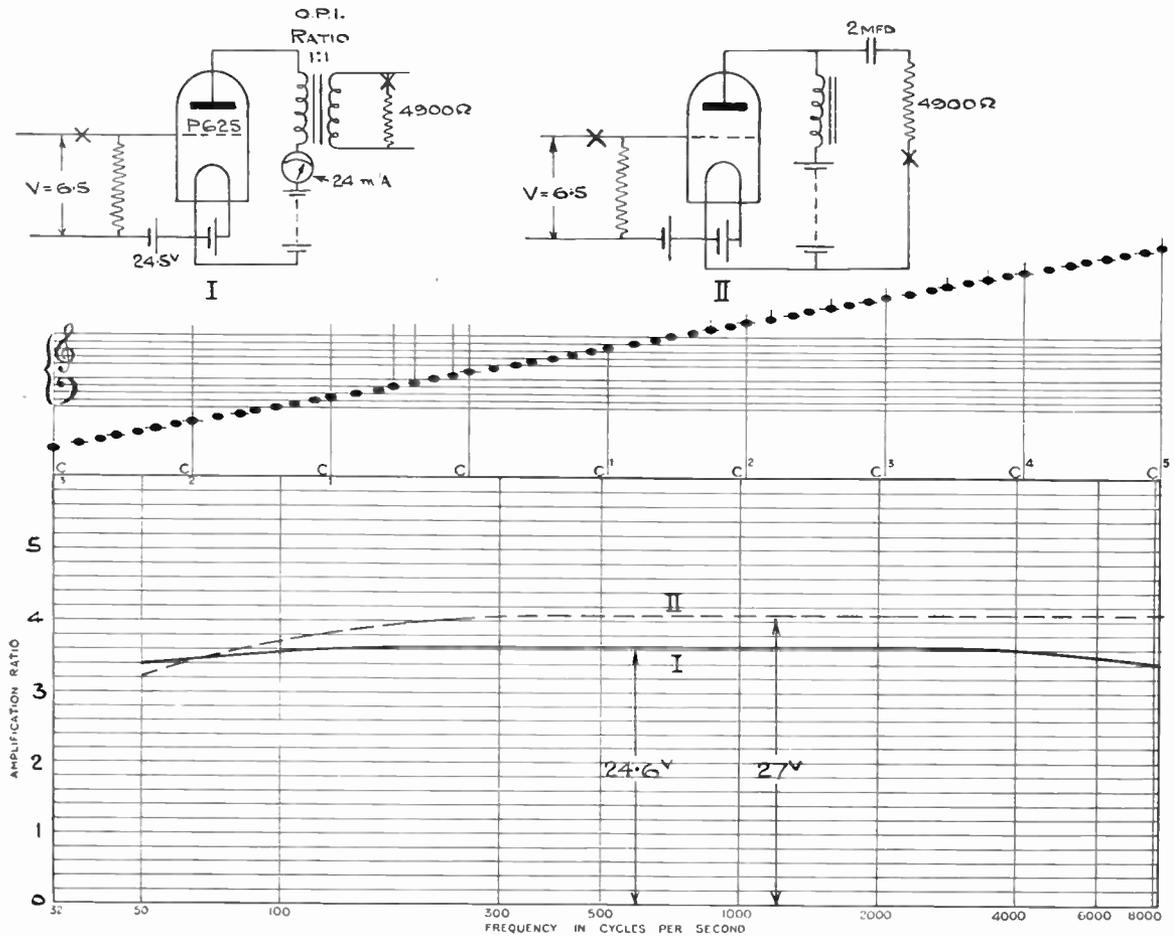
Fig. 103.

being used as the choke. For 500 cycles and above, the 27 volts obtained can be considered as almost the result for a perfect transformer. Hence the OP1 causes a loss of about 10%. In Curve I there is a falling off at both ends of the frequency scale. That at the low frequency end is due to the initial inductance of the transformer windings being insufficient, while that at the high frequency end is due to the magnetic leakage. The greater falling off at the L.F. end in the case of choke capacity coupling is largely due

to the coupling condenser of 2 mfds. not being large enough. We can draw the following conclusions to the design of output transformers for use in the plate circuit of a single valve or valves in parallel.

(1) That the self capacity of the windings has a negligible effect.

(2) That the primary inductance has to be as large as possible to avoid loss of low frequency notes.

(3) That the leakage inductance has to be as low as possible.

(4) That the resistance of the windings must be low.

(5) That the ratio of the transformer must be chosen to suit the speaker and valve.

Unfortunately, to fulfil (2) makes (3) and (4) difficult. The large plate current of the valve increases the difficulty. The design can therefore only be a compromise, and an examination of Curve I with about equal falling off at the two ends of the musical scale, points to OP1 being the best compromise.

Mathematical Section.

For those further interested, we give the mathematical formulæ for the output transformer.

Referring to Figure 102, one finds that

$$n^2 \left(\frac{E}{e}\right)^2 = \frac{1}{\left[\frac{1 + \dfrac{R1}{n^2} + R1}{S}\right]^2 + \left(\dfrac{\omega L}{S}\right)^2}$$

where R_1 and R_2 are the resistances of the windings, S the speaker load, L the leakage inductance, and $\omega = 6\cdot28$ times frequency. L for our transformers varies between $\cdot03$ and $\cdot06$ henries. Hence, for the lower frequencies $\dfrac{L}{S}$ can be neglected and for these

$$\frac{nE}{e} = \frac{1}{\dfrac{1 + \dfrac{R1}{n^2} + R1}{S}}$$

The factor $\dfrac{\frac{R_1}{n} + R_2}{S}$ represents the copper loss.

It has been stated that the usual capacity of the windings has no effect. If it were great enough it would be beneficial, cancelling out the effect of leakage inductance. The experimental proof is, that connecting a ·0011 mfd. condenser across S gave the dotted curve (Figure 102) at the high frequency end, while a ·002 mfd. condenser brought the curve level.

The capacity of the windings can only be a small fraction of the above condensers, hence it can have no appreciable effect. At the same time, contrary to the usual opinion, capacity, such as that due to long leads, etc., across the speaker, can have a beneficial effect if not too great.

Capacity is taken into account in the more accurate formulæ.

$$n^2\left(\frac{E}{e}\right)^2 = \frac{1}{\left[1 - Lc\omega^2 + \dfrac{\frac{R_1}{n^2} + R_2}{S}\right]^2 + \left[\dfrac{\frac{R_1}{n^2}\left(1 + \frac{R_2}{S}\right)}{\omega L_2} - C\omega\left(\frac{R_1}{n^2} + R_2\right) - \frac{\omega L}{S}\right]^2}$$

When the transformer is connected into the plate circuit of a valve, amplification factor μ and plate resistance Rp for a signal Eg applied to the grid is :

$$n^2\left(\frac{E}{\mu Eg}\right)^2 = \frac{1}{\left[1 - Lc\omega^2 + \dfrac{\frac{R_1 + Rp}{n^2} + R_2}{S}\right]^2 + \left[\dfrac{\frac{R_1 + Rp}{n^2}\left(1 + \frac{R_2}{S}\right)}{\omega L^2} - C\omega\left(\frac{Rp + R_1}{n^2} + R_2\right) - \frac{\omega L}{S}\right]^2}$$

or if we neglect the capacity :

$$n^2\left(\frac{E}{\mu Eg}\right)^2 = \frac{1}{\left[1 + \dfrac{\frac{R_1 + Rp}{n^2} + R_2}{S}\right]^2 + \left[\dfrac{\frac{R_1 + Rp}{n^2}\left(1 + \frac{R_2}{S}\right)}{\omega L_2} - \frac{\omega L}{S}\right]^2}$$

PROPERTIES OF OUTPUT TRANSFORMERS.
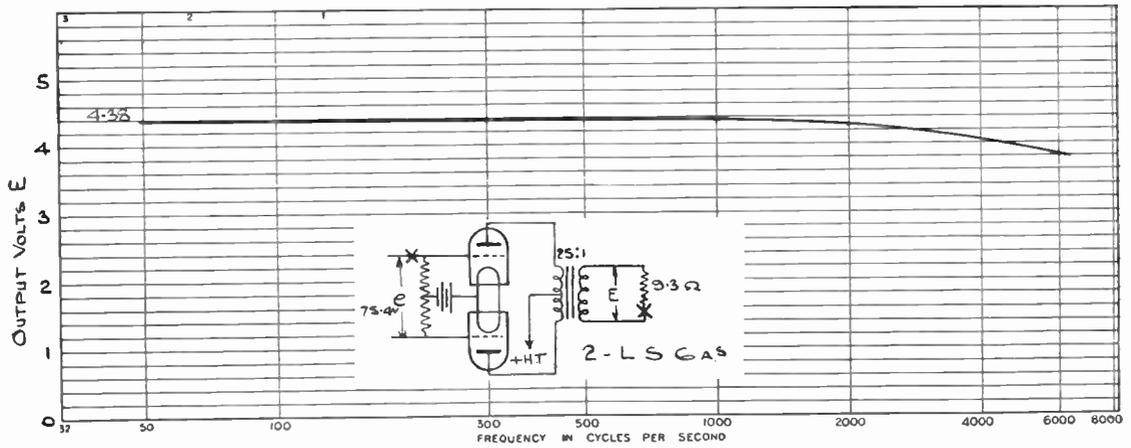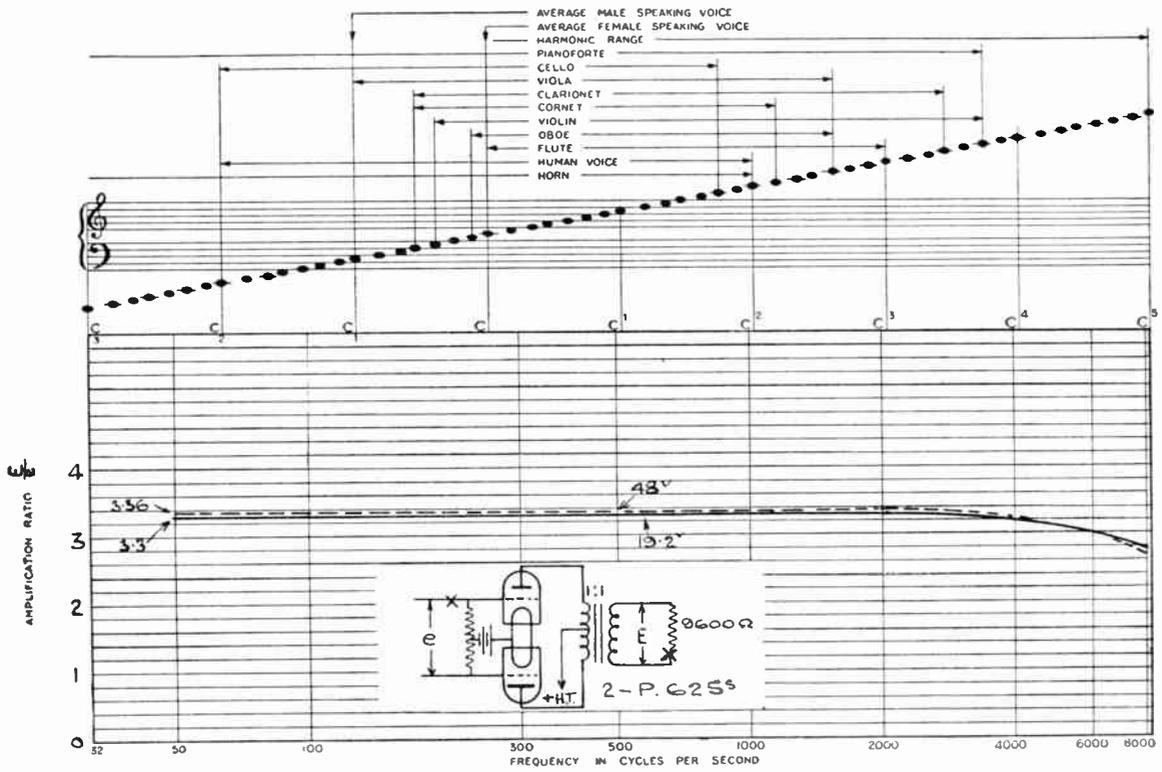
Fig. 104.

Push Pull Output Transformers. The last two formulæ apply equally to transformers in push pull provided Eg is the voltage applied across the two grids and Rp is the resistance of the two valves in series, or twice the resistance of one valve.

Figure 104 shows the measurements for two cases.

(1) An OP1C connected to two P625 valves in push pull.

(2) An OP3C connected to two L.S.6A valves.

It will be noticed that there is no falling off at the low frequencies. This is because the primary inductance is very high, since D.C. magnetisation is absent. Unfortunately, high primary inductance means a rather high leakage inductance, hence there is more falling off at the high frequencies than is desirable. These transformers are now made with less primary turns, giving less fall at the high frequencies.

THE LOUD SPEAKER.

This is the last, but not least, important link in the chain of studio, microphone, transmitter, receiver, and listener.

One is tempted to say that it is the least developed, but thinking of the wonderful reproduction obtainable with a really good moving coil speaker from a good receiver, one hesitates ; we can say that it is the component which gives us the greatest difficulty when attempting to make an accurate measurement of its performance.

All commercial speakers consist of a diaphragm of some sort moved to and fro either by the varying magnetic field caused by the currents obtained as explained from the power stage, or by the varying force obtained from conductors carrying these currents situated in a fixed magnetic field. Development is now beginning on another form in which the signal currents cause a variation of electrostatic field between two diaphragms.

Figures 105, 106, 107, and 108 show four repesentative forms.
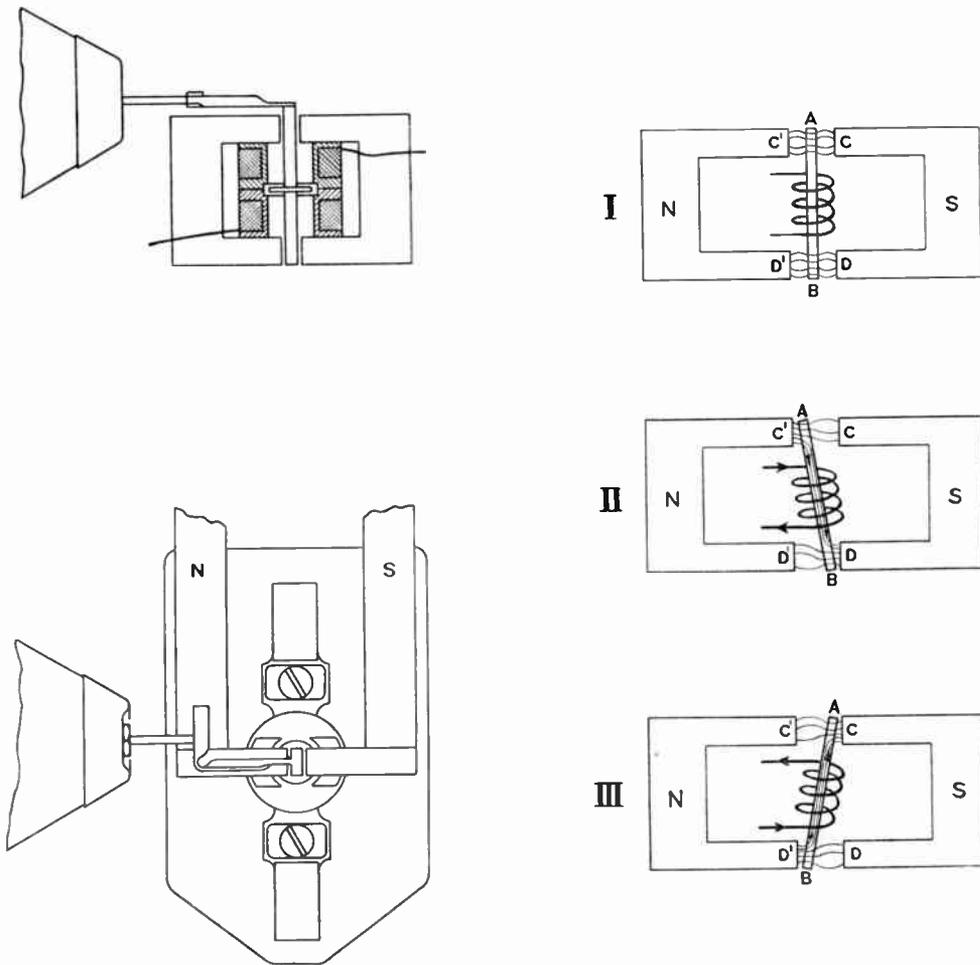
Fig. 105.

