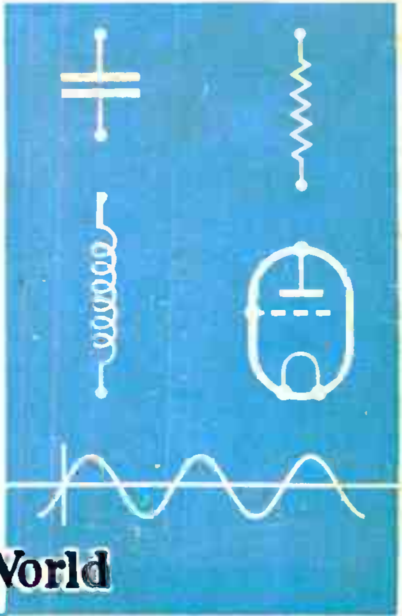
Foundations of Wireless

M. G. SCROGGIE, B.Sc., A.M.I.E.E.

FOURTH EDITION

COMPLETELY REVISED



Wireless World

World Radio History

FOUNDATIONS OF WIRELESS

By

M. G. Scroggie, B.Sc., A.M.I.E.E.

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Revised

"WIRELESS WORLD"

ILIFFE & SONS LTD

DORSET HOUSE, STAMFORD STREET, LONDON, S.E.1

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MADE AND PRINTED IN ENGLAND
AT
THE CHAPEL RIVER PRESS, ANDOVER, HANTS
Kp. 2/45

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PREFACE

ON MATHEMATICAL FORMULÆ

THIS book is non-mathematical in treatment, but nevertheless algebraic formulæ are quite freely used. The justification for this is simply that in no other way can information be recorded so briefly, clearly, and simply.

An algebraic formula is an abbreviated instruction to perform an arithmetical process. How, for example, do we measure the speed of a car ? If it goes 15 miles in half an hour, or 10 miles in 20 minutes, we spot at once that the speed is 30 miles per hour. How ? By dividing distance gone in a given time by the time taken ($15 \div \frac{1}{2} = 30$; $10 \div \frac{1}{3} = 30$). When we recognize this, we have a means of showing anyone who does not know how to find the speed exactly what to do, irrespective of the actual values of the times and distances involved ; we tell him to " Divide distance gone by time taken ".

Probably we forget to tell him that if the answer is to be in miles per hour, distance must be measured in miles and time in hours—it seems too obvious. Yet, in seeing that 10 miles in 20 minutes equals 30 m.p.h., we have automatically regarded 20 minutes as one-third of an hour. (Dividing miles gone by time taken *in minutes* gives speed in miles per minute—one-half in this case.) Similar attention to the units of measurement, not necessarily automatic in all problems, is always needed.

If in our instructions we replace the word " divide " by its mathematical symbol, our words take the form :

$$\text{Speed equals } \frac{\text{Distance Gone}}{\text{Time Taken}}$$

Here we have a brief convenient statement of an extremely general type ; note that it applies not only to cars, but to railway trains, snails, bullets, the stars in their courses, and anything else in heaven or earth that moves.

One small step further, and we are up to our eyes in algebra ; let us write $S = D/T$, where S stands for speed, D for distance, and T for time taken.

Letters stand Proxy for Numbers

Observe that to say that S equals D/T is utterly meaningless unless we say what the letters are meant to stand for. Most people who fail to grasp the essential simplicity of algebraic expression do so because they think that the letters used have some meaning *in themselves*, and do not realize that they only stand for numbers as yet unspecified.

Those faced for the first time with an algebraic expression such as this often say, " But how *can* one divide D by T ? Dividing one letter by another doesn't *mean* anything. If only they were numbers, now . . . ". Well, of course, that is just what they are—ordinary numbers, only we don't yet know their values. But we do know that when these values are found, dividing one by the other will give us the answer we want. So, in place of leaving blanks for the figures ("Blank divided by blank" would be ambiguous, to say the least of it) we put in letters, carefully defined in meaning, to act as temporary substitutes. No question of "dividing one letter by another" ever arises; one waits for the numbers.

Instead of looking on "S = D/T" as an instruction for calculating the speed, we can regard it as a statement showing the relationship to one another of the three quantities, speed, time and distance. Such a statement, always involving the "equals" sign, is called by mathematicians an "equation". From this point of view S is no more important that T or D, and it becomes a mere accident that the equation is written in such a form as to give instructions for finding S rather than for finding either of the other two. Since they are equal, we can divide or multiply the two "sides" of the equation by any number we please without upsetting their equality; if we multiply both by T (using the ordinary rules of arithmetic, since S, T, and D are really unspecified numbers) we get "S × T = D". Our equation now has the form of an instruction to multiply the time of the journey by the speed in order to find the distance gone. (Two hours at 20 m.p.h. takes us 40 miles.)

If we like to divide the new form of the equation by S, we get "T = D/S"—an instruction, now, to find the time consumed on a journey by dividing distance gone by speed. (Thirty miles at 20 m.p.h. would take $1\frac{1}{2}$ hours.)

It is important to note that these conversions of our original equation into new forms are independent of the meanings of the letters; the process is purely arithmetic, and consequently cannot give more information than the original equation contained. But such transformations are frequently made in order to twist the information provided into a form that will be more convenient when we come to put in the numbers for which the letters stand.

Other Symbols in Algebra

Wells's delightful episode of a tramp trying to read algebra —"Hex, little two up in the air, cross, and a fiddlededee"— reminds us that there are algebraic symbols other than letters. These, again, are only instructions to perform certain arithmetical operations on the numbers for which the letters stand.

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For example :

ab	means	" Multiply a by b ". Usually the tramp's " cross " is left out, and mere juxtaposition signifies multiplication. But the cross is restored and we write " $a \times b$ " where its absence might produce ambiguity.
x^2 (" Hex, little two up in the air "—read by the initiated as " x squared ")	means	Multiply x by x. x^n means n x's multiplied together.
$\sqrt{x}$	means	Take the square root of x, or find the number which, when multiplied by itself, makes x. Algebraically, we can say $(\sqrt{x})^2 = x$.
V_1, V_2, V_3		These " subscript " figures have no algebraic meaning : V_1, V_2, and V_3 are *single symbols*, each standing for a different V. They may be voltages at different parts of a circuit, numbered thus to distinguish them.

Unless the square-root sign ($\sqrt{\ }$) can be so called, we have not yet found a fiddlededee. Perhaps Greek letters belong to this mysterious class—several are in frequent use. In particular " π " (read as " pi ") is always used for the ratio of the circumference of a circle to its diameter ; it is mentioned here because its meaning is almost always taken for granted. The corresponding numerical value is 3·1416 approximately (the decimal never ends), or about 22/7. Other " fiddlededees " will be defined, like English letters, when we come to them.

Practical Use of Symbols

Having defined our symbols, let us see how they work. A commonly-used wireless formula is " $\lambda = 1{\cdot}885\sqrt{LC}$ ", λ (lambda) being wavelength in metres, L inductance in microhenries, and C capacitance in picofarads. The formula tells us that if we multiply (the numerical value in any particular case of) L by (the numerical value in that particular case of) C, take the square root of the result, and multiply that by 1·885, we shall be rewarded by (the numerical value in that particular case of) the wavelength. Usually we say, more briefly, " multiply L by C ", omitting the long-winded phrases in brackets. This, though really meaningless, is justified by the fact that we *can't* multiply L by C until we know what numerical values to take. Meanwhile we just write " LC " as an instruction to multiply as soon as that information is known. Note that the square-root sign is extended over both L and C ;

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this means that the extraction of the square root is to be applied to the product, and not to L only.

Brackets have a similar effect in lumping together the letters or figures within them. To find, in any numerical case the value of $a + b\left(\frac{m}{n} + p^2\right)^2$, we proceed thus: Square p, and divide m by n. Add the results, and square the sum so obtained. Multiply this by b, and then add a. Note that b, being outside the bracket, is not squared, but that it stands as multiplier to the whole term $\left(\frac{m}{n} + p^2\right)^2$. Note also that $-x \times -x$ (or $-(-x)$) equals $+x$; and that $1/x$ is called the *reciprocal* of x.

Examples of numerical substitution will be found in the body of the book, so none are given here.