Moving Iron Speaker.

The first type, Figure 105, is known as the "balanced armature moving iron" speaker. The armature AB, of soft iron, is pivoted centrally between the soft iron pole pieces CC¹, DD¹, and is restrained by a spring (often the pivot itself) so as to always return to the central position after displacement. The pole pieces are attached to the N and S poles of a permanent magnet. Around the armature is a bobbin wound with a large number of turns of fine wire. When no current passes through this coil, the armature keeps central and equal magnetic flux flows from N to S through the ends of AB and none flows through AB. This is shown in I. When current flows through the coil in the direction shown in II, magnetic flux flows in the armature as shown and the armature has to tilt as shown to allow some of the flux of C¹ to flow along it back to D. Perhaps a better way to appreciate this is to recognise that the flow of current in II makes AB a magnet, of which B is the N pole and A the south pole. Hence, D¹ repels B from it and C¹ attracts A to it.

When, as in III, the current is reversed, A becomes north and B south, hence armature AB is tilted in the other direction. Therefore the variation of the current caused by the signal in the output valve moves AC to and fro against the spring control, and this pulls the cone to and fro, giving the music by imparting its motion to the air.

Often these speakers are connected direct in the plate circuit of the output valve, as explained on page 151. The steady D.C. plate current of the valve will then pull AB to one side permanently. Hence, an adjustment is provided on the controlling spring, so that this can be counteracted and the armature brought central while the plate current is flowing.

PAGE 184
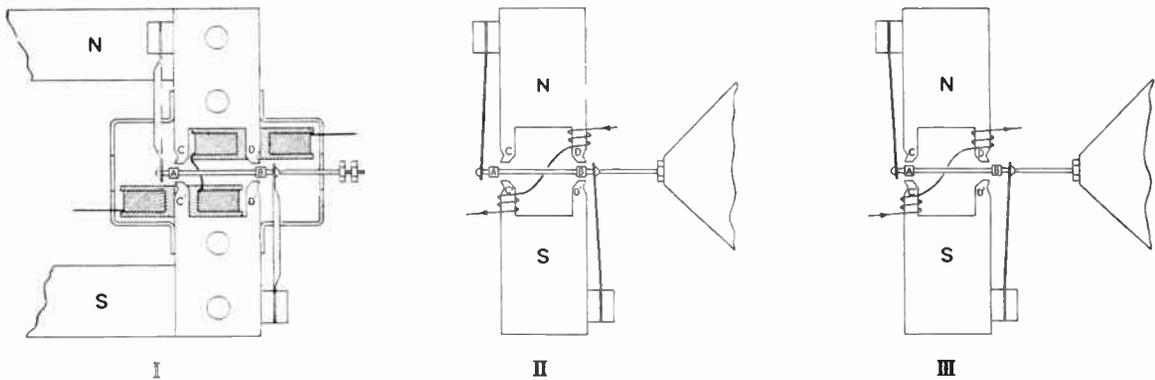
Fig. 106.

The second type of speaker, Figure 106, is known as the "inductor speaker."

The two silicon steel pieces A and B are connected together by a non-magnetic rod and to the cone, and also to the two springs so that they can only move parallel to the faces of the laminated iron pole pieces CC^1 and DD^1; the whole is fitted between the north and south poles of a permanent magnet. The distance between A and B is a little greater than that between the pole pieces C and D, so that when central they "overlap" the latter on their outside and "underlap" on their inside edge, as shown in I. On two of the pole pieces are bobbins of fine wire connected in series. When no current is passing the "armature" AB keeps central and the same magnetic flux flows through CC^1 as through DD^1, the amount being settled by the overlap and the air gap between A to C and B to D.

When current is passed through the bobbins in the direction shown in II it increases the flux in D and decreases it in C. This compels the armature to move to the left against the springs to bring B nearer to the centre of the face of D, and A further away from the centre of C to allow a suitable

path for this change of flux. When the direction of the current is reversed, as in III, the armature AB is compelled to move to the right. Thus, when connected to the valve plate circuit through an output transformer, the signal current fluctuations move AB and with it the cone, to and fro, producing the music. An output transformer is necessary, since if the DC plate current of the valve were allowed to pass through the coils, it would displace the armature completely out of its central position.
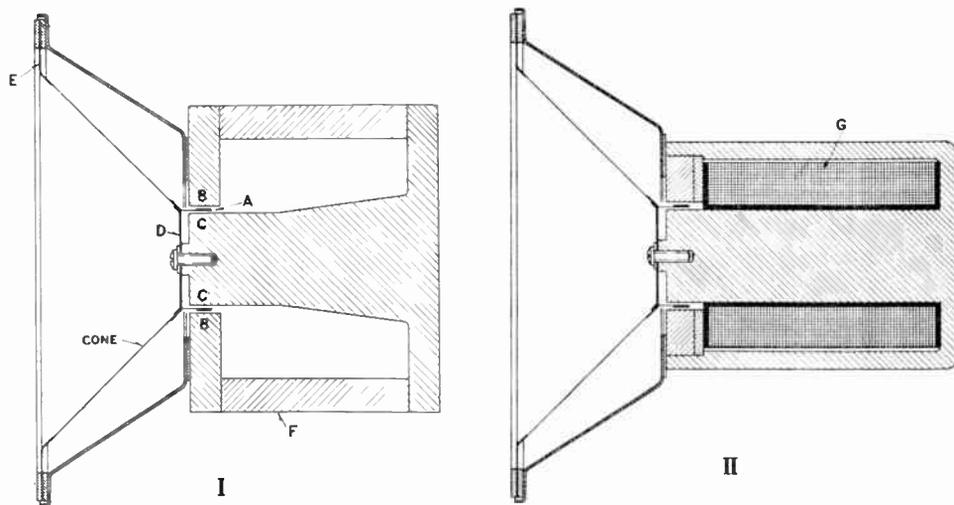
Fig. 107.

Moving Coil or Dynamic Speaker.

The third type of speaker—Figure 107—is known as the moving coil or Dynamic speaker. In it the currents pass through a coil suspended in a magnetic field in such a way that it can move at right angles to the lines of force. Thus in I the coil A is wound on a paper tube cemented to the cone. This coil is central within BC, the pole faces of a magnet, in this case a cylindrical permanent magnet F.

The flexible supports, D a flexible disc fastened to the centre pole piece, and E a soft leather ring to the outer edge of the supporting bowl, allow within limits free movement of the cone and coil, left to right and vice versa. When current passes through the coil in one direction, the interaction between it and the magnetic flux passing across the gap annular BC drives the coils to the left and when reversed to the right. Thus the signal currents drive the coil and cone to and fro, producing the music. In I the magnetic flux is produced by the cylindrical magnet F, passes through the upper pole piece B, across the gap to C, and then down the centre pole piece, through the lower back to F. This type is called a " Magno-dynamic speaker."

Electrodynamic type. In diagram II we have the " Electrodynamic speaker," the only difference being that a large coil of many turns G, through which D.C. current passes, is fitted round the centre pole piece, thus forming an electro magnet, which causes the magnetic flux in the gap.
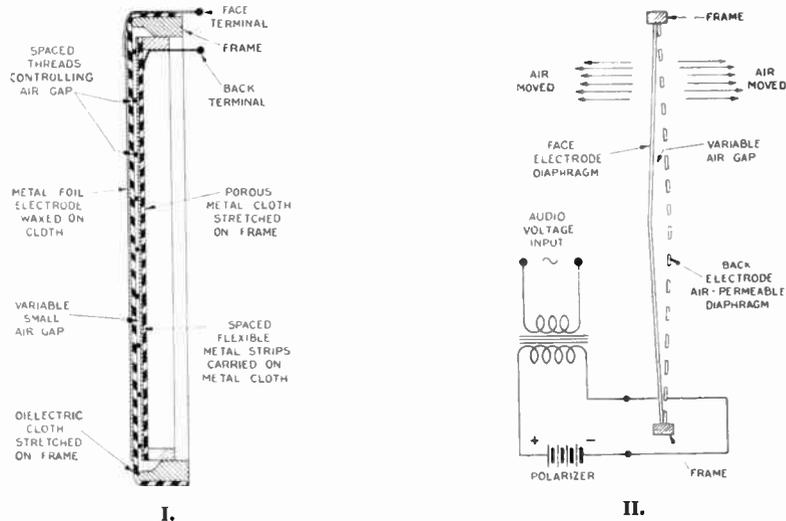
Fig. 108.

Electrostatic Speaker.

The fourth type of speaker (Figure 108) is known as the " Electrostatic speaker," and is not well known commercially in this country. Figure 108 shows the construction of an American sample, and diagram II shows the connections and method of functioning. The two discs held exceedingly close to each other are both conducting. The polarising battery, or other source of D.C. potential, is connected in series with the output transformer of the receiver. The effect of the polarising battery is to cause an electrostatic field between the discs, so that there is a force of attraction between them. The signal being applied in series with this increases and decreases the voltage, so increases and decreases the force. The consequent relative movement of the discs produces the music. The polarising voltage must be greater than the maximum peak signal voltage applied.

The reader having learned how and why these types of loud speakers function, will ask which is best; which shall he purchase? If one can neglect the matter of cost and can be sure that the receiver to work it has a sufficiently large power stage—two power valves of class B in push pull or one of class C, are the minimum (see page 147)—then at present there is only one answer: a properly designed moving coil speaker; all moving coil speakers offered are not good. Despite all statements to the contrary, the reproduction of such a speaker is much superior to that of all the other types. Now let us see the difficulties and faults of each type.

TYPE I. The quality of reproduction of this is the least desirable.

To obtain high efficiency, the clearance between the armature and pole faces is small. Hence the possible movement is small and the spring restoring force has to be large. Small movement means weak reproduction of the low notes.

Also, because the magnetic pull increases inversely as the square of the distance, a strong signal is reproduced much louder than it should be in proportion to a weak signal. There is therefore Amplitude distortion.

The other difficulties apply in varying degrees to all speakers. The armature and attached parts have mass or inertia, there is a returning spring, and hence the moving structure has natural resonances at various frequencies, giving much larger movements to the cone than the proper amount due to the impressed signal. The notes corresponding to these frequencies will be exaggerated. The cone itself is an elastic structure, and can vibrate along the length of its face, leading to more resonances and exaggerated notes.

TYPE II. Here the armature moving parallel between the pole faces is capable of quite large movement ; hence it can reproduce low notes more satisfactorily than Type I. The returning spring force need not be large, so that structure resonances are of lower frequency and less important. By the special shaping of the pole faces, the change of magnetic flux can be made nearly proportional to displacement, so that there need be no serious amplitude distortion. The greatest objection is its low efficiency, necessitating the use of two power valves of class B or one of class A in the output stage. The writer feels that this difficulty will be reduced in the near future by better design.

Both Types I and II have a serious defect leading to the weak reproduction of the high musical notes. Their impedance increases very rapidly at the higher frequencies (see Figure 87), which so limits the current that the speaker cannot reproduce these higher notes.

TYPE III. This has everything in its favour except cost. The driving part consists of a coil of wire, in the Ferranti Dynamic Speakers, of 120 turns of 37's enamel wire on

$1\frac{1}{2}''$ mean diameter, weight 3·02 grammes, thus adding little weight to that necessitated by the cone itself. The suspension can be made to have very little returning force or spring, so that the mass resonance frequently is very low, usually 50 cycles or less. Such resonances as exist are those due to vibrations of the cone structure itself, and are not very serious. The response is good from 50 cycles to 5,000 cycles, and even the 8,000 cycles note is in evidence. There need be no amplitude distortion if properly designed ; apart from the few resonances, the movement of the cone is strictly proportional to the strength of the input signal.
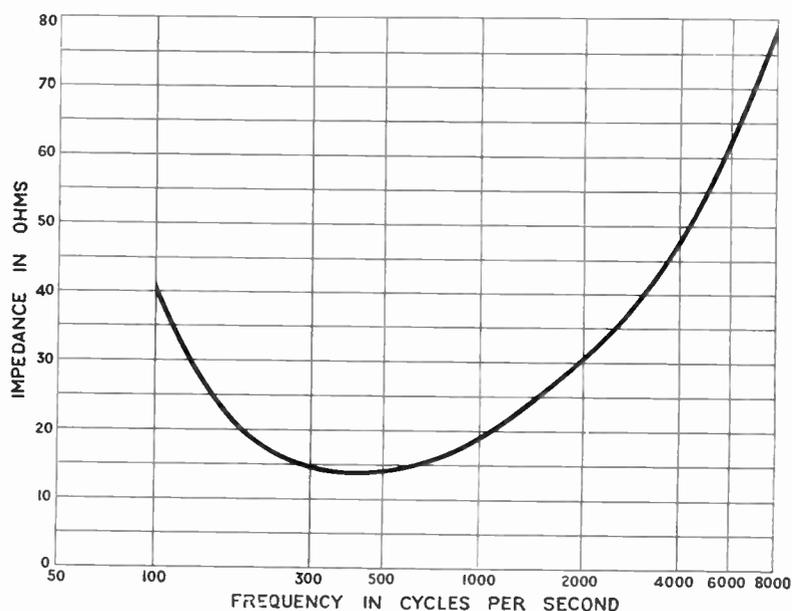
Fig. 109.

Its impedance varies least of any type over the frequency range. Figure 109 gives the impedance curve of the Ferranti Magno-dynamic Speaker.

Type 4. The writer has no experience of this type, and can say but little. The possible motion is somewhat limited, hence the lower notes can hardly be well reproduced.

ENERGY SUPPLIES TO A RECEIVER.

We have dealt with a receiver from all points of view with the exception of the energy supply—that is, current for the filament, voltage and current for the plates, and the necessary voltage for the grid bias.

These are usually obtained from dry batteries, accumulators, Direct Current or Alternating Current electric mains by the use of suitable apparatus.

It is quite evident that the filament supply and the H.T. supply for each stage of the receiver must come from a common source. This can lead to serious trouble, and the next section deals with this regarding high tension batteries or accumulators.

ANODE RESISTANCE FEED SYSTEM, AND ITS ADVANTAGES.

Effect of Common Impedances for all Valves.

All high tension batteries, especially the cheap variety, have high internal resistances. When the H.T. to all the valves of a receiver is supplied by one common battery, the large alternating current generated in the plate circuit of the last valve or valves causes an A.C. voltage to be produced across the H.T. battery owing to its internal resistance.

This A.C. voltage is passed by the battery connections to the earlier stages of the receiver, and is equivalent to audio frequency regeneration, more commonly known as " back coupling."

If this regeneration is great enough, self-oscillation occurs, resulting in the volume of the output being considerably reduced and a high pitched whistle is heard. This trouble is more pronounced with good transformers than with inferior ones, owing to the greater amplification given by good transformers.
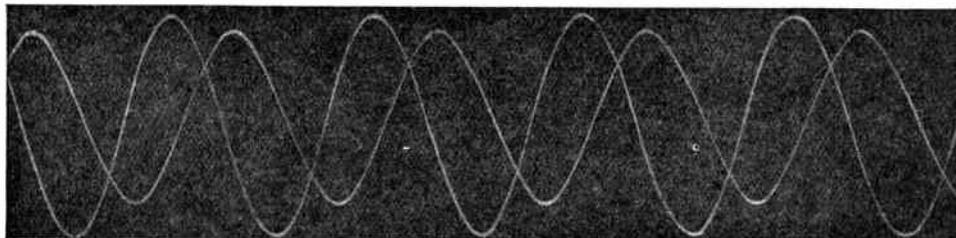
Fig. 110.

The oscillogram, Figure 110, shows the input and output signals from a set consisting of a detector and two audio frequency stages. The H.T. battery was composed of four –30 volt accumulators in series. The output curve appears slightly smaller than the input because the scales are different, but it will be seen that the curves are similar in shape.
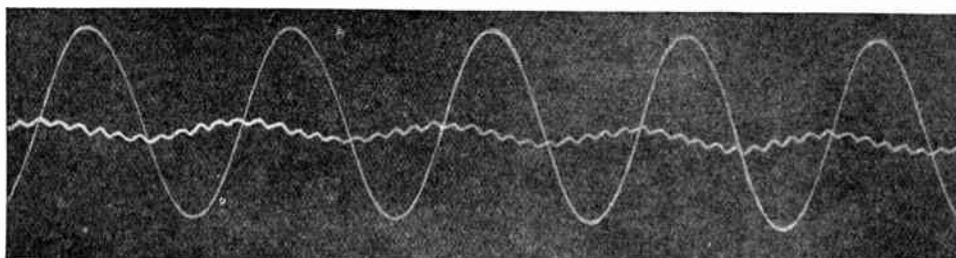
Fig. 111.

The oscillogram, Figure 111, shows the input and output of the same set when a resistance of 60 ohms was connected between each H.T. accumulator—that is, a total of 180 ohms resistance. It will be seen that the output is very much smaller than before, and that it has a high frequency ripple imposed upon it. This ripple represents the whistle heard in the speaker.

Regeneration due to this back coupling may not be enough to cause self-oscillation, but nevertheless it alters the frequency characteristic of the set, making the amplification more uneven.

This trouble can be avoided completely by using a separate H.T. battery for each valve, but this is too costly and cumbersome.

The Anode Feed System.

The other method of avoiding back coupling and regeneration is the Anode Feed System, which has been developed by Ferranti. This system consists of taking the H.T. Supply for a Receiver from the maximum H.T. tapping, to each valve, through a resistance great enough to reduce the voltage to that required for each valve, and connecting a 2 mfd. by-pass condenser from the end of the resistance

Fig. 112.

nearest to the transformer primary, to the H.T. negative; these condensers are usually incorporated in well designed sets. Figure 112 shows an ordinary 3-valve circuit, the anode feed resistances, and associated condensers, being more heavily lined. The principle on which this system depends is as follows :—

The audio frequency impulses from each valve have normally two paths open to them, one being through the H.T. battery to H.T. negative, and the other through the by-pass condensers connected across the H.T. tappings. If the return path is through the battery, regeneration results, and it is therefore necessary to restrict the return path to the by-pass condensers, which, however, have an appreciable impedance at very low frequencies. If, therefore, anode feed resistances are inserted as shown in the diagram, they so increase the resistance of the return path through the H.T. battery, or other source of H.T. supply, that almost the whole of the impulses are by-passed through the condensers. When this scheme is carried out correctly, the overall amplification of the set at all frequencies is exactly the same, as when each valve is fed from a separate H.T. battery.

This has been confirmed by measurement, which proves that the overall amplification of a receiver employing the anode feed system equals exactly the product of the amplification of each stage when measured under working conditions.

To calculate the resistance necessary in a particular case, consider the following example :—

The detector valve DE.210, with 60 volts on the plate, when arranged for grid leak rectification at this voltage, passes a plate current of 4 milliamps. Assume that the H.T. voltage is 120. Then the voltage to be dropped in the resistance is 60 volts.

$$\text{Resistance} = \frac{\text{Voltage to be dropped}}{\text{current (expressed as a fraction of an ampere)}}$$

$$= \frac{60}{\cdot 004} = 15{,}000 \text{ ohms.}$$

For the first L.F. stage, with the same valve but with 90 volts on the plate and 3 volts negative bias, the plate current is 2·5 milliamps. The voltage to be dropped in the resistance is 30 volts.

$$\text{Resistance} = \frac{\text{Voltage to be dropped}}{\text{current (expressed as a fraction of an ampere)}}$$

$$= \frac{30}{\cdot 0025} = 12{,}000 \text{ ohms.}$$

(a 10,000 ohm resistance will be suitable.)

By means of the Chart, Figure 113, the anode feed resistance required in any particular case may readily be determined.

The following example shows how to use this chart :—

Assuming that the total D.C. voltage available is 220 volts, the voltage required in a certain valve is 100 volts, and under these conditions it will consume 2 milliamps. Refer to the curve marked 220 volts at the bottom right-hand side of the chart. Follow that curve to the line against which is given the voltage actually required. Then follow the diagonal line upwards across the chart until the vertical line is reached at the top of which is specified the required milliamp. consumption. The horizontal line which intersects the vertical line at that point gives the value of anode feed resistance required.

It will be noted that zones are marked on either side of the horizontal lines, throughout the width of each zone the value of resistance given against the horizontal line should be employed.
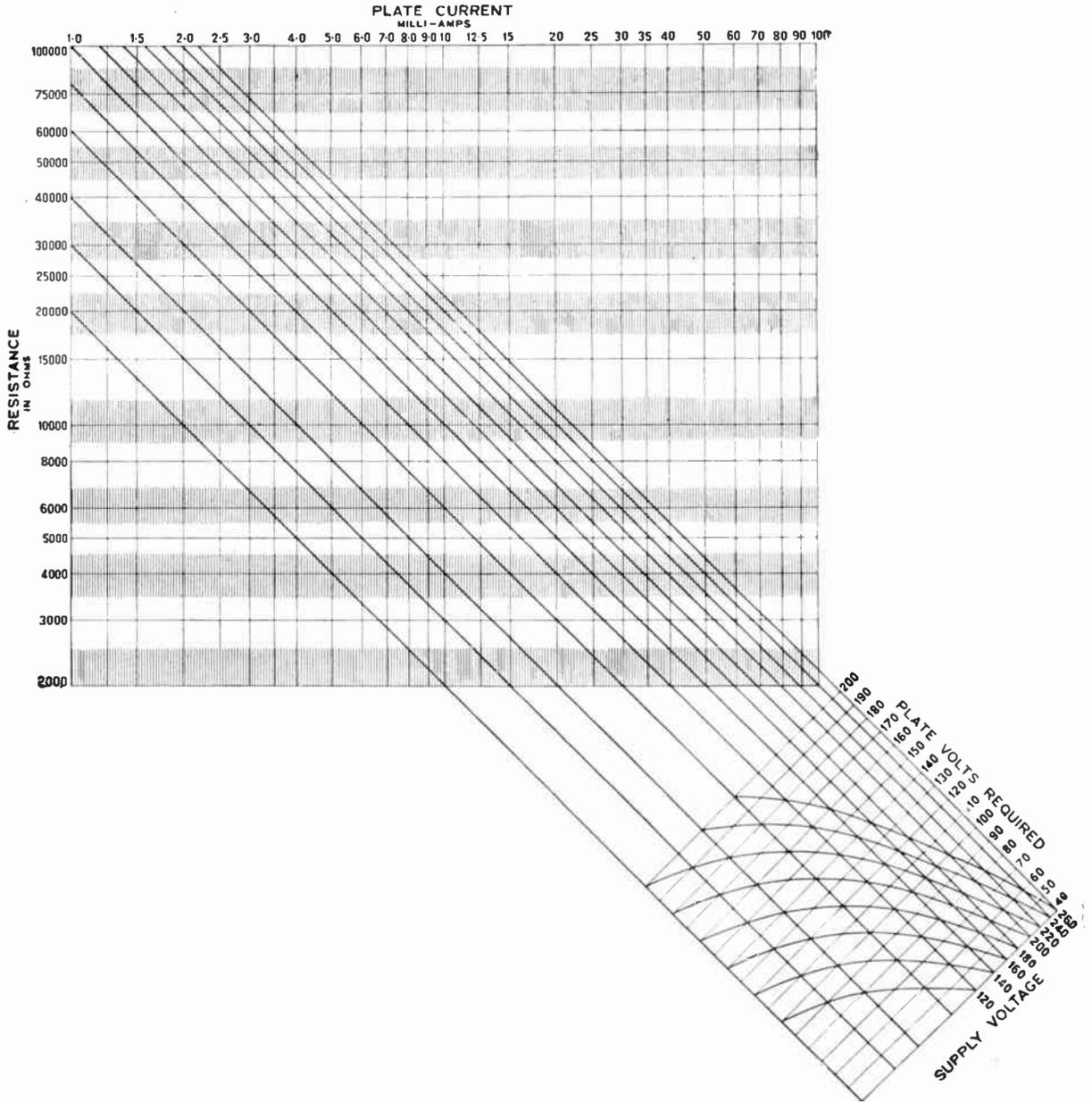
Fig. 113.

In the case of an anode rectifier or screen grid of a screened grid H.F. valve, the current flowing is small, and therefore, instead of using one resistance as in other cases, two resistances should be used. These are connected in series across the H.T. supply, and a 2 mfd. condenser must be connected from the joint of the two resistances to the H.T. negative. The H.T. for the anode or screened grid, as the case may be, is taken from the junction of the two resistances.

It is convenient to make the total value of the two resistances in series about 100,000 ohms, the values of the respective resistances being proportional to the voltages across them.

EXAMPLE: Assume that 200 volts are available and that 80 volts is required for the screen of the screened grid valve, the resistances will be of the ratio 80 to 120, or, in other words, the voltage across one resistance will have to be two-fifths, and across the other three-fifths, of 200 volts.

This will necessitate the use of resistances having values of 40,000 ohms and 60,000 ohms respectively, the 40,000 ohm resistance being connected to the H.T. negative.

The Anode Feed System has the further advantage of necessitating the use of only 2 H.T. terminals in the set.

Moreover, when an accumulator is employed for providing the H.T., the use of the Anode Feed System avoids unequal discharge of the cells and materially prolongs their useful life.

The other sources of energy supply mentioned, namely D.C. and A.C. electric mains, require smoothing apparatus consisting of chokes, resistances, and condensers which can

cause feed back and regeneration in just the same way as high tension batteries. In fact, the regeneration effect is usually much more marked, but owing to the large amount of inductance and capacity contained in the necessary smoothing apparatus, the self-oscillation produced is not of the nature of a high whistle, but is one of very low frequency, appearing as a sort of popping noise usually known as motor-boating. Either the receiver or the smoothing apparatus must contain the anode feed system.

The Ferranti H.T. supply units, to be described later, for the purpose of connecting the receivers to electric mains, are so arranged that back coupling does not exist.

HIGH TENSION SUPPLY UNITS.

An H.T. supply unit, or H.T. battery eliminator, is a piece of apparatus to enable high tension voltage, and current, for a radio receiver to be obtained from the Electricity Supply Mains.

There are two distinct types :—

(1) For use on Direct Current Supply.

(2) For use on Alternating Current Supply.

It is essential, before either purchasing or constructing an H.T. supply unit, to ascertain the nature of the supply ; whether alternating current or direct current, the voltage, and, if alternating current, the frequency.

Before installing an H.T. supply unit, if the nature of the supply is unknown, write to the Electricity Supply Authority informing them of your intention, and asking for the required information. This information must definitely apply to the premises on which the H.T. supply unit is to be installed.

Very serious consequences may arise if an H.T. supply unit is connected to a supply for which it is not suitable.

Since H.T. supply units, and all electrical apparatus, can be dangerous unless certain simple precautions are taken, the attention of the constructor is drawn particularly to the following points :—

(1) All apparatus must be enclosed, so that accidental contact with any live part, terminals, wires, etc., is impossible.

(2) Suitable fuses must be incorporated in the apparatus.

(3) All metal cases must be earthed.

Safety Box.

These requirements are insisted upon by Electricity Supply Authorities before they will certify apparatus for connection to their Mains. To enable these requirements to be complied with, we supply a fool-proof metal Safety Box within which any H.T. supply unit of reasonable size, including all we illustrate, may be built A 3-way flexible cable is used for connection to the Mains through a suitable plug, the third wire (WHITE) being available for earthing. A switch, and fuses of a simple renewable pattern, are provided, the switch being arranged in such a manner that raising the lid automatically disconnects the Mains from the rest of the apparatus. The illustration facing page 144 shows the appearance of an eliminator built into this box.

H.T. Supply Units for A.C. Mains.

There is one additional precaution to be noted, namely, that the speaker must be so connected that there is no high tension in the output terminals of the set if these are accessible, nor in the leads from these to the speaker, nor in the speaker terminals.

This precaution is insisted upon by the Authorities, and is incorporated in the Regulations of the Institution of Electrical Engineers. The best method of isolating the

speaker is to use an output transformer, as shown in Figure 89, or a choke filter may be used, as indicated in Figure 88. If a choke filter is employed, a blocking condenser of at least 2 mfds. capacity, and preferably larger, must be connected, as shown in Figure 88.

Contact with the terminals or connections to a dry battery giving 120 or 150 volts can be unpleasant, but not very dangerous, owing to the limited power, but with apparatus connected to the Electricity Supply Mains, there may be the whole Power Station behind it, and contact can be dangerous to life.

In the case of H.T. supply units for alternating current supplies, using a rectifying valve, there will usually be more than 400 volts across the secondary terminals of the transformer feeding the rectifying valve. In view of this, we feel reluctant to supply this type of transformer unless the safety box is ordered at the same time.

On D.C. supply, the following additional precautions are required :—

Three Wire System.   A great many of the D.C. mains in this country are connected on what is known as the three-wire system. In some houses the negative wire entering a house may be earthed, and the positive wire may be 200 or 230 volts above earth potential. If this is the case, one additional and very important precaution *must* be taken. The negative L.T. or earth terminal of the set should be disconnected from the ordinary earth connection and should be connected to one terminal of a condenser of at least 2 mfds. capacity of the high voltage type C.1. The earth wire which has normally been connected to the Set must be connected to the other terminal of this condenser. This condenser should be

incorporated inside the Receiver or filter in a position such that contact cannot be made with its terminals This condenser is shown clearly in position in the Figures showing the circuits of D.C., H.T. supply units. *Failure to carry out this precaution will lead to the shorting of one of the mains, with serious consequences.*

Other houses in the same neighbourhood will be connected to the other pair of mains, in which case the positive main will be earthed, and the negative will be minus 200, or minus 230 volts below earth.