Symbols for Verbal Convenience

It only remains to point out that when there is used a phrase like "the resistance R" it is not to be assumed that by virtue of some superior knowledge the writer is assured that this resistance *is* R, and that the reader has to accept that fact as one more of the unsolved mysteries of wireless. It only means that it is proposed to save space by using the symbol "R" to stand for "the numerical value of the resistance, whatever it may eventually turn out to be", or perhaps for "the numerical value of the resistance, whatever may be the value we choose to make it". Sometimes, indeed, the letter is just a handy label, meaning "the particular resistance marked R on the diagram". Often it will combine these meanings, and E/R may stand for "E divided by the numerical value of the resistance marked R in Fig. So-and-so".

Space is often saved also by using the "index notation" for very large or very small numbers. 10^6 means six tens multiplied together (see definition of x^2), which comes to one million. As an extension of this, 10^{-6} means "one divided by 10^6", or, in mathematical terms, "the reciprocal of" 10^6. It is, of course, one-millionth part. "$3{\cdot}2 \times 10^{-12}$" thus means "3·2 divided by one million million". The justification for this notation is that "0·0000000000032" is extremely difficult to read.

Note that $10^6 \times 10^6 = 10^{12}$, that $10^{12} \times 10^{-6} = 10^6$, and that $\frac{10^{12}}{10^6}$ is only another way of writing $10^{12} \times 10^{-6}$. In short, multiply by adding indices, and divide by subtracting them.

September, 1943.

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FOUNDATIONS OF WIRELESS

CHAPTER I

AN OUTLINE OF BROADCASTING

WIRELESS—or radio—is used for many things : broadcasting ; communication with and between ships, aircraft, tanks, trains, and cars ; direction finding ; radar (radiolocation) ; television ; photograph and "facsimile" transmission ; telephone and telegraph links ; meteorological probing of the upper atmosphere ; to mention some. All these are based on the same foundations, and it is necessary to understand most if not all of the subject matter of this book, no matter which application of wireless one intends to study in detail.

Most readers find purely abstract principles very heavy going ; it is more stimulating to have in mind some application of those principles. As it would be confusing to have all the applications of radio in mind at once, broadcasting has been selected most often, especially in dealing with reception, because it directly affects the largest number of people. If, therefore, references are made in some places to "music", the reader whose activities fall in a different sphere has only to substitute a different word.

1. What Wireless is Not

In discussing the day's wireless programmes one might easily remark to a friend : "There's some good music on the air to-night". This suggests a point of view that must be utterly abandoned before even beginning to grasp the mechanism of wireless transmission.

"Music on the air" suggests that the transmitting station sends out music as a disturbance of the air, which is music as we understand it in every-day life. But a transmitter is not a super-megaphone bawling out music ; its aerial emits no more sound than does an ordinary telephone wire. "Music" must therefore be sent out from a wireless station in some altered state, from which it can be converted back into ordinary audible music by the listener's receiving equipment.

Anyone who has watched a cricket match will recall that the smack of bat against ball is heard a moment after bat and ball are seen to meet ; the sound of the impact has taken an appreciable time to travel from the pitch to the grandstand.

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If the pitch were 1,100 feet away from the observer the time delay would be one second. Yet it is found that a watch may be set with apparently perfect accuracy by a wireless time signal from New York, providing, of course, that we allow for the fact that Americans do not use Greenwich Mean Time. That time signal has hurtled across the Atlantic in about a fiftieth part of a second. Comparing this with the three hours that would be required by any air-borne impulse we are driven to the conclusion that wireless transmissions travel in some other medium.

In the light of these facts "music on the air" has resolved itself into a silent substitute for music, carried by something that is not air.

2. Nature of Wireless Signals

The clue to the real nature of wireless signals is given by their rate of travel, which is the same as that of light. Light is one of the many possible disturbances in a mysterious and rather debatable medium called the "ether of space"; besides light there exist both longer and shorter ether waves which do not affect the eye at all.

The shortest waves, a few millionths of an inch long, affect only the smallest things, and are used by physicists to evoke disturbances within the atoms of which matter is composed, or to peer into atomic structure. The longer waves, which may be many yards long, also act on objects of physical dimensions comparable with their own. In particular, they affect metallic objects, such as wireless aerials, losing energy to them and setting up in them electric currents. All these waves, since they are all carried by the ether, travel at the same rate, which is about 186,000 miles per second.

3. Transmission and Reception

Natural processes are mostly reversible, so that the fact that ether waves of long wavelength set up electric currents in an aerial wire at once suggests that if by any means electric currents of a suitable kind can be made to flow in an aerial, that aerial will very probably radiate waves into the ether. In actual fact it does so, and recognition of this at once makes it evident that communication can be carried out between two points, even though separated by many miles, provided that we have some means of generating the currents at the transmitting end and recognizing them at the receiver.

The whole process is no more and no less wonderful than ordinary speech, during which air waves are set up by the motions of the speaker's vocal cords, transmitted over a distance

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of a yard or two by the intervening air, and reconverted into mechanical movements when they strike the listener's ear drum. The sequence "electric currents—electric waves—electric currents" is exactly analogous to the sequence "mechanical motions—air waves—mechanical motions" Communication by air waves, for which we use our own natural organs, seems merely commonplace; communication by electric waves is still something of a novelty, because it is only in this century that man has learnt to build himself transmitting and receiving stations, which are the electrical equivalents of mouth and ears.

The long distances over which wireless communication is possible is a result of the natural properties of the longer ether-waves; in communication by signal fires and heliograph the shorter (visual) waves have been used for generations for the sake of their ability to span greater distances than can conveniently be reached by waves in the air.

4. Waves

Of the various types of wave that we meet in daily life those formed when still water is disturbed are the nearest in character to the invisible air or ether waves. If we drop a stone into a pool and watch the resulting ripples carefully we shall observe that as they pass a twig or other small object floating on the surface they cause it to bob up and down. But the twig is not carried along bodily by the ripples.

The waves, therefore, do not consist of water flowing outwards from the point where the stone hit the surface, although they certainly give the impression that this is happening. As the twig shows, all that the water at any one point does is to move up and down rhythmically a few times before the wave dies away. The point is that nothing moves outwards from the centre but *energy* passed on from one part of the water to the next.

The behaviour of an air-wave is very similar. Suppose someone seated in the middle of a large room claps his hands. A listener seated against the wall will hear that hand-clap almost immediately. It is not to be imagined that the air suddenly compressed in the act of clapping has shot across the room to the listener's ear in that brief time. What has happened is that the body of air suddenly compressed by the clap has rebounded, compressing in the process the air immediately surrounding it. This, rebounding in its turn, has passed on the wave of compression in the same way until it has eventually reached the listener. All that has actually travelled across the room is *energy* in the form of compression of the air.

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5. Frequency and Wavelength

In wireless work one is more largely concerned with rhythmic waves than with irregular disturbances like that caused by a hand-clap. A stretched string, which emits a definite musical note, gives rise to a more important type of air wave.

When such a string is plucked or bowed it vibrates in the manner indicated in Fig. 1*a*. The movement of the string is rhythmic in the sense that each complete *cycle* of movements, from the highest position of A to the lowest *and back again*, occupies the same period of time. Moreover, each of these cycles is exactly like the last in every respect save that as the vibration dies away the amplitude of movement of the string becomes progressively less.

The length of the time occupied by each cycle determines the pitch of the note heard; if it is short, so that many vibrations take place each second, the note is high; while if it is long, so that only a few cycles of the movement occur in a second, the note is low. In scientific work of all kinds it is customary to specify a note in terms of the number of complete vibrations that occur in each second, this being known, for the sake of brevity, as the *frequency*.

Suppose the string vibrates at the rate of 20 cycles per second; in each second it will send out 20 compressions and 20 rarefactions of the air. The rate at which the wave that these compose will travel forward depends only on the medium through which it is passing; in air the velocity is about 1,100 feet per second. If we imagine that the string has been in vibration for exactly one second the wave corresponding to the first vibration will have reached a distance of 1,100 feet from the string

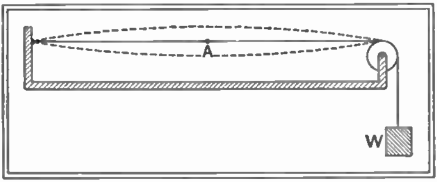

Fig. 1*a* (left): A stretched string vibrates in a regular manner when plucked or bowed, giving rise to a musical note of definite pitch. The size of the weight W controls the tension of the string, and therefore the pitch of the note

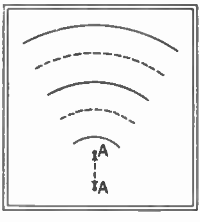

Fig. 1*b* (right): End view of the vibrating string at A in Fig. 1*a*. As it moves up and down over the distance AA it sends out alternate waves of compression (full line) and rarefaction (dotted line), which carry some of the energy of vibration to the listener's ear

just as the last wave (the 20th) is leaving it. There are, therefore, in existence 20 complete waves extending over a distance of 1,100 feet, from which it is evident (see Fig. 2) that each wave must be 55 feet long. If the string had executed 1,100 vibrations in the same period, the first would still have travelled 1,100 feet in the second of time occupied, and there would have been 1,100 complete waves in the series—each, therefore, one foot long. Since the velocity of sound in air is constant the higher frequencies correspond to the shorter wavelengths, and vice versa. It is specially to be noticed that it is the *frequency* of the vibration that is fundamental, and that the wavelength is a purely secondary matter depending on the

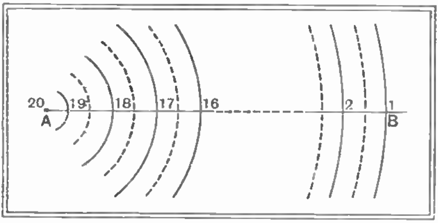

Fig. 2 : Twenty successive waves from the string. If the string is vibrating 20 times per second, the 1st wave has been travelling for one second by the time it reaches B, and the 20th is just leaving the string at A. Since sound travels 1,100 feet in one second, AB = 1,100 feet, and the distance between one wave and the next (wavelength) is 1/20th of 1,100 feet

velocity with which the wave travels. That it really is frequency, and not wavelength, that settles the musical note heard can be shown by sending a sound through water, in which the velocity is 4,700 feet per second ; the wavelength corresponding to a 1,100-cycle note is much greater than in air, but the pitch, as judged by the ear, remains the same as for a 1,100-cycle note in air.

The range of musical sound with which a wireless engineer has to deal runs from a low note of frequency about 25 cycles per second to a high note of frequency some 8,000 cycles per second, since if this range is fully reproduced music is sufficiently natural to give real pleasure to even a critical listener. The musical frequency-scale of Fig. 3 indicates, for reference, the frequencies corresponding to various notes.

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6. Wireless Waves

When we turn to the production of the wireless waves, by whose aid music is transmitted from place to place, we find frequencies of a very different order. These waves, as has already been mentioned, are set up by the surging to and fro of electric current in the aerial of the transmitter. Since the flow of electric current does not involve the movement of material objects, as does the vibration of the strings and reeds used in music, there is no great barrier to the production of very high frequencies indeed. If the current in the aerial vibrates at such a rate as to complete the double motion a million times in a second, it is *oscillating* at quite an ordinary *radio-frequency.* In such a case the surging current sends out into the ether a wave which has the electrical equivalent of compressions and rarefactions, the " compressions " following each other every millionth of a second.

Being a wave in the ether, our wireless wave travels at the invariable speed of all ether waves, 186,240 miles, or 300,000,000 metres, in each second. If, during one second, one million complete waves are radiated by the aerial, then at the end of that

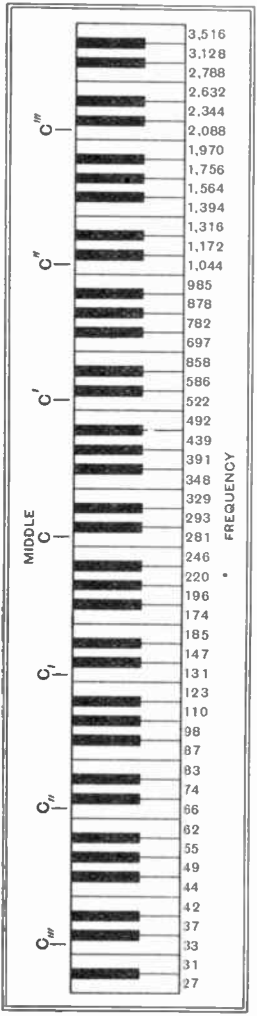

Fig. 3 : Showing the frequency corresponding to each musical note. Harmonics (multiples of the fundamental frequency shown) give notes their distinctive character ; hence the need to reproduce frequencies outside the range of music as written

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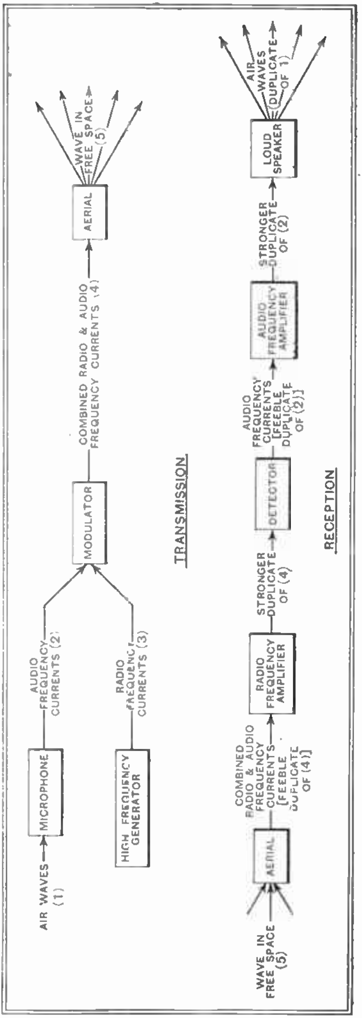

Fig. 4 : Schematic outline of broadcasting, showing how air-waves in the transmitting studio are duplicated, after many transformations, in the listener's home. Many stages of amplification have been omitted, for the sake of simplicity, from the diagram of transmission

time the first wave has travelled 300 million metres and the millionth is just leaving the aerial. Each wave, therefore, is 300 metres long. Just as in the case of sound, a lower frequency of electrical oscillation in the aerial will give rise to fewer waves each second, though the distance over which the waves emitted in one second will stretch remains the same. The waves, therefore, are longer. In symbols, the relationship is $\lambda = \frac{300{,}000{,}000}{f}$, where λ = wavelength in metres and f = frequency in cycles per second.

In dealing with sound, frequency is always used to specify the pitch of the note ; in wireless matters both frequency and wavelength are in common use. Since in this book we shall be much less concerned with the waves themselves than with the rapidly oscillating electric currents from which they are born and to which they give rise, we shall exhibit a definite bias towards the use of frequency rather than wavelength, on the grounds that the specification of wavelength is really meaningless except when considering a wave in free space.

7. From Studio to Listener

With a knowledge of the nature and relative frequencies of sound and wireless waves we can trace through, in the broadest outline, the whole process of broadcast transmission and reception. It is summed up, with almost ludicrous absence of detail, in the crude scheme of Fig. 4.

We begin in the studio, where we will imagine that an orchestra is playing a symphony. The result, brutally ignoring the æsthetic side, is a complicated mixture of air waves. These impinge on the diaphragm of a *microphone*, and this diaphragm, being thin, light and flexible, takes on the movements of the air in which it stands. The task of the microphone is to convert these movements of its diaphragm into movements of electricity, just as though the wire leading from it were a pipe filled with water pushed to and fro by the diaphragm.

The complicated air-waves are thus eventually translated into corresponding movements of electricity, so making a complex mixture of currents at frequencies which may lie anywhere within the range 25 to 8,000 or more cycles per second. They cannot be radiated from the aerial in their present form, partly because they are too weak and partly because the frequencies they represent are far too low to radiate well.

From another source a single regularly oscillating current, of a frequency suitable for wireless purposes (150,000 to 100,000,000 or more cycles per second) is produced, and the currents from the microphone are superposed on this in such a way that they mould (or *modulate*) it into their shape. The result is finally fed to the aerial, so that the wave sent out bears upon it, in the form of variations of strength, the impress of the currents derived from the microphone. These are then carried, in their new form, to any point on the globe to which the wireless wave itself can reach.

It has already been pointed out that a transmitter sends out a silent substitute for music; this complex wave is that substitute. The nature of this wave is discussed in detail in Chapter 9.