In addition to these precautions, others are necessary. It must be remembered that the whole set, the aerial, and the low tension batteries, now have a voltage minus 200/230 relative to earth and, therefore, contact with any part may give a shock. The low tension battery must stand on insulating material. A dry sheet of glass or ebonite is a satisfactory insulator.

The aerial wire should be taken from the set and connected to one terminal of a condenser (usually ·005 mfd.), the other terminal of the condenser being connected to the aerial terminal of the set. This condenser should preferably be incorporated inside the set, as it isolates the aerial and prevents shock from the aerial terminal and from any connections to the aerial itself, but the other terminals, wires, and parts of the set, including the low tension accumulators, must be protected from accidental contact, otherwise a serious shock may be obtained.

D.C. Mains. With regard to D.C. H.T. supply units, there are D.C. mains so bad, particularly where the Supply Authority uses mercury arc rectifiers to supply the current, that the mains hum can only be reduced by considerably increasing the capacity of the earthing condenser. This should usually be of 2 mfds. capacity, but in certain very bad cases the capacity may have to be increased to 20 mfds., or even more.
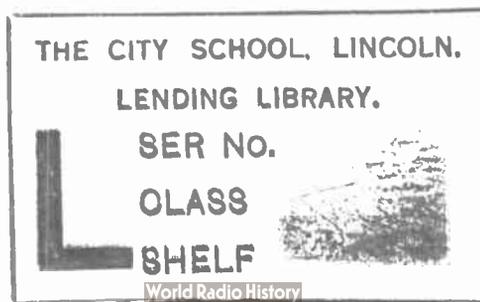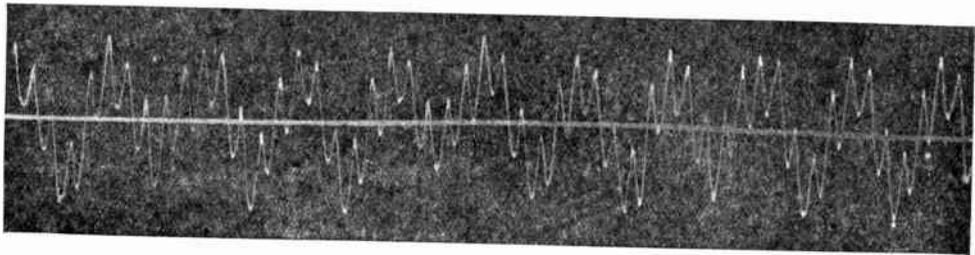
Fig. 115.

When an attempt is made to use the direct current mains for high tension supply without any smoothing, a violent noise is heard in the speaker; this is due to the voltage of the mains varying rapidly. The oscillogram, Figure 115, shows the variations of a particular supply. The bright central line shows the mean voltage of 230 and the wavy line shows the peculiar way in which the instantaneous voltage varies. It rises in this particular case to a maximum of about 236 volts and falls to a minimum of about 224 volts, and the sound corresponding to this wave form, or ripple, is heard in the speaker.

A D.C. H.T. supply unit is necessary to smooth out the ripples in the Mains Supply. It consists of a combination of impedances (either chokes or resistances), in series with the H.T. leads to the set, and having condensers connected in shunt across the leads before the chokes and resistances.

The chokes and resistances offer great opposition to the variations, while the condensers, owing to their property of charging and discharging rapidly, in effect by-pass the variations from the positive lead to the negative so that the current flowing from the choke or resistance to the set has very much less ripple in it.

Since the smoothing apparatus is much the same for H.T. supply units for A.C. and D.C. supplies, details of the additional apparatus required in connection with A.C. eliminators are now given.

A.C. Mains. The voltage from alternating current mains varies continually from 325 volts positive to 325 volts negative 50 times per second in the case of a 230 volt 50 cycle supply.

50 ~ SUPPLY
(338)
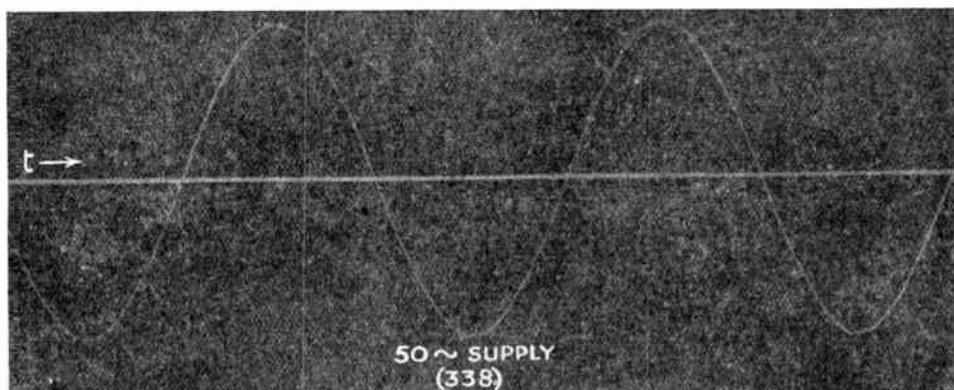
Fig. 116.

The oscillogram, Figure 116, shows such a current, which cannot be used for wireless sets until it is changed. The apparatus used for this purpose is a rectifier, which possesses the property of passing current in one direction only, or very much more on one direction than in the other. The two most satisfactory types are the Westinghouse Metal Rectifier, Figure 117, and the Valve Rectifier Figure 118.

In both cases a 4 mfd. condenser is connected across the rectifier output. The voltage and current obtained is of the wave form shown by the oscillogram, Figure 119, there being 100 peaks per second when operating from a 50 cycle supply.

Referring to Figure 118, which shows a valve rectifier with a mains transformer, the secondary is connected to the two plates of the valve rectifier (each plate containing a rectifier, so the valve is called a full wave rectifying valve),
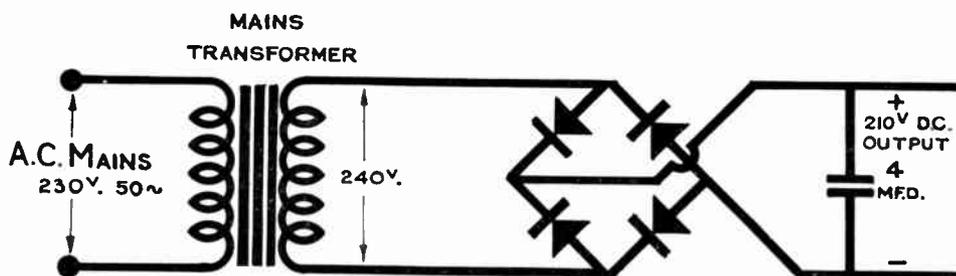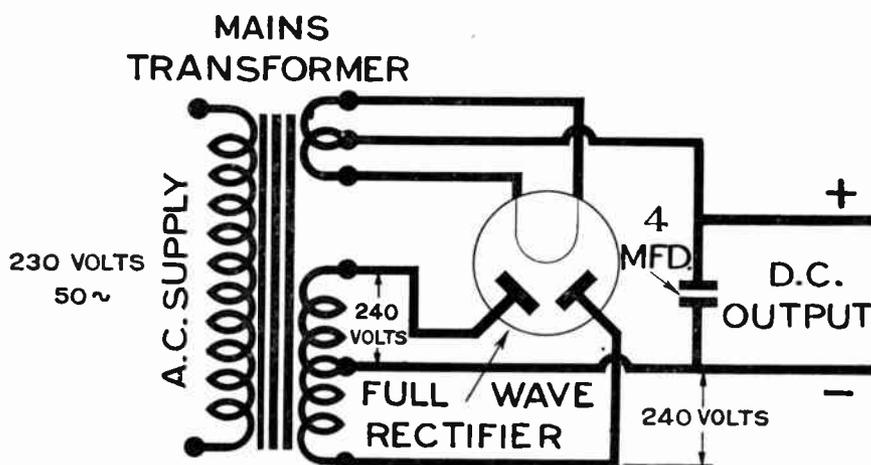
Fig. 117.

Fig. 118.

the output is of the form shown in oscillogram, Figure 119. Since each plate has the full voltage (240 volts in the Figure) impressed upon it, the secondary winding has, across its outer ends, 480 volts. This is the reason why we are reluctant to supply this transformer without the safety box, as apparatus with 400/480 volts or more should not be connected in a house, without making it absolutely safe.

In Figure 117 it will be seen that four separate metal rectifiers are employed and connected in such a manner that the transformer feeding them needs to give only 240 volts across its secondary winding.

50∿ SUPPLY, FULL WAVE RECTIFICATION.
57 m.A. MEAN VALUE
(344)
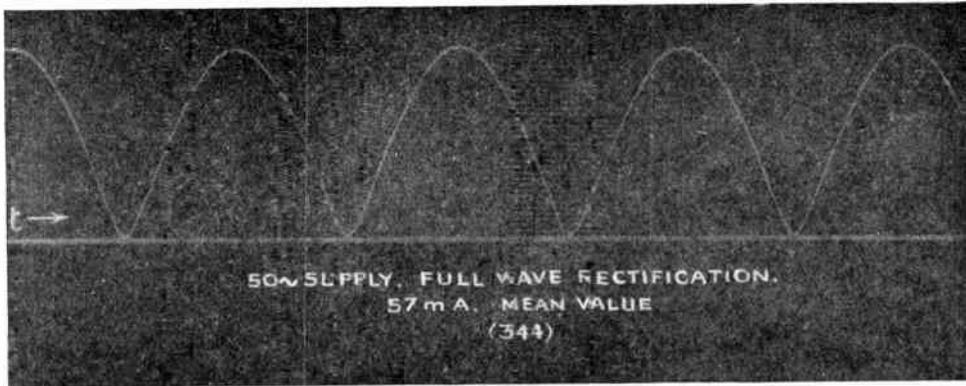
Fig. 119.

Comparison of Valve and Metal Rectifiers. The following are briefly the relative characteristics of the two types :—

WESTINGHOUSE METAL RECTIFIER.

1. More efficient than the valve rectifier.
2. Has no moving parts of filament to heat.
3. It requires a cheaper mains transformer.
4. For a given D.C. output voltage the input voltage need not be much greater. For example : an input voltage of 240 will usually enable an output voltage of 200 to be obtained.
5. For a given output, it is more expensive than the valve rectifier.
6. Its life is practically unlimited, no replacements being required.

VALVE RECTIFIER.

1. It is cheaper than the metal rectifier.
2. It requires a more expensive and elaborate type of transformer.
3. The voltage across the secondary of the mains transformer has to be about double that required for the same output from a metal rectifier.
4. Replacement is required at intervals.

It should be noted that metal rectifiers must be well ventilated to prevent any serious rise in temperature; the maximum temperature should not exceed 30°C. or 86°F.

These smoothers, or filters, are much the same whether they are connected to direct current mains or to rectified alternating current obtained from rectifiers of the form shown in Figures 117 and 118.

Fig. 119.

Fig. 120.

Their principles are now dealt with :—

Figures 119 and 120 show two common and simple forms of H.T. supply units. The chokes and condensers are arranged as shown, and across the smoothed D.C. output a high resistance with tappings is connected. The H.T. leads of the set are connected to these tappings A, B, and C, and across each tapping is a condenser of 1 or 2 mfds. The tappings are adjusted to give the desired voltages.

These arrangements have a serious defect. The valve plate currents have all to pass through some portion of the resistance, and, notwithstanding the by-pass condensers, there is still an impedance between A and B, A and C, and B and C ; in consequence, a small portion of the strong signal from the last stage of the receiver can pass by way of the battery connections from B and C to the earlier stages of the set and cause in effect audio frequency regeneration. If the regeneration or feed-back is great enough, oscillations of very low frequency are set up in the set and smoothing apparatus, giving the popping noise known as " Motor-*"Motor-* boating." Even if the feed-back is not sufficient to cause *boating."* self - oscillation, or " Motor - boating," it makes the amplification curve of the amplifier more uneven than it should be : this may be overcome to a great extent by the use of the anode feed system developed by us and previously described. When this system is used, the result is still not ideal, and therefore, after considerable experiment and measurement in our laboratories, we decided that satisfactory smoothing of H.T. current obtained from the Mains necessitates a separate feed, or filter system, for each stage of the receiver, the only exception being that the high frequency high tension tappings may be common.

CIRCUITS FOR H.T. SUPPLY UNITS.
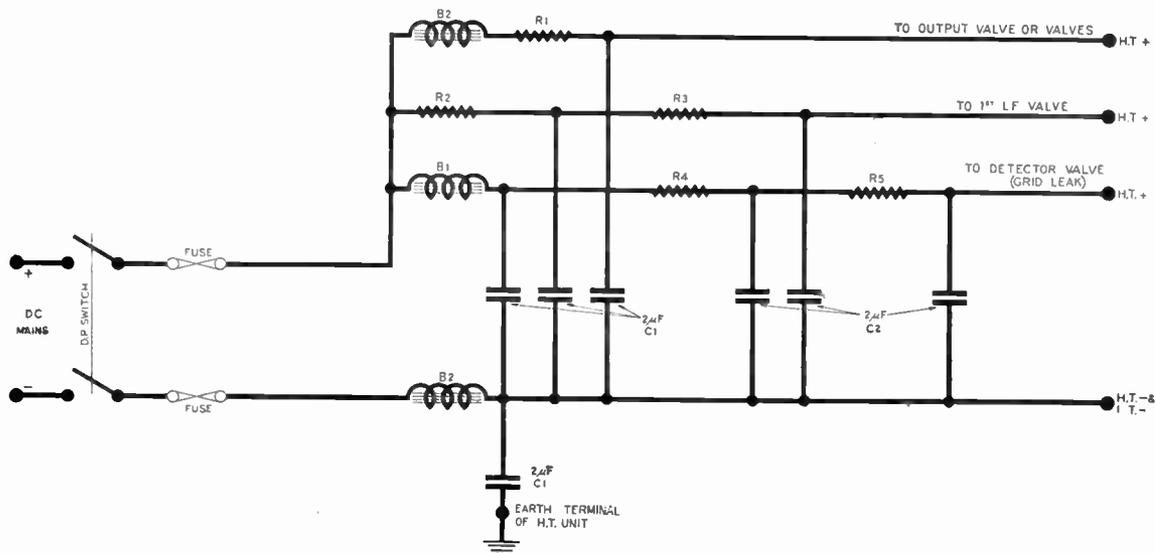
Fig. 121.

Figure 121 shows an H.T. supply unit for use on D.C. mains. This will supply a three-valve set with a detector valve using grid leak rectification and two audio frequency stages. The last stage can consist of the largest power valve or two such valves arranged in push-pull. The limit of output is 100 milliamps, through the chokes type " B.2." It will be observed that a type " B.2 " choke is connected in the negative lead to prevent ripples often present in the negative main from affecting the set.