On striking an aerial this wave is partially absorbed by it, the energy so abstracted from the wave serving to set up in it a current which is an exact replica in miniature of the far more powerful current surging back and forth in the aerial of the

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transmitter. If the received signals* are very feeble, as they may be if the transmitter is distant or the aerial small, the first need is to strengthen them without changing their character. This is done by a *radio frequency amplifier* (Chapter 12). When sufficiently amplified the signals are passed to a *detector* (Chapter 10), which sorts out from the complex current representing the wave as a whole those parts of it which are directly due to the original music, rejecting those more rapidly oscillating currents which, in enabling the music to be transported from transmitter to receiver on the wings of a wireless wave, have now done all that is required of them.

The currents we now have left are as exact a copy of those given by the microphone in the studio as can be had after so many transformations ; they only require to be magnified up by another valve or two until they are strong enough to operate a loudspeaker (Chapter 14). To this they are accordingly passed, where they serve to push and pull a diaphragm (usually of paper) in such a way that its movements are a mechanical replica of the movements of the electric currents supplied to it. The diaphragm of the speaker thus performs the same movements as did that of the microphone a fraction of a second earlier ; in doing so it sets up in the listener's home air waves which are, as nearly as may be, identical with those produced by the orchestra.

* In the absence of a better word, the meaning of " signals " has been extended to include the electrical equivalents of speech and music.

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CHAPTER 2

ELEMENTARY ELECTRICAL NOTIONS

8. Electrons and the Electric Charge

THE exact nature of electricity is a mystery that may never be fully cleared up, but from what is known it is possible to form a sort of working model or picture which helps us to understand how it produces the results it does, and even to think out how to produce new results. The reason why the very existence of electricity went unnoticed until comparatively recent history, although it lay all around, is that it consists of two opposite kinds which in any piece of material exist in equal quantities and cancel one another out. For want of a better description, these kinds are called positive (+) and negative (—). In arithmetic + 1 — 1 equals just nothing; and so although both positive and negative electricity produce very remarkable effects when they are separate, a combination of equal quantities shows no apparent electrification.

One of the most startling conclusions is that all matter is composed from electricity. The atoms of all substances are built up from three kinds of particles: particles of positive electricity, called *protons*; particles of negative electricity, *electrons*; and particles which, being electrically neutral, are called *neutrons*. The differences between the chemical elements —oxygen, carbon, aluminium, and so on—result solely from the differences in the number and arrangement of those particles in their atoms.

Each atom can be imagined as a sort of ultra-microscopic solar system in which a number of electrons revolve round a *nucleus* (consisting only of protons and neutrons), rather as our earth and the other planets revolve round the sun.

The gas, hydrogen, is the lightest element and has the simplest atom; it consists of one proton (+ 1) with one electron (— 1) revolving round it. The next lightest substance, helium, has a nucleus consisting of two protons (+ 2) and two neutrons, with two electrons (— 2) revolving round it (Fig. 5). And so on for more complicated atoms; in every case there is the same number of electrons revolving round the nucleus as there are protons inside the nucleus, so that all atoms in their normal state contain equal numbers of + and — charges and so show no electrification.

Certain agencies—heat, light, friction, etc.—can, however, detach one or more of the electrons from atoms, leaving them deficient in negative electricity, and as a result showing all the

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characteristics of positive electrification, or, as it is usually called, a positive charge. The detached electron, or any piece of matter harbouring it, displays every evidence of a negative charge, including an attraction for positive charges.

The space between them, across which this attraction is exerted, is said to be subject to an *electric field*. The greater the number of opposite charges, and the closer they are, the more intense the field and the greater the attractive force. The force tends to drive electrons, which are much more mobile than their positive counterparts, from places where they are in the majority to places where they are relatively fewer, until there are equal proportions on both sides, when the attractive

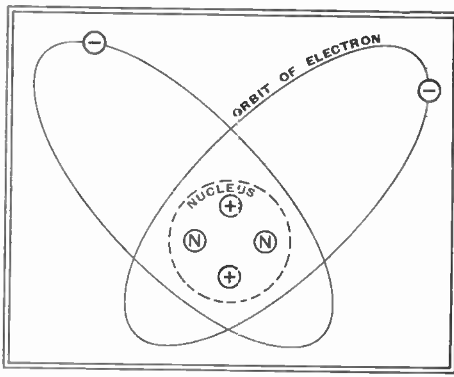

Fig. 5 : Diagram of imagined structure of helium atom
+ Proton N Neutron
— Electron

force ceases. The action is rather like that of water, which always tends to flow between any two points at different levels until the levels are equivalent.

It needs a certain amount of energy to force electrons to go to a place where they are already in a majority, against the repulsion between like kinds ; and this energy is stored ready to be released if the attraction for electricity of the opposite kind is allowed to assert itself; reuniting the electrons with their positive mates. Water power, too, can be stored, by pumping it up to a higher level against the force of gravity, and can then be used to drive a water-wheel or turbine in its descent.

9. Conductors and Insulators

For this flow to take place a pipe or other channel must be provided, while for an electron flow an electrically conducting path must be provided between the two points. A conducting material is one in which electrons are very readily detached from their parent atoms, so that if a wire is stretched between two oppositely charged bodies, electrons can enter the wire at one end and cause a displacement of free electrons all down the wire, resulting in the emergence of an equal number of electrons at the other. Picture a long pipe, already filled with water. If an extra teaspoonful of water is forced into it at one end a teaspoonful will emerge at the other—but not the same actual water. If milk had been forced in instead of water, water would still have emerged. In the same way, the wire in its normal state must be pictured as already filled with electrons, all in continuous random movement from atom to atom. The passage of electricity through the wire amounts to no more than the superposition upon this vast random movement of a trifling drift in one direction ; the emerging electrons may only have moved a thousandth of an inch.

If the atoms of a substance have their electrons so firmly fixed that this exchange is not possible, the material will not conduct ; it is called an *insulator*. All metals are *conductors* ; to the class of insulators belong ebonite, bakelite, rubber, the silk or enamel covering on wire, and, indeed, most non-metallic substances.

The flow of electrons through a conductor constitutes a *current of electricity*.

10. Fundamental Electrical Units

So far we have considered the current as originating from a body which has a small and temporary excess of electrons ; when the charge is dissipated the current must inevitably stop. There are, however, certain appliances, such as batteries and dynamos, that have the power of continuously replenishing the surplus of electrons ; so that if the two terminals of a battery are joined by a conductor a current flows continuously, or at least until the chemicals in the battery, which are responsible for the action, are used up. The difference in electrical level or pressure tending to drive a current through any continuous path, or *circuit*, leading from one terminal to the other, is called the electromotive force or E.M.F. and is measured in *volts*.

The current that flows might very reasonably be measured in terms of the number of electrons passing from the battery into the circuit each second, but the electron is so extremely small that such a description of any useful current would lead to inconveniently large numbers. In consequence, it has become customary to take as the practical unit a body of about six million billion (6,000,000 000,000,000,000) electrons. This

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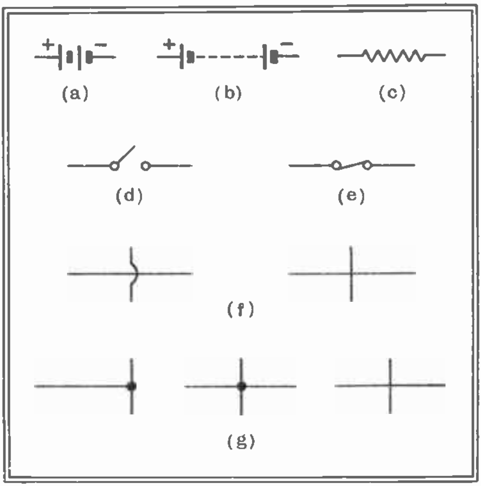
is called the *coulomb*, and is a unit of *quantity of electricity*, just as the gallon is a unit of quantity of water.

Just as one might speak of a flow of water of so many gallons per second, one can quite correctly describe an electric current as so many coulombs per second. Such a description, however, is rather cumbersome for frequent use, and the composite unit coulombs-per-second, as a measure of the rate of flow of electricity, is replaced by the more briefly named unit, the *ampere*. The statement that a current of one ampere is flowing means

Fig. 6: Some conventional signs used in constructing electrical diagrams. (*a*) A battery of few cells, used for filament accumulator or grid battery. Two cells are shown, making either a 4-volt accumulator or a 3-volt dry battery. (*b*) A battery of many cells, e.g., a high-tension battery. (*c*) A resistor. (*d*) A switch, shown open. (*e*) A fuse. (*f*) Wires crossing: the sign on the left is more usual. (*g*) Wires joining: the "dot" is generally used. Note that a simple line always indicates an electrical connection of negligible resistance



that one coulomb of electricity, or about 6×10^{18} electrons, flows past any point in the path of the current in each second.

Water, driven through a pipe by a constant pressure, will flow at a rate depending on the frictional resistance between the water and the inside of the pipe. Further, a pipe of large diameter will offer less resistance than one of small bore, and so will carry a larger flow at any given pressure. In just the same way, the magnitude of the current of electricity driven through a conductor by a battery depends on the electrical resistance offered by that conductor to its flow, and a thick wire offers less resistance than a thin one of equal length.

Circuits, especially the more complex ones, are more easily grasped from a diagram than from a description in words. Fig. 6 shows some of the conventional symbols from which

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electrical diagrams are constructed. A circuit element designed to provide resistance is known as a *resistor*. Each type of component—battery, resistance, switch, etc.—has its own sign, and the way they are joined up to make the complete circuit is indicated by lines representing the wiring. A wire is always supposed to provide an electrical connection of negligible resistance. Other symbols will be introduced into diagrams as they are needed.

11. Ohm's Law

Suppose an experiment is performed with the very simple circuit shown in Fig. 7, in which the E.M.F. driving current through a resistance can be varied by using varying numbers of cells, and the strength of current is measured by a meter, such as is de-

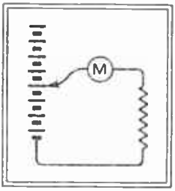
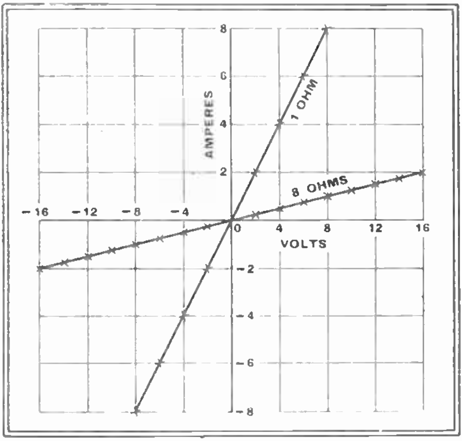


Fig. 7 : Apparatus for carrying out an experiment leading to Ohm's Law. A variable number of cells can be used to drive current through a resistance, and the strength of current is measured by a meter M.

scribed in Sec. 26. In Fig. 8, the results of such an experiment are shown in the form of a graph. If each cell gives 2 volts, then readings of current can be taken at 2, 4, 6, 8, 10, etc., volts, and at each of these voltages (indicated by distance to the right from the centre) the current is indicated by height

Fig. 8 : Results of experiment with circuit Fig. 7 shown as a graph, for two different resistances.



above the centre. The battery can then be reversed, causing the current to flow the opposite way round the circuit. These readings are distinguished from the first set by being called negative, and are plotted to the left and downwards.

When all the dots are joined up by a line (representing the readings that might be taken at every intermediate voltage) it will generally be found—unless the resistance is of a special class considered in Chapter 7—that the line is straight and inclined at an angle. Changing the resistance, a line will be obtained which slopes at a different angle. Fig. 8 shows two possible samples resulting from such an experiment. Each of these lines shows how much current will flow, in the resistance to which it refers, when a certain E.M.F. is supplied ; or the amount of E.M.F. required to cause a certain current to flow

Referring to the steep line, the experiment showed that 2 volts caused 2 amps., 4 volts 4 amps., and so on. The result of dividing the number of amps. by the number of volts is always 1. This being so, it is not necessary to have a graph to find out the current due to any E.M.F. ; we can express the same result as :

$$\frac{\text{Volts}}{\text{Amperes}} = 1 \text{ or Amperes} = \frac{\text{Volts}}{1}$$

The corresponding statement for the other line is :

$$\frac{\text{Volts}}{\text{Amperes}} = 8 \text{ or Amperes} = \frac{\text{Volts}}{8}$$

The quantity—1 or 8—determining the current that will flow is the numerical value of the resistance, and the unit is called the *ohm*, so we can say :

$$\frac{\text{Volts}}{\text{Amperes}} = \text{Ohms}, \quad \text{or Amperes} = \frac{\text{Volts}}{\text{Ohms}},$$
$$\text{or Volts} = \text{Amperes} \times \text{Ohms}.$$

This relationship, by which, if two of these quantities are known, the third can be found, is known as Ohm's Law, after the discoverer.

For brevity, electrical quantities such as these are denoted by letters :

Quantity	*Symbol for Quantity*	*Unit of Quantity*	*Symbol for Unit*
E.M.F.	E	Volt	V
Current	I	Ampere	A
Resistance	R	Ohm	Ω

So Ohm's Law is usually written as $I = E/R$ or $R = E/I$ or $E = IR$, it being understood that the above units are used in working out any particular case.

If, for example, we have a 2-volt accumulator connected to a length of wire having a resistance of 10 ohms (Fig. 9), the current flowing will be 2/10ths of an ampere. If the resistance had been only half this value, the current would have been twice as great, and it would have had this same doubled value if the original resistance had been retained and a second accumulator cell had been added to the first to make a total E.M.F. of 4 volts.

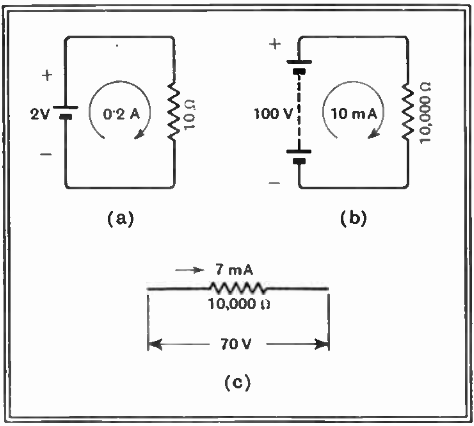


Fig. 9: Circuits, illustrating Ohm's Law, constructed from symbols of Fig. 6. *a* and *b* show the application of Ohm's Law to a complete circuit, including the E.M.F. of the battery. (*c*) The application of the Law to part of a circuit; if 7 mA flows through 10,000 ohms the voltage across the resistance must be 70 volts

Taking another case, we might find, in investigating the value of an unknown resistance, that when it was connected across the terminals of a 100-volt battery a current of 0·01 ampere was driven through it. Using Ohm's Law in the form $R = E/I$, we get for the value of the resistance $100/0{\cdot}01 = 10{,}000$ ohms. Alternatively, we might know the value of the resistance and find that an old battery, nominally of 120 volts, could only drive a current of 0·007 ampere through it. We could deduce, since $E = I \times R$, that the voltage of the battery had fallen to $10{,}000 \times 0{\cdot}007 = 70$ volts.

12. Practical Units

It is not usual to describe a current as 0·007 ampere as was done just now; one speaks of "7 milliamperes", or, more familiarly still, of "7 milliamps". A milliampere is thus seen

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to be a thousandth part of an ampere. Several other such convenient prefixes are in common use ; the most frequent are :

Prefix	*Meaning.*	*Symbol.*
milli-	One thousandth of	m
micro-	One millionth of	μ
micromicro- or pico-	One billionth of (million millionth)	$\mu\mu$ or p
kilo-	One thousand	k
mega-	One million	M

These prefixes can be put in front of any unit ; one speaks commonly of milliamps., microamps., kilocycles per second, megohms, and half a dozen other such units. " Half a megohm" comes much more trippingly off the tongue than " Five hundred thousand ohms ", just as $\frac{1}{2}M\Omega$ is quicker to write than $500{,}000\Omega$.

It must be repeated, however, that Ohm's Law refers to volts, ohms, and amperes. If a current of 5 milliamps is flowing through 15,000 ohms, the voltage across that resistance will *not* be 75,000 volts. The current must be expressed as 0·005 amp. before the correct result, 75 volts, is obtained for the magnitude of the potential difference.