Figure 122 shows an H.T. supply unit for use on A.C. mains. It is suitable for supplying a set using grid leak rectification and two audio frequency stages. A permanent full wave Westinghouse Metal Rectifier is employed. It is capable of supplying the plate current of the largest power valve or two such valves arranged in push-pull up to 200 volts, 100 milliamps.

Fig. 122.

Figure 123 shows a similar unit, but incorporating a valve rectifier instead of a metal rectifier :—

With a mains transformer, type EV2, and rectifying valve, type U8. Output 200 volts, 80 m/a.

With a mains transformer, type EV3, and rectifying valve, type U5 or R5. Output 120 volts, 20 m/a.
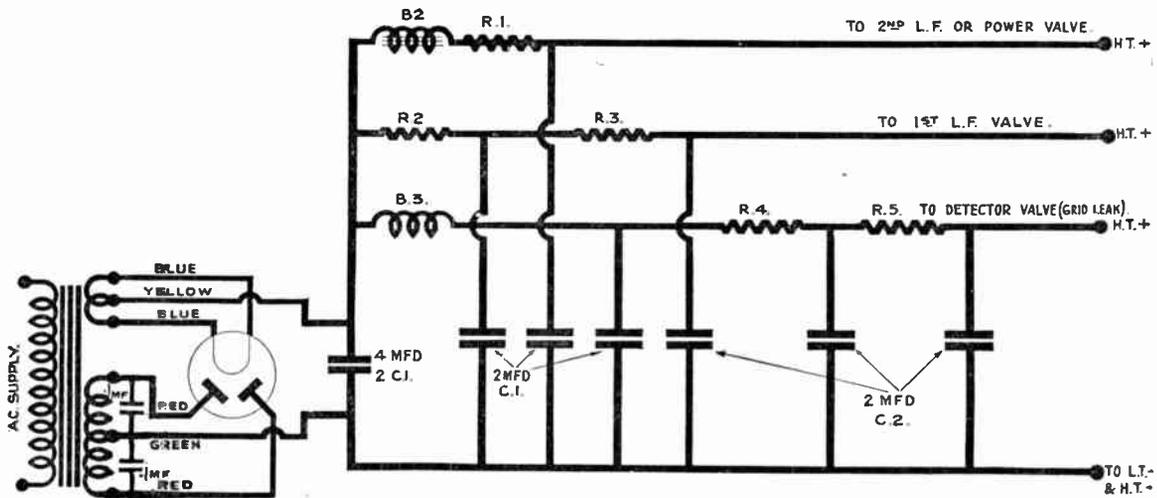
Fig. 123.

O
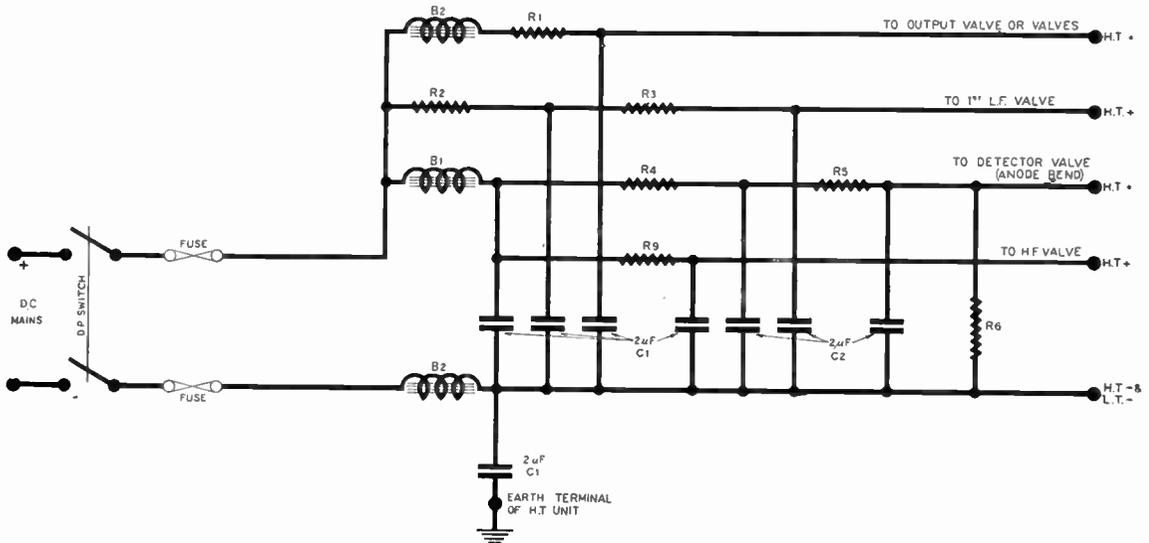
Fig. 124.

Figure 124 shows an H.T. supply unit for use on D.C. mains for supplying a set consisting of one ordinary high frequency transformer-coupled stage, an anode bend rectifier, and two audio frequency stages. The maximum output is 100 milliamps. through the " B.2 " chokes.

Figure 125 shows an H.T. supply unit for use on A.C. mains but for operating the same kind of receiver as the unit shown in Figure 124. The output is up to 200 volts, 100 milliamps., depending on the rectifier used.

Fig. 125.

Figure 126 shows a similar unit but incorporating a valve
rectifier instead of a metal rectifier :—

With a mains transformer, type EV2, and rectifying
valve, type U8 or R8. Output 200 volts 80 m/a.
With a mains transformer, type EV3, and rectifying
valve, type U5 or R5. Output 120 volts 20 m/a.
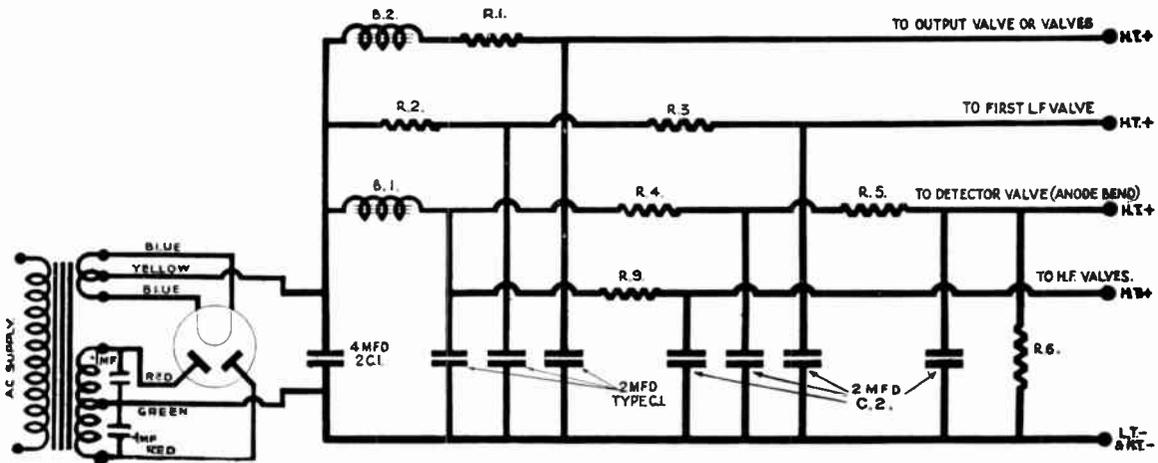
Fig. 126.
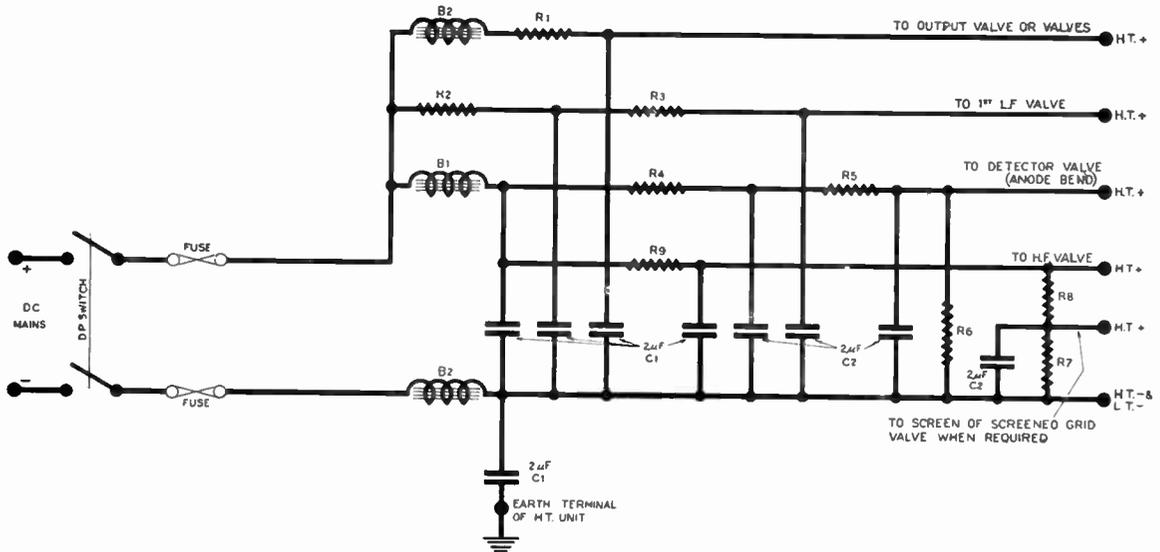
Fig. 127.

Figure 127 shows an H.T. supply unit for use on D.C. mains for supplying a set containing one or two stages of high frequency amplification using screened grid valves, an anode bend rectifier, and two stages of audio frequency amplification. An additional terminal is provided to give the necessary voltage for the screen of the H.F. valves, and, since this has to be fairly accurately adjusted, a potentiometer arrangement is used. The maximum output from this supply unit is 200 volts, 100 milliamps., through the " B2 " chokes.
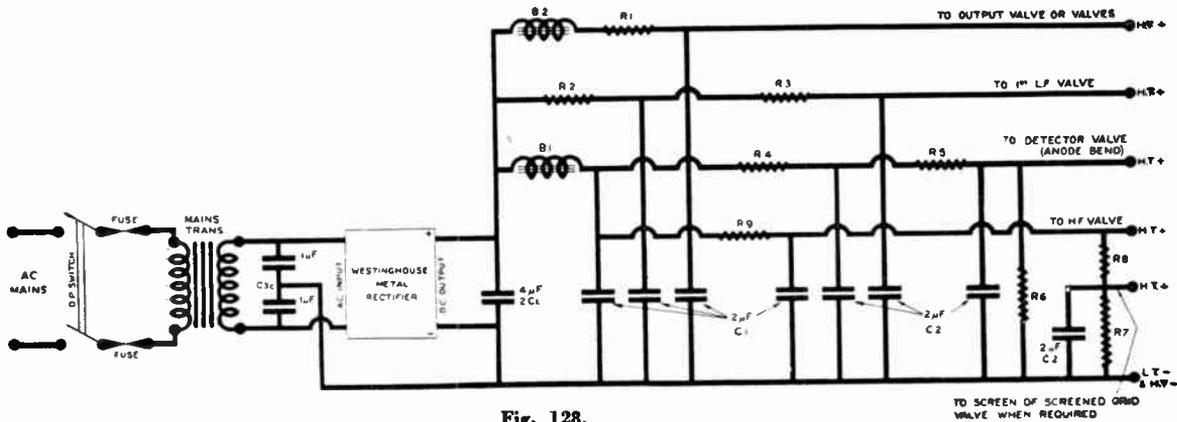
Fig. 128.

Figure 128 shows an H.T. supply unit for use on A.C. mains for the same purpose as the D.C. unit in Figure 127.
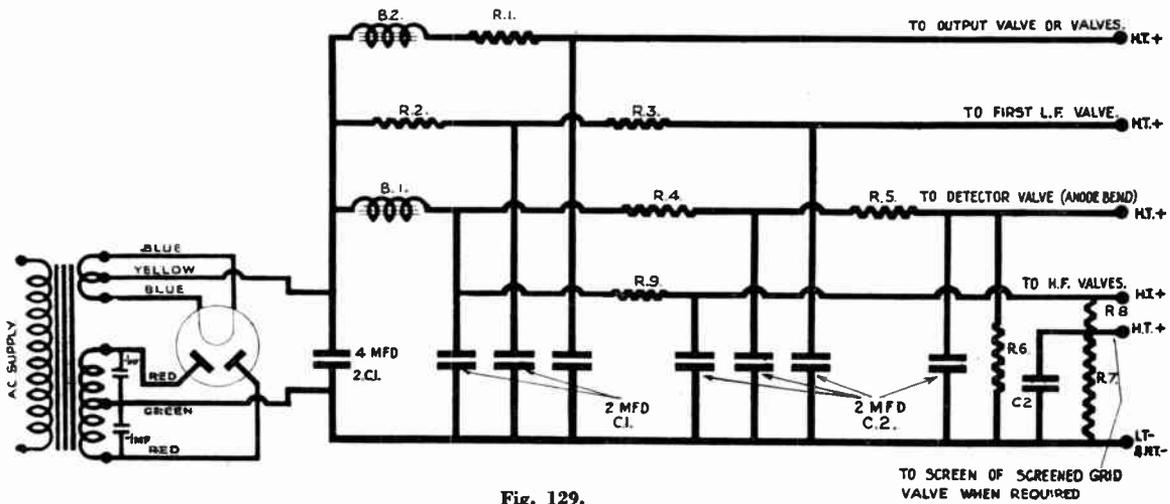
Fig. 129.

Figure 129 shows a similar unit but incorporating a valve rectifier instead of a metal rectifier :—

With a mains transformer, type EV2, and rectifying valve, type U8.

Output 200 volts 80 m/a.

The last section dealt wholly with the high tension supply for a receiver, but where A.C. mains are available all the supplies for a receiver can be obtained direct from the mains.

The introduction of indirectly heated valves enables one to use raw A.C. to heat the filaments. In the case of the last or power stage, because there is no amplification after the plate circuit of this stage, one is able to use raw A.C. as a means of heating the filament direct. There is this precaution required. The return for the grid and the H.T. circuit must be to a point on the mean potential of the filament. The windings on the mains transformer supplying the filaments is usually centre tapped to provide this point. If such a centre tapping is not available, a resistance of 40–50 ohms connected across the ends of the filament, with a connection brought out from its centre, will serve the same purpose.

Automatic Bias.

The grid bias voltage in such sets is usually obtained by inserting a suitable resistance in series with the cathode or centre point in the case of the power stage. The grid return is made to the bottom end of this resistance, and the flow of plate current through this resistance provides the necessary bias. This system has been given the name of "Automatic Bias."

The whole of the above has been shown carried out in the circuit diagram of A.C. Mains 32 (see Figure 11, page 29), and no further description is necessary here.