13. Kirchhoff's Law

The term *potential difference* is used above instead of E.M.F. because the voltage across the resistance is a *result* of the current, and not the cause of it. The E.M.F. driving the current must be due to a battery or other source somewhere in the circuit. Going back to our water-flow analogy ; if a circulating pump forces water through a narrow pipe or a filter offering resistance to its flow, a difference in pressure is caused between the two ends of the pipe. This difference in pressure (or potential) is exactly equal to what may be called the water-motive-force, or pressure in pounds per square inch, produced by the pump. Water pressure can be measured by the height at which the liquid will rise in a vertical pipe open at the top. So it is sometimes specified as a " head " of so many feet, and a pressure difference can be regarded as a difference in level. The greater the difference in level (either feet or volts) the greater the tendency to cause a current (of water or electricity) to flow.

Whenever there is an E.M.F. in a circuit, the resulting current causes a potential difference (P.D.) that is exactly equal and opposite. The P.D. may be the result of the resistance offered by one or more sections of the circuit, or of other circuit effects that will be considered in the next chapter. So if one starts at any point in a closed circuit, reckoning the total of E.M.F.'s

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as (say) positive, and the opposing P.D.'s as negative, one must always find the results adds up to nil. If it doesn't, a mistake has been made somewhere ; just as a surveyor must have made an error if he starts off from a certain spot, reckons his increases in height above sea level as positive and his descents as negative, and after following a circular route to his starting point finds that (on paper) he has made a net ascent or descent. Obviously he can be neither higher nor lower when he arrives back at the same spot ; and in the same way the total voltage round any complete circuit, when account is taken of positive and negative, must be zero. This is a very valuable check on one's working, especially in complicated circuits. This principle, in fact, although it follows directly from Ohm's Law, is regarded as of sufficient importance to be called a Law, too, associated with the name of Kirchhoff.

The reader must be warned against a possible cause of confusion with regard to the direction of current flow. In the early days of electrical knowledge, before it was clear whether electric currents consisted of positive electricity flowing one way, or negative electricity flowing in the opposite direction, or both at once, it was agreed to guess that positive electricity flowed from one terminal of a battery (which was therefore marked +) through the external circuit to the negative or — terminal. Unfortunately it was a bad guess, for in most cases the current consists of electrons flowing from — to + terminals, or in the opposite direction to that supposed to be taken by the current. As it has become so firmly established to talk about current flowing from positive to negative terminal, this convention is adhered to in this book (note direction of arrows in Fig. 9) ; so it must be remembered that it is likely to be electrons flowing in the opposite direction.

14. Electrical Power

It would be a commonplace to point out that to pump water along a horizontal pipe some small amount of *power* would be required to overcome the friction. It is equally true to say that if electricity is driven through a conductor some power is required to overcome the resistance of that conductor. A rise either in voltage (pressure), current (flow of water), or resistance (friction) will naturally increase the power necessary to maintain the flow. Since these three are related by Ohm's Law, the power needed can be expressed in terms of any two of them. Using standard symbols the power is :— $P = I^2R$, or EI, or E^2/R.

Any of these expressions can be used for calculating the power

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expended in a circuit, according to whether current and resistance, voltage and current, or voltage and resistance are known. Once again the units to be used are ohms, amperes, and volts, while the unit of power is the *watt* (symbol W). One watt is the power expended when a current of one ampere is driven by an E.M.F. of one volt.

Take the case of an electric fire having a resistance of 20 ohms, connected to 200-volt mains. By Ohm's Law the current will be 10 amperes. The three expressions for power work out, for this case, as follows :—

$I^2R = 10^2 \times 20 = 100 \times 20 = 2{,}000$ watts.

$EI = 200 \times 10 = 2{,}000$ watts.

$E^2/R = 200^2/20 = 40{,}000/20 = 2{,}000$ watts.

When electrical energy is consumed, some other form of energy must necessarily appear in its place (Law of Conservation of Energy). In the case given it is fairly evident that the electricity consumed is converted into heat. This is equally true of any case where a current passes through a resistance, though if the dissipation of power is small, the rise in temperature may not be noticed. For example the heat developed by a 15,000 ohm resistor carrying 5mA, which only dissipates 375 milliwatts (0·375 watt) would be quite difficult to detect.

It is important to note that the watt is a unit of power, which is rate of doing work, and not of simple work or energy. A ten-horse-power engine exerts ten horse-power, no matter whether it runs for a second or a day ; if it continues for an hour the work done is ten horse-power-hours. Similarly, one coulomb per second under a pressure of one volt is one watt, no matter how long it flows. If the 2,000-watt fire were left on for eight hours the power would be 2 kilowatts at any moment during that time, and the total energy expended would be 16 *kilowatt-hours*. A kilowatt-hour is the " unit " charged for in the quarterly electric-light bill.

15. Resistances in Series or Parallel

It is only in the simplest cases that a circuit consists of no more than a source of E.M.F. and a single resistance. A battery lighting a single lamp or a single valve-filament is one of the few practical examples. The circuits with which we shall have to deal will in most cases contain several resistances or other circuit elements, and these may be connected either *in series* or *in parallel*. Fortunately Ohm's Law can be applied to every part of a complex circuit as well as to the whole. Beginners often seem reluctant to make use of this fact, being scared by the apparent difficulty of the problem.

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Two elements are said to be in series when in tracing out the path of the current we encounter them serially, one after the other. In Fig. 10 the two resistances R_1 and R_2 are connected in this way. Remembering that an electric current is an electron-flow, it will be evident that *the same current flows through both of them.*

Two elements are said to be in parallel if they are so connected in the circuit that they form two alternative paths for the current flowing between a pair of points. In Fig. 11, for example, R_1 and R_2 are alternative paths for conveying current from A to B. It will be evident, from the nature of things, that *the same potential difference exists across both of them.*

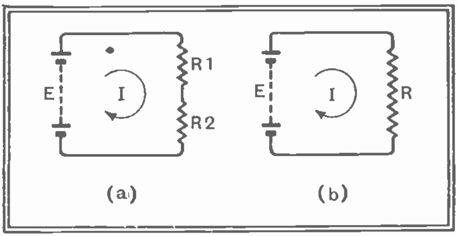

Fig. 10 : Resistances in Series. The circuit *b* is equivalent to the circuit *a*, in the sense that both take the same current from the battery E, if $R = R_1 + R_2$

16. Resistances in Series

In Fig. 10 *a* two resistances, R_1 and R_2, are shown connected in series with one another and with the battery of voltage E. To relate this circuit to the simpler ones already discussed we need to know what single resistance R (Fig. 10 *b*) can be used as a substitute for R_1 and R_2 taken together.

We know that the current in the circuit is everywhere the same; call it I. Then the potential difference across R_1 is IR_1, and that across R_2 is IR_2 (Ohm's Law). The total voltage-drop is the sum of these two, namely, $IR_1 + IR_2$, or $I(R_1 + R_2)$, and is equal to the voltage E of the battery. In the equivalent circuit of Fig. 10 *b*, E is equal to IR, and since, to make the circuits truly equivalent, the currents must be the same in both for the same battery-voltage, we see that $R = R_1 + R_2$. Generalizing from this result, we conclude that: *The total resistance of several resistances in series is equal to the sum of their individual resistances.*

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17. Resistances in Parallel

Turning to the parallel-connected resistances of Fig. 11 *a*, we have the fundamental fact that they have the same voltage across them; in this case the E.M.F. of the battery. Each of these resistances will take a current depending on its own resistance and on the E.M.F. of the battery; the simplest case of Ohm's Law. Calling the currents respectively I_1 and I_2, we therefore know that $I_1 = E/R_1$ and $I_2 = E/R_2$. The total current drawn is the sum of the two: it is $I = E/R_1 + E/R_2 = E\ (1/R_1 + 1/R_2)$. In the equivalent circuit of Fig. 11 the current is E/R, which may also be written $E\ (1/R)$. Since, for true equivalence between the circuits, the current must be the same for the same battery voltage, we see that $1/R = 1/R_1 + 1/R_2$. Generalizing from this result, we may conclude that: *If several resistances are connected in parallel the sum of the reciprocals of their individual resistances is equal to the reciprocal of their total resistance.*

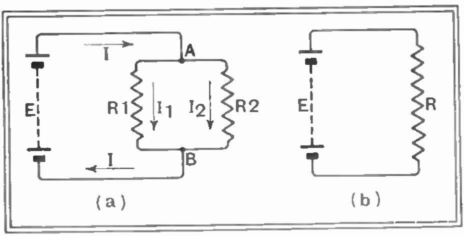

Fig. 11: Resistances in parallel. The circuit *b* is equivalent to the circuit *a* in the sense that both take the same current from the battery E, if $1/R = 1/R_1 + 1/R_2$

If the resistance of Fig. 11 *a* were 100 and 200 ohms, the single resistance R that, connected in their place, would draw the same current is given by $1/R = 1/100 + 1/200 = 0{\cdot}01 + 0{\cdot}005 = 0{\cdot}015$. Hence, $R = 1/0{\cdot}015 = 66{\cdot}67$ ohms. This could be checked by summing the individual currents through 100 and 200 ohms, and comparing the total with the current taken from the same voltage-source by 66·67 ohms. In both cases the result is 0·015 ampere per volt of battery.

Summing up, we have the two rules which, expressed in symbolic form, are :—

1. Series Connection. $R = R_1 + R_2 + R_3 + R_4 + \ldots\ldots\ldots$
2. Parallel Connection. $1/R = 1/R_1 + 1/R_2 + 1/R_3 + \ldots\ldots\ldots$

For two resistances in parallel, a more easily calculated form of the same rule is $R = R_1R_2/(R_1 + R_2)$. More than two resistances in parallel can be dealt with by first working out the equivalent of two of them, and then combining this with the next ; and so on.

18. Series-Parallel Combinations

These rules can be extended to cover quite a complicated network of resistances. In such cases the algebra required,

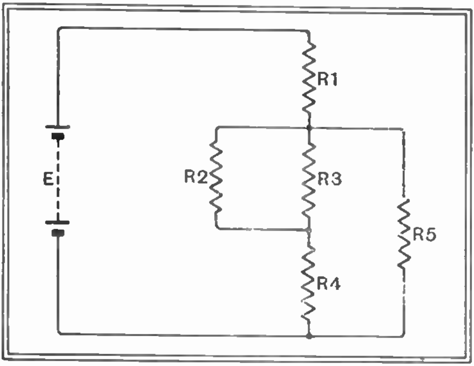

Fig. 12 : A complicated network of resistances. The current through and voltage across each can be computed with the aid of the rules already discussed. R1 = 100 ohms, R2 = 200 ohms, R3 = 500 ohms, R4 = 150 ohms, R5 = 1000 ohms, E = 40 volts

though perfectly simple, is inclined to get very long-winded if an attempt is made to work out a general formula ; we will therefore content ourselves with one example, worked out numerically. The example will be the circuit of Fig. 12 ; we will find the total current flowing, the equivalent resistance of the whole circuit, and the voltage and current of every resistor individually.

The bunch R_2, R_3, R_4, R_5 is obviously going to be our stumbling block, so we will begin by simplifying it. In doing this it is always necessary to work from the inside outwards. Writing R_{23} to symbolize the combined resistance of R_2 and R_3 taken together, we know that $R_{23} = R_2R_3/(R_2 + R_3) = 200 \times 500/700 = 142{\cdot}8$ ohms. This gives us the simplified circuit of Fig. 13 *a*. If R_{23} and R_4 were one resistance, they and R_5 in parallel would make another simple case, so we proceed to combine R_{23} and R_4 to make R_{234}.

$R_{234} = R_2 + R_4 = 142{\cdot}8 + 150 = 292{\cdot}8$ ohms. Now we have the circuit of Fig. 13*b*. Combining R_{234} and R_5, $R_{2345} = 292{\cdot}8 \times 1000/1292{\cdot}8 = 226{\cdot}5$ ohms. This brings us within sight of the end ; Fig. 13 *c* shows us that the total resistance of the network now is simply the sum of the two remaining resist-

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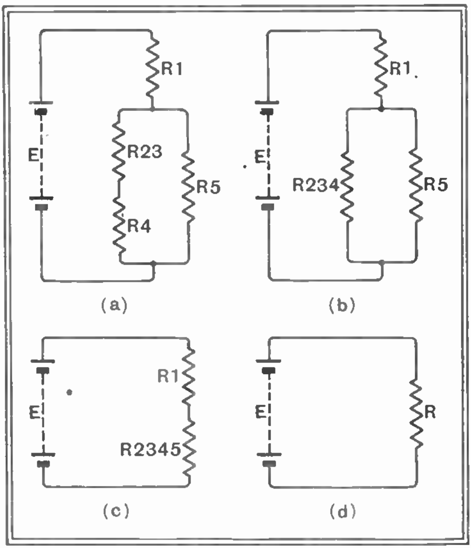
ances ; that is, R of Fig. 13 *d* is $R_{2345} + R_1 = 226{\cdot}5 + 100 = 326{\cdot}5$ ohms.

From the point of view of current drawn from the 40-volt source the whole system of Fig. 12 is equivalent to a single resistor of this value. The current taken from the battery will therefore be $40/326{\cdot}5 = 0{\cdot}1225$ amp. $= 122{\cdot}5$ mA.

To find the current through each resistor individually now merely means the application of Ohm's Law to some of our previous results. Since R_1, carries the whole current of $122{\cdot}5$ mA, the potential difference across it will be $100 \times 0{\cdot}1225 = 12{\cdot}25$ volts. R_{2345} also carries the whole current (13 *c*) ; the p.d. across it will again be the product of resistance and current, in this case $226{\cdot}5 \times 0{\cdot}1225 = 27{\cdot}75$ volts. This same voltage also exists, as comparison of the various diagrams will show, across the whole complex system $R_2R_3R_4R_5$ in Fig. 12. Across

Fig. 13: Successive stages in simplifying the circuit of Fig. 12. R_{23} stands for the single resistances equivalent to R_2 and R_3; R_{234} to that equivalent to R_2, R_3 and R_4, and so on. R represents the whole system



R_5 there lies the whole of this voltage ; the current through this resistor will therefore be $27{\cdot}75/1000$ amp. $= 27{\cdot}75$ mA.

The same p.d. across R_{234} of Fig. 13 *b*, or across the system $R_2\,R_3\,R_4$ of Fig. 12, will drive a current of $27{\cdot}75/292{\cdot}8 = 94{\cdot}75$ mA through this branch. The whole of this flows through R_4 (13 *a*), the voltage across which will accordingly be $150 \times$

0·09475 = 14·21 V. Similarly, the p.d. across R_{23} in Fig. 13 *a*, or across both R_2 and R_3 in Fig. 12, will be 0·09475 × 142·8 = 13·54 volts, from which we find that the currents through R_2 and R_3 will be respectively 13·54/200 and 13·54/500 amp., or 67·68 and 27·07 mA, making up the required total of 94·75 mA for this branch.

This gives a complete analysis of the entire circuit ; we can now collect our scattered results in the form of the following table :—

RESULTS OF SOLVING FIG. 12

Resistance	Current [milliamps.]	Voltage [volts]	Power [watts]
R_1	122·5	12·25	1·501
R_2	67·68	13·34	0·916
R_3	27·07	13·54	0·367
R_4	94·75	14·21	1·346
R_5	27·75	27·75	0·771

It is instructive to apply Kirchhoff's Law to the several closed circuits that are included in this network. Consider that formed by E, R_1, and R_5. If we take the route clockwise we go "uphill" through the battery, becoming more positive by 40 volts. Coming down through R_1 we move from positive to negative so add — 12·25 volts, and, through R_5, — 27·75 volts, reaching the starting point again. Check : 40 — 12·25 — 27·75 = 0.

Taking another clockwise route via R_3, R_5, and R_4, we get 13·54 — 27·75 + 14·21 = 0. And so on for any closed loop.

It should be noted that by using suitable resistors any potential intermediate between those given by the terminals of the battery can be obtained. For instance, if the lower and upper ends of the battery in Fig. 12 are regarded as 0 and + 40 (they could equally be reckoned as — 40 and 0, or — 10 and + 30, with respect to any selected level of voltage), the potential of the junction between R_3 and R_4 is 14·21 volts. The arrangement is therefore called a *potential divider*, and is often employed for obtaining a desired potential not given directly by the terminals of the source. If a sliding connection is provided on a resistor, to give a continuously variable potential, it is generally known—though not always quite justifiably—as a *potentiometer*.