CALCULATIONS OF IRON CORED STRUCTURES WITH NECESSARY NUMERICAL DATA.
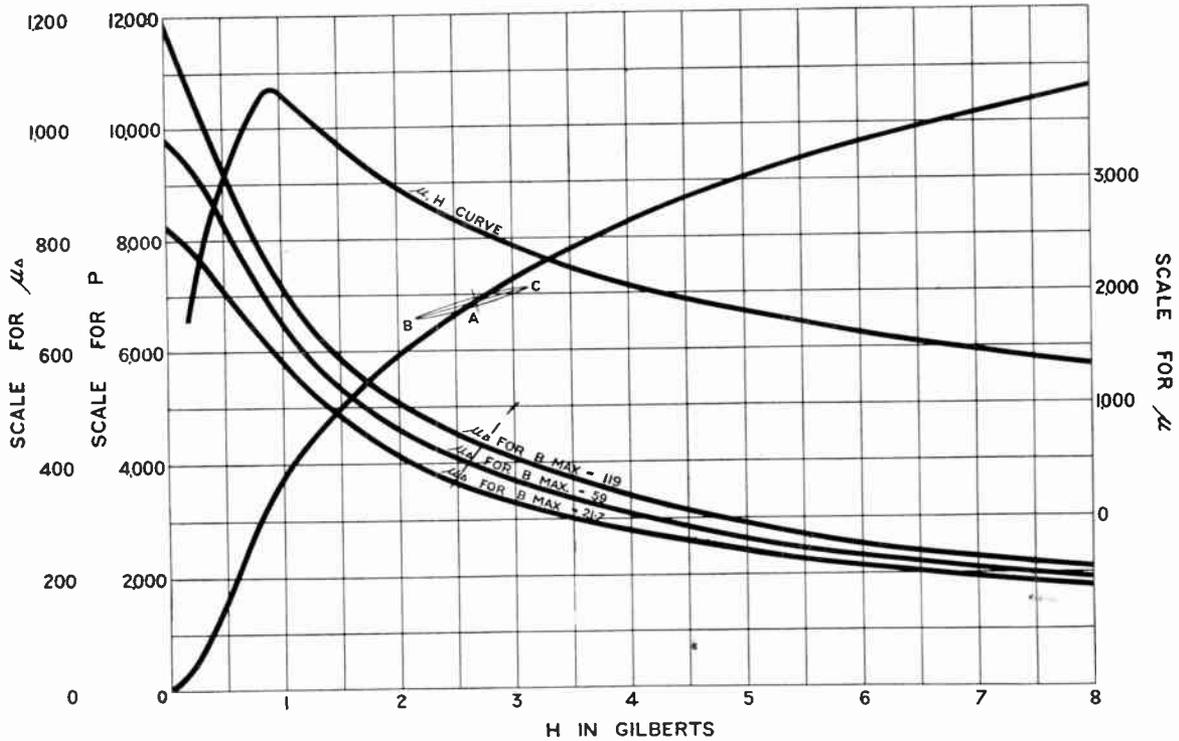
Fig. 130.

The curves of Figure 130, which have been obtained by laboratory measurements, supply the necessary data.

Curve I is the well known " B.H." curve for Armco, the iron used for A.F. transformers. H is the magnetising force in Gilbert's, due to a certain current flowing through a given number of turns round the core, per centimetre of mean magnetic length of core. B is the flux density in lines per square centimetre.

Curve II are μH curves, where $\mu = \dfrac{B}{H}$ for the same iron, and is called the *permeability*.

Curves III are curves of $\mu \Delta$, or what is known as incremental permeability. If a steady D.C. current passing through the windings of $H = 2 \cdot 7$, then the value of B is obtained by looking in the B.H. curve above $H = 2 \cdot 7$—say the point A—and is 6,900 lines per square centimetre.

Now, if the value of the current rises and falls a little—that is, if it is a D.C. current with a small A.C. superimposed—the change of flux caused does not take place, as might be imagined, along the slope of the B,H curve, but around a loop as A B C. The mean slope of this loop is the incremental permeability for this particular value of H. The curves have been obtained by measuring the inductance under specified conditions and calculating back.

FORMULÆ :—

$$H = \frac{1 \cdot 257 \times \text{current in amperes} \times \text{turns}}{1} \text{ gilberts per centimetre}$$

where l is the mean magnetic length in centimetres. When *the magnetic path is wholly of iron,* and is of constant cross section, then l is just the mean length of the core.

Where the magnetic patch is wholly of iron but of different cross sections, then

$$1 = \left[\frac{l_1}{A_1} + \frac{l_2}{A_2} + \text{ etc.} \right] A$$

where A is the area in sq. cms., to which we wish to refer the flux density.

If the path contains air gaps, the mean magnetic length is not easy to find.

$1 = l_2 + \left[\dfrac{l_a}{A_a} \mu \right] A$ where l_2 is the mean magnetic length of the iron as calculated above and $l_a = $ length of the air gap, Aa its area, and μ the permeability of the iron under

the existing conditions for H. The difficulty is to find μ, since H is not yet known. It is a matter of trial and error. The inductance L henries

$$= \frac{1 \cdot 257 \times \text{Nett core area} \times (\text{turns})^2 \, \mu_\Delta}{10^8 \left[l_2 + \frac{l_a}{A_a} A \, \mu_\Delta \right]}$$

the nett core area being in square cms., and μ_Δ being obtained from the published curves.

L is the property of the transformer which we usually set out to calculate and involves first finding H and μ_Δ.

Variation of μ_Δ:
therefore of
inductance with
strength of signal.

But μ_Δ, the incremental or A.C. permeability, varies with the strength of the A.C. flux. We give three sets of curves for three values of A.C. flux.

The latter is defined by the B max. or lines per sq. cm. corresponding to the peak value of the A.C. current.

$$\text{B max.} = \frac{\text{A.C. volts (R. M. S.) across the coil} \times 10^8}{4 \cdot 44 \times \text{frequency} \times \text{core area} \times \text{turns}}$$

core area in sq. cm.

The general reader will see that calculations on iron core structures are not simple, and the formulæ have been included for the technical reader. There is one phase which the general reader must consider—the value B max. for the transformers used in a receiver, under normal working conditions.

First, from the last formula, it will be noticed that it depends on the frequency of the applied note, being greatest at the lowest frequency; hence, one must consider it at the lowest note to be properly reproduced. We usually take this to be 50 cycles.

Now, if the A.C. flux is too great, the sloping loop A B C (Figure 130) is not reasonably straight, and the μ_Δ varies during one period of the A.C. fluctuation. Hence, the

output under these circumstances, could not be as pure in wave form as the input. However, if the A.C. flux B max. is kept low enough, this effect is negligible.

The Tables below give values for each stage of a very powerful receiver used for cinema amplification, the last stage being two LS6A valves in push-pull, with 400 volts on their plates.

We have considered two cases:—

1st: When the receiver is just giving the mean level of sound volume :—19 volts across the primary of the output transformer.

2nd : When the last valves are just being overloaded during the loud passages of music. This means 190 volts across the primary of the output transformer.

(1) MEAN LEVEL OF SOUND (19 VOLTS).
$B_{max.}$ lines per sq. cm.

Frequency of Note Cycles.	In AF5, 1st A.F. Transformer.	2nd AF5c.	In Standard Output Transformer.	In Large Output Transformer, OPcXX.
50	0·62	34	340	200
150	0·21	11·3	113	67

(2) JUST OVERLOADING (190 VOLTS).
$B_{max.}$ lines per sq. cm.

Frequency of Note Cycles.	In AF5, 1st AF Transformer.	2nd AF5c.	In Standard Output Transformer.	In OPcXX.
50	6·2	340	3,400	2,000
150	2·1	113	1,130	670

There is no doubt that if one ever got a voltage on the primary of the standard output transformer of 190 volts for a 50 *cycle note* (a rare and transient occurrence) the flux, 3,400 lines per sq. cm., would give a somewhat distorted 50 note. We prefer, therefore, to use the larger output transformer OPCXX in such a receiver. Of course, in a receiver with a smaller power stage, the A.C. flux densities will be less. In any case, it is only the output transformer that need be considered.

WARNING.

As shown, the inductance of a given structure depends on the $B_{max.}$—that is, on the volts across it when measured. The greater this voltage the greater the inductance, though the variation is not great for the low flux usual in receivers.

The inductances of all our components are measured with $3\frac{1}{2}$ volts across (corresponding to 32 lines per sq. cm. in an AF5), which give flux densities well within the range occurring in receivers. But components are offered to the public for which large inductances are claimed, undoubtedly measured at large voltages, and so are misleading as to their use in the receivers.

BATTERY SET.

For use with H.T. accumulators : 120 volts.

Two stages, Detector, and power stage of two valves, class A, in push pull.
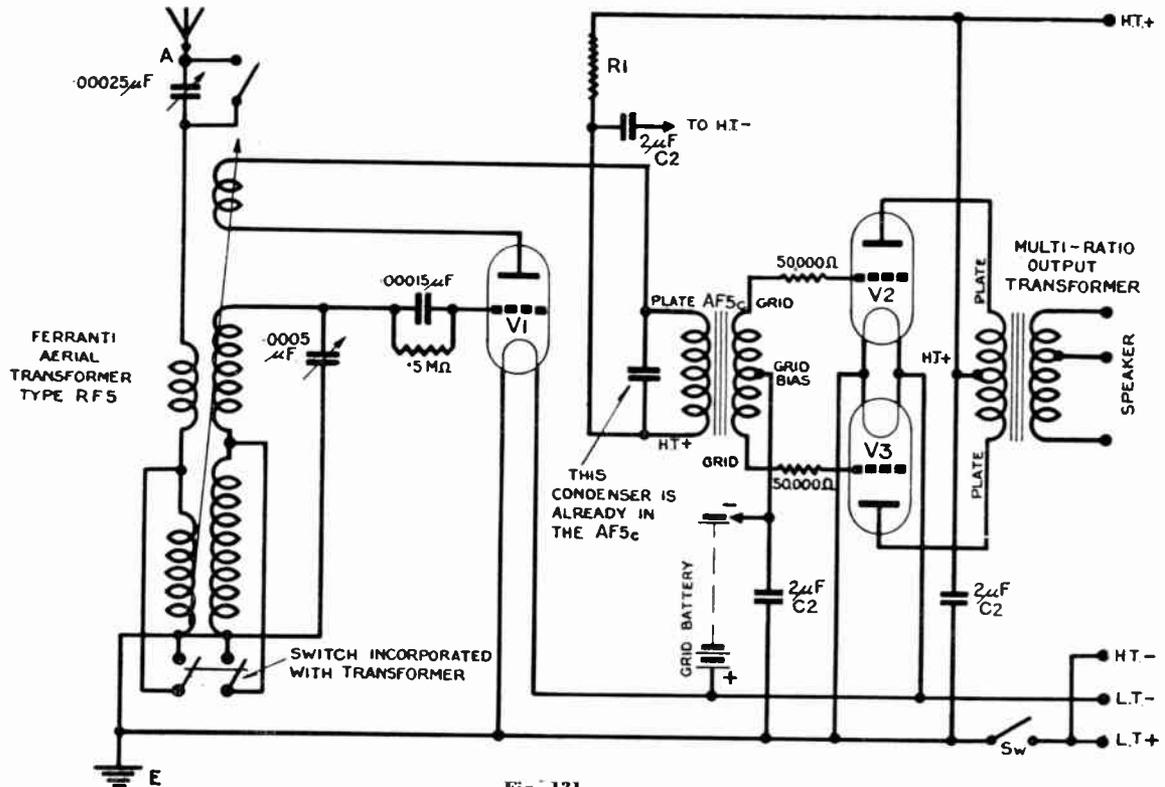
Fig. 131.

				Grid bias.	Milliamps, approx.	
V2	LP2. Marconi Osram	... ...	$-4\frac{1}{2}$ v. ...		12	
and	220P. Ediswan	... ...	$-4\frac{1}{2}$ v. ...		14	
V3	PM2A. Mullard	... ...	$-4\frac{1}{2}$ v. ...		13	
either	220P. Cossor	... ...	$-7\frac{1}{2}$ v. ...		14	

For balance one must only pair valves of the same make.

					R1.
V1 either	L210. Marconi Osram	$+2$ ...	2	... 60,000 ohms.
	L210. Ediswan	$+2$ ...	2	... 60,000 ohms.
	PM1.LF., Mullard	$+2$ ...	2	... 60,000 ohms.
	210.LF. Cossor	$+2$ ...	2	... 60,000 ohms.

A.C. MAINS SET.

Two valves.
Power stage : 1 valve Class C.

Fig. 132.

Typical set of valves and values of resistances :

V2 ... P625 Marconi Osram.
V1 ... MH4 Marconi Osram.
VR ... U10 Marconi Osram.

The mains transformer must be capable of giving 28 milliamps and 260 volts D.C. output when smoothed.

Values of the resistances :—

R4 = R5	R22.	R23.	R24.	R25.
40,000 ohms.	100,000 ohms.	1,100 ohms.	750 ohms.	100,000 ohms.

The power valve will overload when a signal of $17\frac{1}{2}$ volts R.M.S. is applied to the grid.

The maximum audio signal across the grid of V1 (for gramophone pickup) $= \cdot146$ volt.

The radio frequency signal at the grid of V_1, when modulated 80% to overload the last valve $= \cdot29$ volts

BATTERY SET.

THREE VALVE SET.

H.T. power available, 120 volts, dry batteries.

Power stage valve class A.
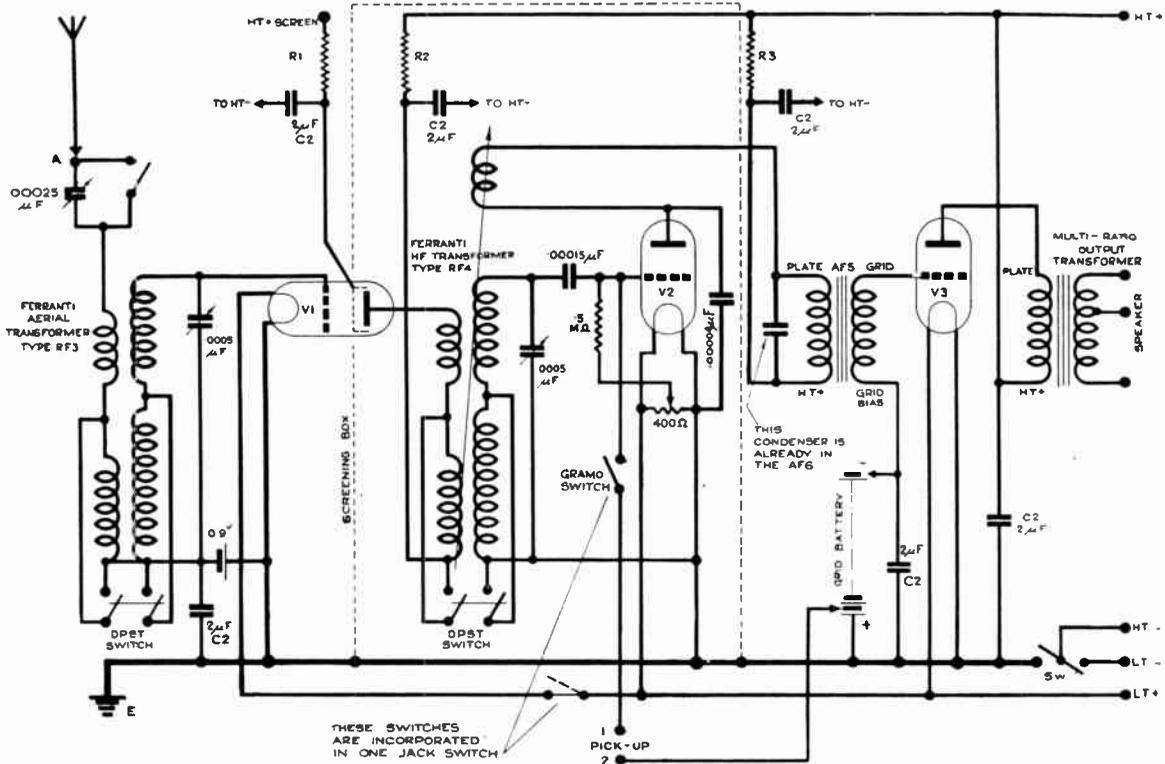
Fig. 133.

		Grid bias.	Milliamps, approx.
V3 either	LP2. Osram Marconi	$-4\frac{1}{2}$ v. ...	6
	220P. Ediswan	$-4\frac{1}{2}$ v. ...	7
	PM2A. Mullard	$-4\frac{1}{2}$ v. ...	$6\frac{1}{2}$
	220P. Cossor	$-7\frac{1}{2}$ v. ...	7

									R3.
L210. Osram Marconi	...	...	...	...	+2	...	2	...	60,000 ohms.
P210. Ediswan ...	...	...	...	...	+2	...	2	...	60,000 homs.
PM1.LF. Mullard	...	...	...	...	+2	...	2	...	60,000 homs.
210. LF., Cossor	...	...	...	...	+2	...	2	...	60,000 ohms.

V2 either

		Screen.						R2.
	V.	R1.						
S.215. Osram Marconi	75	10,000	...	0	...	2½	...	2,000 ohms.
215.SG. Ediswan ...	75	10,000	...	0	...	3	...	2,000 ohms.
PM.12. Mullard	75	10,000	...	0	...	2½	...	2,000 ohms.
215.SG. Cossor	75	10,000	...	0	...	3½	...	2,000 ohms.

V1 either

In each set, approximate total milliamps required from battery : 12.

BATTERY SET.

For use with H.T. accumulators : 120 volts.

THREE VALVE SET.

Power stage, two power valves, class A, in push-pull.
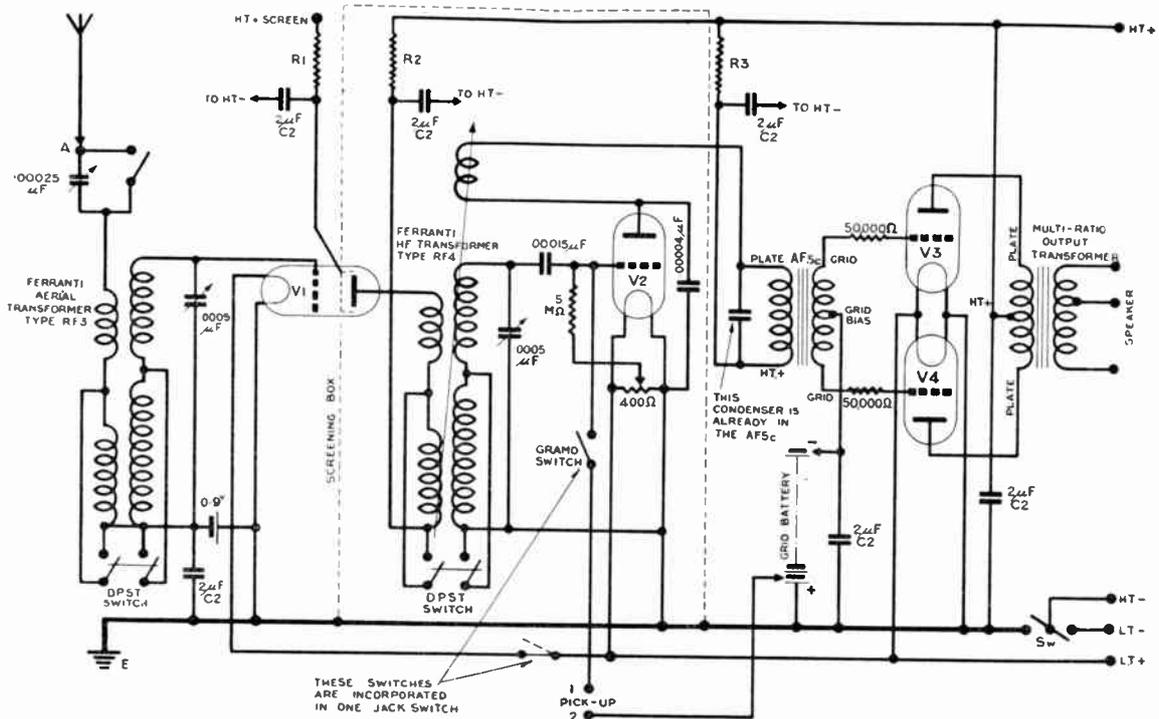
Fig. 134.

		Grid bias.	Milliamps, approx.
V3 and V4 either	LP2. Marconi Osram	$-4\frac{1}{2}$ v.	...6 ×2=12
	P220. Ediswan Mazda	$-4\frac{1}{2}$ v.	...7 ×2=14
	PM2A. Mullard	$-4\frac{1}{2}$ v.	...$6\frac{1}{2}$×2=13
	220.P. Cossor	$-4\frac{1}{2}$ v.	...7 ×2=14

For balance one must only pair valves of the same make.

				R3.
V2 either	L210. Marconi Osram	+2 ...	2	... 60,000 ohms.
	L210. Ediswan Mazda	+2 ...	2	... 60,000 ohms.
	PM1.LF. Mullard	+2 ...	2	... 60,000 ohms.
	210.LF. Cossor	+2 ...	2	... 60,000 ohms.

		Screen.					R2.
		V.	R1.				
V1 either	S215. Marconi Osram	75	10,000 ...	0 ...	$2\frac{1}{2}$	... 2,000 ohms.
	215.SG. Ediswan Mazda	75	10,000 ...	0 ...	3	... 2,000 ohms.
	PM12. Mullard	75	10,000 ...	0 ...	$2\frac{1}{2}$	... 2,000 ohms.
	215.S.G. Cossor ...	75	10,000 ...	0 ...	$3\frac{1}{2}$	... 2,000 ohms.

In each set, approximate total milliamps required from battery = 18.

P

A.C. MAINS SET.

H.F., Detector and push-pull output.

Power stage : Two power valves, class C, in push-pull.
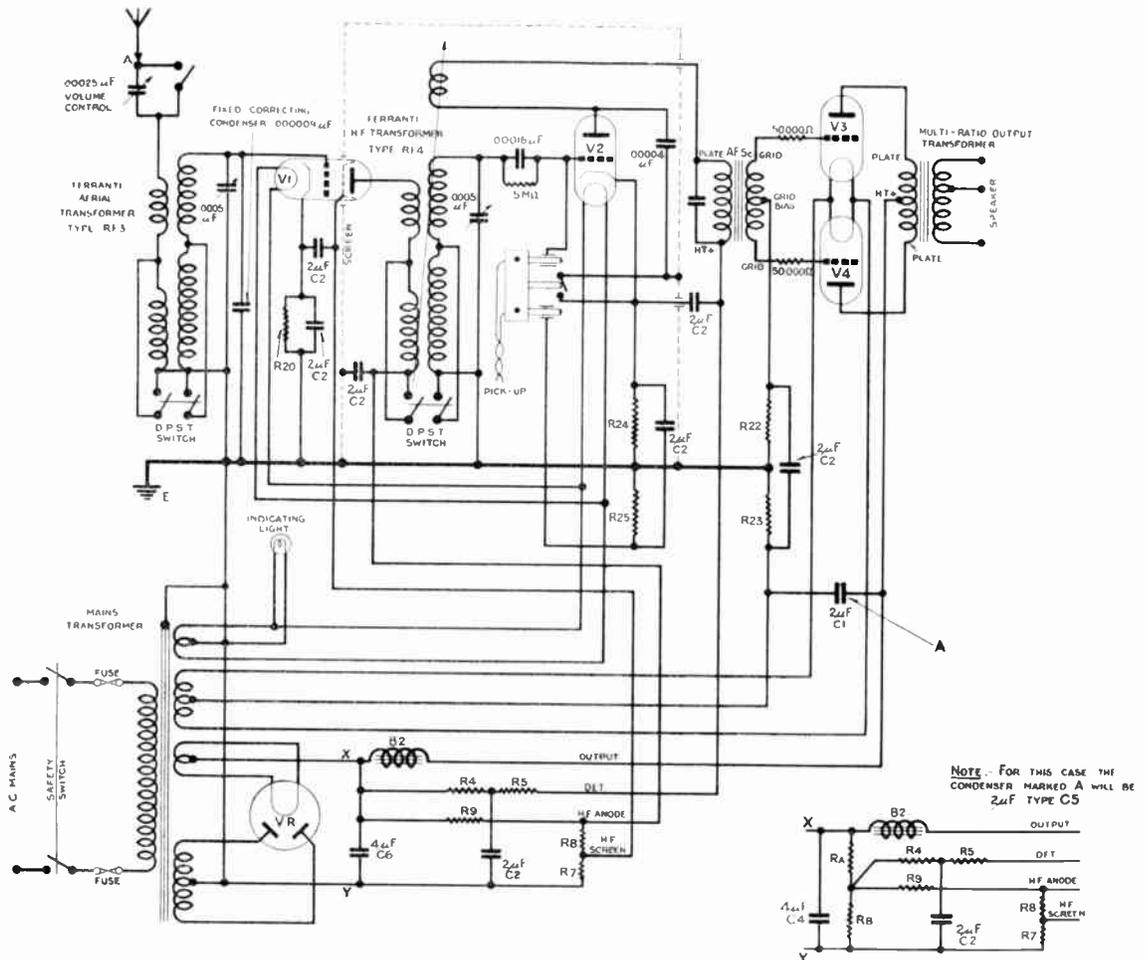
Fig. 135.

Typical set of valves and values of resistances :

V3 and V4	...	P625's Marconi Osram.
V2	... ...	MH4 Marconi Osram.
V1	... ...	S4VA Mullard.
VR	... ...	U10 Marconi Osram.

The mains transformer must be capable of giving 55 milliamps and 260 volts D.C. output when smoothed.

Values of the Resistances :—

R4 = R5	R7.	R8.	R9.	R20.	R24.	R25.	R22.	R23.	
40,000	60,000	40,000	30,000	1,100	750	100,000	100,000	550	ohms.

The power valves will just overload when a signal of 35 volts R.M.S. is applied across the grids.

The maximum audio signal across the grid of the detector V_2 (for gramophone pickup) $= \dfrac{35}{120} = \cdot292$ volt, since the amplification of MH4 and AF5c $= 120$. Using the figures of the section on grid leak rectification, we find if the R.F. input to V_2 is $\cdot58$ volt, the overloading will occur when the modulation at the station is 80%. This will be the normal signal used to give moderate loudness.

Such a signal will give perfect results with grid leak rectification and the one stage of H.F. amplification will be sufficient, except for very weak stations.

SCREENED GRID H.F. VALVES.

TYPE OF VALVE AND MAKER.	Filament.		Grid Bias. Volts.	Anode.		Screen Volts.	Screen Current. mA.	Nominal Impedance. Ohms.	Amplification Factor.	Mutual Conductance : mA. per Volt.
	Volts.	Amps.		Volts.	Current. mA.					
MARCONI OSRAM :										
S.215	2	·15	0	120	2·7	80	1·0	300,000	180	·6
				150	2·8	80				1,000
S.22	2	·2	0	120	3·9	75	1·5	200,000	350	1·75
				150	4·1	75				1,000
S.410	4	·10	0	120	4·6	80	1·0	200,000	180	·9
				150	4·7	80				1,000
S.610	6	·10	0	120	5·7	80	1·0	200,000	210	1·05
				150	5·8	80				1,000
MULLARD :										
PM.12	2	·15	0	120	2·7	75	1·0	212,000	200	·94
				150	2·8	75				1,000
PM.14	4	·075	0	120	2·7	75	1·0	230,000	200	·87
				150	2·8	75				1,000
PM.16	6	·075	0	120	3·0	75	1·0	200,000	200	1·0
				150	3·2	75				1,000
EDISWAN (MAZDA) :										
215.SG.	2	·15	0	120	3·3	70	1·0	270,000	300	1·1
				150	3·4	70	1·0	400,000	450	1,000
COSSOR :										
215.SG	2	·15	0	120	3·6	75	1·0	300,000	330	1·1
				150	3·7	75				1,000
220.SG	2	·2	0	150	3·1	75	1·0	200,000	320	1·6
										1,000
410.SG.	4	·1	0	120	4·1	75	1·0	200,000	200	1
				150	4·2	75				1,000
610.SG.	6	·1	0	120	4·2	75	1·0	200,000	200	1
				150	4·3	75				1,000

The above are Battery Heated Valves. Since grid current flows only when grid is ½ to 1 Volt positive, grid bias has been chosen as Zero for simplicity.

SCREENED GRID A.C. HEATED VALVES.

TYPE OF VALVE AND MAKER.	Filament.		Grid Bias. Volts.	Resis- tance. R20 Ohms.	Anode.		Screen Volts.	Screen Current. mA.	Nominal Im- pedance. Ohms.	Ampli- fica- tion. Factor.	Mutual Conduc- tance : mA. per Volt.
	Volts.	Amps.			Volts.	Current. mA.					
MARCONI OSRAM :											
MS4	4	1·0	−1½	520	150	2·4	60	·4	500,000	550	$\dfrac{1\cdot1}{1,000}$
MULLARD :											
S4V	4	1·0	−1·2	800	150	1·25	75	—	909,000	1,000	$\dfrac{1\cdot1}{1,000}$
S4VA... ...	4	1·0	−1·0	700	150	1·5	75	—	430,000	1,500	$\dfrac{3\cdot5}{1,000}$
S4VB... ...	4	1·0	−1·2	250	150	5·0	75	—	257,000	900	$\dfrac{3\cdot5}{1,000}$
EDISWAN (MAZDA) :											
AC/SG ...	4	1·0	−1·5	240	150	6·2	70	—	390,000	900	$\dfrac{2\cdot3}{1,000}$
COSSOR :											
41MSG ...	4	1·0	−1·5	—	150	·5	60	—	400,000	1,000	$\dfrac{2\cdot5}{1,000}$

DETECTOR VALVES FOR ANODE BEND RECTIFICATION.

TYPE OF VALVE AND MAKER.	Filament.		Grid Bias Volts at given Anode Volts.	Anode Volts.	Maximum Anode Current with Strong Signal. mA.	Nominal Impedance. Ohms.	Nominal Amplifi- cation Factor.
	Volts.	Amps.					
MARCONI OSRAM :							
L.210	2	·1	−9	120	·9	12,000	11·0
L.410	4	·1	−7½	120	1·0	8,500	15·0
L.610	6	·1	−7½	120	1·2	7,500	15·0
MULLARD :							
PM.1LF	2	·1	−10½	120	1·0	12,000	11·0
PM.4DX	4	·1	−7½	120	1·0	7,500	15·0
PM.6D	6	·1	−7½	120	1·0	9,000	18·0
EDISWAN (MAZDA) :							
L.210	2	·1	−6	120	1·0	10,000	15·5
COSSOR :							
210.LF	2	·1	−9	120	·9	12,000	10·0
410.LF	4	·1	−6	120	1·0	8,500	15·0
610.LF	6	·1	−6	120	1·0	7,500	15·0

DETECTOR VALVES FOR GRID LEAK RECTIFICATION.

Grid Condenser ·00015 mfd.   Grid Leak 500,000 Ohms.

TYPE OF VALVE AND MAKER.	Filament.		Grid Bias at given Anode Volts.	Anode.		Nominal Impedance. Ohms.	Amplification Factor.
	Volts.	Amps.		Volts.	Current. mA.		
MARCONI OSRAM :							
L.210	2	·1	+2	60	3·0	12,000	11·0
L.410	4	·1	+2	60	2·8	8,500	15·0
L.610	6	·1	+2	60	2·7	7,500	15·0
MULLARD :							
PM.1LF	2	·1	+2	60	3·0	12,000	11·0
PM.4DX	4	·1	+2	60	2·8	7,500	15·0
PM.6D	6	·1	+2	60	2·9	9,000	18·0
EDISWAN :							
L.210	2	·1	+2	60	3·0	10,000	15·5
COSSOR :							
210.LF	2	·1	+2	60	2·2	12,000	10·0
410.LF	4	·1	+2	60	2·0	8,500	15·0
610.LF	6	·1	+2	60	2·0	7,500	15·0

A.C. HEATED VALVES.

MARCONI OSRAM :							
MH4	4	1·0	0	70	2·4	16,000	35·0
MHL4	4	1·0	0	60	3·1	8,000	20·0
MULLARD :							
354V	4	1·0	0	70	3·2	11,700	35·0
164V	4	1·0	0	50	4·0	6,650	16·0
EDISWAN (MAZDA) :							
AC/HL	4	1·0	0	60	2·7	11,700	35·0
COSSOR :							
41.MHF.	4	1·0	0	60	2·3	14,000	32·0

FIRST L.F. VALVES.

TYPE OF VALVE AND MAKER.	Filament.		Grid Bias. Volts.	Anode.		Nominal Impedance. Ohms.	Amplifi- cation Factor.
	Volts.	Amps.		Volts.	Current. mA.		
MARCONI OSRAM :							
L.210	2	·1	−3	100	3·0	12,000	11·0
L.410	4	·1	−1½	100	3·5	8,500	15·0
L.610	6	·1	−1½	100	3·4	7,500	15·0
MULLARD :							
PM.1LF	2	·1	−3	100	3·2	12,000	11·0
PM.4DX	4	·1	−1½	100	3·0	7,500	15·0
PM.6D	6	·1	−1½	100	3·0	9,000	18·0
COSSOR :							
210.LF	2	·1	−3	100	2·6	12,000	10·0
410.LF	4	·1	−1½	100	2·7	8,500	15·0
610.LF	6	·1	−1½	100	3·0	7,500	15·0
EDISWAN (MAZDA) :							
L210	2	·1	−1½	100	3·5	10,000	15·5

A.C. HEATED VALVES.

MARCONI OSRAM :							
MHL4	4	1	−1½	100	5·0	8,000	20·0
MULLARD :							
164V	4	1	−3	100	5·0	6,650	16·0
EDISWAN (MAZDA) :							
AC/HL	4	1	−1½	120	3·5	11,700	35·05
COSSOR :				−3½	100	4·0		
41.MLF	4	1	−4	120	4·5	7,900	15·0

SMALL POWER VALVES. CLASS A.

TYPE OF VALVE AND MAKER.	Filament.		Anode.		Grid Bias.	Nominal Im-pedance.	Amplifi-cation Factor.	Mutual Con-ductance.	Grid Swing
	Volts.	Current. Amps.	Volts.	Current. mA.	Volts.	Ohms.		mA. per Volt.	Volts. (RMS).
OSRAM :									
P.215 ...	2·0	·15	120	6·5	−9	5,000	7·0	1·4	6·25
P.410 ...	4·0	·1	120	6·0	−9	5,000	7·5	1·5	6·25
P.610 ...	6·0	·1	120	5·5	−7½	3,500	8·0	2·3	5·25
COSSOR :									
220.PA ...	2·0	·2	120	5·0	−4½	4,000	16·0	4·0	3·25
215.P ...	2·0	·15	120	7·0	−6	4,000	9·0	2·25	4·25
220.P ...	2·0	·2	120	6·0	−7½	4,000	8·0	2·0	5·25
410.P ...	4·0	·1	120	7·0	−7½	4,000	8·0	2·0	5·25
610.P ...	6·0	·1	120	7·5	−6	3,500	8·0	2·28	4·25
MULLARD :									
PM.2 ...	2·0	·2	120	5·0	−9	4,400	7·5	1·7	6·25
PM.2A ...	2·0	·2	120	6·0	−4½	3,600	12·5	3·5	3·25
PM.4 ...	4·0	·1	120	5·0	−9	4,450	8·0	1·8	6·25
PM.6 ...	6·0	·1	120	6·5	−7½	3,550	8·0	2·25	5·25
MAZDA :									
P.220 ...	2·0	·2	120	3·0	−6	3,700	12·5	3·4	4·25

The small power valves listed above are suitable, because of their small anode current, for use with dry H.T. batteries, and the figures are given for 120 volts, which is that most generally used.

MEDIUM POWER VALVES.   CLASS B.

TYPE OF VALVE AND MAKER.	Filament.		Anode.		Grid Bias.	Nominal Im-pedance. Ohms.	Amplifi-cation Factor.	Mutual Con-ductance. mA. per Volt.	Grid Swing Volts (R.M.S).
	Volts.	Current. Amps.	Volts.	Current. mA.	Volts.				
OSRAM :									
P.240 ...	2·0	·4	150	17·0	−24	2,500	4·0	1·6	17·0
LP.2 ...	2·0	·2	150	11·5	−4½	3,900	15·0	3·85	3·25
P.2 ...	2·0	·2	150	17·0	−10½	2,150	7·5	3·5	7·5
P.425 ...	4·0	·25	150	17·0	−16½	2,300	4·5	1·95	11·75
COSSOR :									
230.XP ...	2·0	·3	150	22·0	−18	1,500	4·5	3·0	12·75
415.XP ...	4·0	·15	150	22·0	−18	1,500	4·5	3·0	12·75
425.XP ...	4·0	·25	150	11·0	−13½	2,000	7·0	3·5	9·5
610.XP ...	6·0	·1	150	23·0	−15	2,000	5·0	2·5	10·5
625.P ...	6·0	·25	150	14·0	−9	2,500	7·0	2·8	6·25
MULLARD :									
PM.252 ...	2·0	·3	150	16·5	−15	2,600	5·4	2·1	10·5
PM.202 ...	2·0	·2	150	14·0	−13½	2,000	7·0	3·5	9·5
PM.254 ...	4·0	·18	150	14·0	−22½	2,000	4·2	2·1	16·0
PM.256 ...	6·0	·25	150	12·0	−12½	1,850	6·0	3·25	8·75
MAZDA :									
P.220 A...	2·0	·2	150	12·0	−15	1,850	6·5	3·5	10·5
P.240 ...	2·0	·4	150	11·0	−13½	1,900	7·0	3·7	9·5
P.425 ...	4·0	·25	150	13·5	−27	1,950	3·5	1·8	19·0
P.625 A...	6·0	·25	150	15·0	−20	1,600	4·0	2·5	14·0
P.625 B...	6·0	·25	150	13·5	−10½	2,500	7·0	2·8	7·5

VALVES FOR A.C. MAINS.

	Volts.	Current. Amps.	Volts.	Current. mA.	Volts.	Ohms.	Factor.	mA. per Volt.	(R.M.S).
MULLARD :									
104.V ...	4·0	1·0	200	17·0	−12½	2,850	10·0	3·5	9·0
MAZDA : ...									
A/CP ...	4·0	1·0	200	13·0	−15	2,650	10·0	3·75	10·5
AC/P1 ...	4·0	1·0	200	16·0	−30	2,000	5·0	2·5	21·0

The medium power valves listed above, having a greater anode current than Class A, are more suitable for use with H.T. accumulators, and the figures are given for 150 volts, which is that generally advised.

LARGE POWER VALVES.   CLASS C, D, E, and F.

TYPE OF VALVE AND MAKER.	Filament.		Anode.		Grid Bias.	Nominal Impedance. Ohms.	Amplification Factor.	Mutual Conductance. mA. per Volt.	Grid Swing Volts (R.M.S).
	Volts.	Current. Amps.	Volts.	Current. mA.	Volts.				
OSRAM :									
PX4 ...	4·0	·6	200	50	−33	1,050	3·5	3·3	23
			150	37	−23				16
PX4 ...	4·0	1·0	300	40	−48	830	5·0	6·0	34
			200	40	−27				19
P625 ...	6·0	·25	250	24	−26	2,400	6·0	2·5	18
			200	19	−20				14
P625A ...	6·0	·25	200	25	−39	1,600	3·7	2·3	27
			150	19	−26				18
LS5A ...	5·25	·8	400	33	−112	2,750	2·5	·91	79
			300	25	−80				56
			200	16	−50				35
LS6A ...	6·0	2·0	400	63	−91	1,300	3·0	2·3	64
			300	55	−60				42
			200	40	−34				24
COSSOR :									
41.MP ...	4·0	1·0	200	24	−7½	2,500	18·7	7·5	5·25
41.MXP	4·0	1·0	200	40	−12½	1,500	11·2	7·5	8·75
MULLARD :									
PM.256A	6·0	·25	200	32	−33	1,400	3·6	2·6	23
MAZDA :									
PP5/400	4·0	2·0	400	63	−32	1,500	9·0	6·0	22
P650 ...	6·0	·5	200	23·5	−40	1,300	3·5	2·7	28

The large power valves listed above, because of their much greater anode voltage and anode current, usually receive their H.T. supply from the mains.

ELECTRICAL TERMS.

As this book may come into the hands of those with little or no knowledge of the meaning of the various terms used, we give below brief definitions of the more common ones :—

Coulomb. The Unit of Quantity of Electricity.

Ampere. The Rate of Flow of Electricity. One ampere equals one Coulomb per second.

Milliampere. 1/1000th of an ampere.

Volt. The Unit of Electrical Pressure.

Ohm. The Unit of Resistance. All materials offer resistance to the flow of Electricity through them. In the case of a wire, the resistance is proportional to the length and inversely proportional to the cross section. Different materials have different resistances, copper having the lowest, except silver.

Current (Amperes), Pressure (Volts), and Resistance (Ohms) are connected together by the very simple formula :—

$$\text{Amperes} = \frac{\text{Volts}}{\text{Ohms},} \text{ usually written } I = \frac{E}{R},$$

so that, knowing any two quantities, the other may be calculated.

Megohm. 1,000,000 Ohms.

Decibels. The Power Gain, or the ratio of output to input power of an amplifier, is sometimes expressed in Decibels or Transmission Units.

1 decibel or 1 T.U. $= 10 \log_{10} \frac{B}{A}$

where A and B are input and output powers respectively. Thus, if $\frac{B}{A} = 100$, then $10 \log_{10} 100 = 10 \times 2 = 20$ decibels, and if $\frac{B}{A} = 1,000$, then $10 \log_{10} 1,000 = 10 \times 3 = 30$ decibels.

D.C.	The symbol for Direct Current, that is, a Current that flows in one direction only.
A.C.	The symbol for Alternating Current. An alternating current is one which periodically rises to a maximum in one direction, falls to zero, and rises to a maximum in the opposite direction, and so on. The rise and fall of current is called a Period, or Cycle. The number of Periods, or Cycles, per second is called the frequency, or periodicity ; 50 periods is a common frequency, and is usually written 50 $\sim$.
Watt.	The Unit of Power in the product of Volts and Amperes. 746 Watts equal 1 horse-power.
Kilowatt.	1,000 Watts.
Kilowatt hours.	Kilowatts $\times$ hours. 1 kilowatthour (KWH) $=$ 1 Unit, the basis on which Electricity is sold.
Ampere-hours.	The product of Amperes and Hours, i.e., 1 ampere flowing for 1 hour is 1 ampere-hour, 2 amperes for 2 hours is 4 ampere-hours, and so on.
Insulator.	A material that does not allow an appreciable electric current to flow through it, e.g., Ebonite, Porcelain, Bakelite, Cotton, Silk, etc.
Capacity.	When two wires, coils, or sheets of metal, such as tinfoil, are placed close to one another but not touching, they have the property of storing a small amount of Electricity, and such an arrangement is said to have capacity. When a number of metallic sheets are assembled together with insulation separating them, they form a Condenser. The unit of capacity is the Farad, but as this is too big for ordinary use, we usually speak of Micro-Farads, i.e., 1,000,000ths of a Farad.
Inductance.	When a current flows through a coil it produces a certain amount of magnetism depending on the strength of the current, the size and number of turns in the coil. If the current increases or decreases the amount of magnetism varies accordingly, and the magnetic lines are said to cut

the turns of the coil. This cutting produces a voltage in the coil, the voltage always being in a direction tending to oppose a change of current; thus, if the current is increased the induced voltage opposes it, whilst if the current is reduced the voltage tends to keep it at its original value. This property is known as the inductance. Inductance is greatly increased by placing iron inside a coil. The unit of Inductance is the Henry.

Impedance. This is measured in Ohms, but Impedance differs from Resistance by being governed by other factors besides the dimensions and materials of which an article is made. The Resistance of a Valve between its Plate and Filament is termed its impedance; this Resistance is not constant, but varies with different H.T., G.B., and L.T. voltages. The impedance of a Coil or Transformer is the combined opposition of the winding, due to its Resistance, Inductance, and capacity, to the flow of an alternating current through it.

When R = Resistance.
L = Inductance.
K = Capacity.
$\sim$ = Frequency.
π = 3·14.

Then Impedance $= \sqrt{R^2 + \left(2\pi \sim L - \dfrac{1}{2\pi \sim K}\right)^2}$

A good Transformer must have High Primary Impedance. As the Impedance of a Transformer varies with the Frequency, the frequency should always be specified.

Amplification Factor. *Amplification Factor,* usually written μ or M, is the relation between the signal on the plate of a valve and that on its grid. If a signal of one volt on the grid results in a signal of 10 volts on the Plate the Valve is said to have an amplification factor of 10.

Mutual Conductance. The ratio $\dfrac{\text{Amplification Factor}}{\text{Anode Impedance}}$ is known as the Mutual Conductance of a valve, and is expressed in milliamps per volt.

Amplification
Ratio.

The *Amplification Ratio* of a Transformer or other device is the relation between the signal impressed on the grid of the preceding valve and the signal impressed on the grid of the following valve under working conditions. It is best shown by a curve drawn on the musical scale, as the amplification ratio varies somewhat with the frequency. In good coupling components such as Ferranti Transformers, the variation is slight and the amplification throughout the audible range is nearly equal to the amplification factor of the preceding valve multiplied by the Transformer ratio, $3\frac{1}{2}$.

The formula giving the approximate amplification of a Valve and Transformer is :—

$$Amplification = M \times Ratio \frac{Z}{\sqrt{R^2 + Z^2}}$$

Where M = Valve Amplification Factor.
Ratio = $3\frac{1}{2}$ in Ferranti Transformers.
R = Valve Impedance.
Z = Transformer primary impedance at any specified frequency.

Oscillograph.

A delicate Instrument capable of responding to variation of voltage or current occurring with great rapidity.

Oscillogram.

A photograph taken through the medium of an Oscillograph.

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