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CHAPTER 3

CAPACITANCE AND INDUCTANCE

19. Capacitance : What it is

WE have seen that because positive and negative electricity attract one another they are normally together, cancelling one another out. It needs something special—an electromotive force—to separate them. Directly they are separated, this mysterious attraction comes into play. Unless the EMF is maintained, they will drift together again; very rapidly through a conductor, slowly through an insulator.

The most convenient method of specifying the strength of an electric attraction or *potential* is in terms of the E.M.F. required to overcome it ; that is to say, in volts. How is this related to the magnitude of the charges between which the potential exists ?

Suppose a battery has its positive terminal connected to earth, which is normally regarded as zero potential. The negative terminal is connected to any insulated piece of metal, say a plate suspended over the ground, as in Fig. 14. The E.M.F. of the battery, say 50 volts, drives electrons up from earth on to the plate. The loss of electrons from the earth has no appreciable effect on its potential, any more than pumping a few gallons of water up from the ocean would affect sea level. When the electrons get on to the plate their mutual repulsions cause them to spread all over it. The negative potential of the plate, as one would expect, increases the more it becomes occupied by electrons over and above the normal electronic population that is neutralized by positive counterparts. When the potential due to this surplus of electrons is

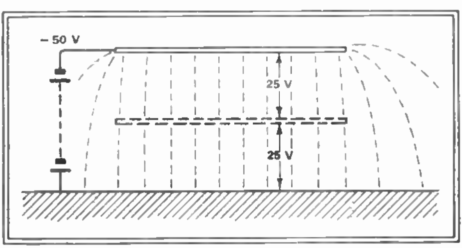

Fig. 14 A metal plate suspended above the ground and charged negatively. Half-way between, the voltage is half that of the plate

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sufficient to stop any more from coming we say that it is — 50 volts.*

The attractive force or field between electrons and the ground causes the plate to be attracted with a certain force. The directions along which the attraction is felt are indicated on diagrams by lines, as shown, termed *lines of electric force*. Except near the edges of the plate they are practically parallel. To represent a more intense field more lines are drawn in a given space.

If now the voltage of the battery is doubled, it will drive more electrons on to the plate. Actually twice as many electrons are needed to charge the plate to — 100 volts ; *the potential is directly proportional to the charge*. The quantity of electrons required to charge the plate to 1 volt is called the *capacitance* of the plate. As capacitance is not necessarily related to earth, it is more general to say that the capacitance of one conductor relative to another is equal to the quantity of electricity it is necessary to transfer from one to the other in order to establish a difference of potential between them of 1 volt. Our unit of electrical quantity (Sec. 10) is the coulomb or ampere-second. So the obvious unit of capacitance is that which requires 1 coulomb to charge it to 1 volt, and is named the *farad* in honour of Faraday, who contributed so much to electrical science. In symbols, $Q = CV$ where Q is the charge in coulombs required to charge a condenser of C farads to a potential difference of V volts. Just as the embodiment of resistance is a resistor, a device designed for its capacitance is called a *capacitor* or more commonly a *condenser*.

20. Factors Determining Capacitance

Referring again to Fig. 14, suppose a second plate, exactly the same as the first, is suspended at an equal distance above earth. To charge it also to — 50 volts it has to receive the same quantity of electrons as the first. The capacitance of two equal condensers connected in parallel is twice that of one. Stated more generally, the capacitance is proportional to the

* Incidentally, if we join the plate to earth by some conductor containing no E.M.F., there is nothing to stop electrons going down it. Directly the number on the plate decreases, its potential falls below that of the battery, which immediately drives more up from earth. And so on, a continuous circulation or flow of current begins. The rate of flow adjusts itself to the amount that results in a back pressure of 50 volts due to the resistance of the conductor. Then there is a perfect balance between the E.M.F. of the battery and IR (the current multiplied by the resistance of the circuit), both being equal to the potential of the plate. This is just another way of looking at Ohm's Law.

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area, other things being equal. (This statement is subject to a correction to allow for "edge effect," but is very nearly true when the area is large compared with the spacing.)

Next, suppose a second plate is placed midway between the first and ground, as shown dotted. Assuming all the plate and earth surfaces are flat and parallel, and the spacing relatively small, the potential is distributed at a uniform rate across the space; therefore, from middle plate to top the potential is 25 volts. To raise it to 50, twice as many electrons would have to be driven on to the top plate; in other words, halving the spacing has doubled the capacitance, or, more generally, the capacitance is inversely proportional to the distance between the parallel plates.

Suppose now that instead of air between the plates we use some other insulating material such as glass, rubber, or oil. It is found that the capacitance is thereby multiplied by some figure which is called the *permittivity* of the material. Strictly speaking, the comparison should be made with a vacuum, but the permittivity of air is so nearly equal to 1 that the difference is generally ignored. Insulating material occupying the space between condenser plates, and subject to the stress of the potential, is called a *dielectric*, and another name for permittivity is *dielectric constant*. The permittivity of most solid and liquid dielectrics is between 3 and 10, but with special ceramic materials it may be as much as 80.

The capacitance of a condenser thus depends on three factors: area and spacing of plates and permittivity of the material between. In the simplest form of condenser, with parallel plates having a total area of A square centimetres, spaced uniformly apart at a distance of D centimetres by material of permittivity K, the capacitance is $\frac{0 \cdot 0885 AK}{D} \times 10^{-12}$ farads. Supposing, for example, the dielectric has a permittivity of 5 and is 0·1 cm. thick, the area necessary to provide a capacitance of 1 farad would have to be about 9 square miles. In practice this unit, denoted by the symbol F, is much too large for convenience, and is divided by a million into microfarads (μF), or even by 10^{12} into micromicrofarads ($\mu\mu$F), sometimes called picofarads (pF).

When a condenser is connected to a source of E.M.F., E volts, a current flows into (not *through*) it, for sufficient time to transfer a charge which raises its potential by the same number of volts. If the E.M.F. ceases, the condenser will discharge through any available path, giving a temporary current in the opposite direction.

If a water analogy is insisted on, a condenser can be regarded

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as a rubber balloon. When connected to a pump (E.M.F.), water flows (current) due to the distention (charging) of the balloon (condenser), and ceases when the back-pressure (potential) of the balloon has risen to equal that which the pump can exert. The larger the balloon (area of condenser plates), and the thinner and more stretchable (greater permittivity) of the material (dielectric), the greater its capacity (capacitance) and the greater the quantity of water (electricity) needed to raise it to a given pressure (voltage). The same amount of water pumped into a smaller balloon would raise it to a higher pressure. If the balloon (condenser) is disconnected and sealed off (insulated), its pressure remains steady. Its pressure cannot be changed unless water flows in or out. If there is a small (high resistance) leak, the pressure slowly falls as the water flows out. If pumped to too high a pressure, the balloon bursts (condenser breaks down) and the pressure drops very rapidly.

21. Practical Forms of Condenser

The construction of a condenser depends on the use to which it is to be put.

A variable tuning condenser, denoted in diagrams as in Fig. 15*a*, consists of two sets of metal vanes which can be

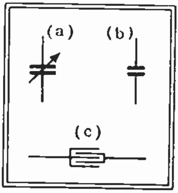

Fig. 15 : Circuit symbols for (a) variable condenser, (b) fixed condenser, (c) alternative symbol for fixed condenser, generally used of a large capacitance.

progressively interleaved with one another, thus increasing the effective area, to obtain any desired capacitance up to the maximum available, usually about 500 pF.

Fixed condensers (Fig. 15*b*) consist of two or more metal plates interleaved with thin sheets of mica (which is a particularly good dielectric), or, if a large capacitance is desired, of two long strips of metal foil separated by waxed paper and rolled up into a compact block. Fig. 15*c* is a symbol used, especially in the older books, to denote a large fixed condenser.

A large capacitance, up to hundreds of microfarads, can be obtained economically by using aluminium plates in a chemical solution or paste which causes an extremely thin insulating film to form on one of them. Apart from this film-forming, the

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solution also acts as a conductor between the film and the other plate. These *electrolytic* condensers must be connected and used so that the terminal marked positive never becomes negative. Even when correctly used there is always a small leakage current.

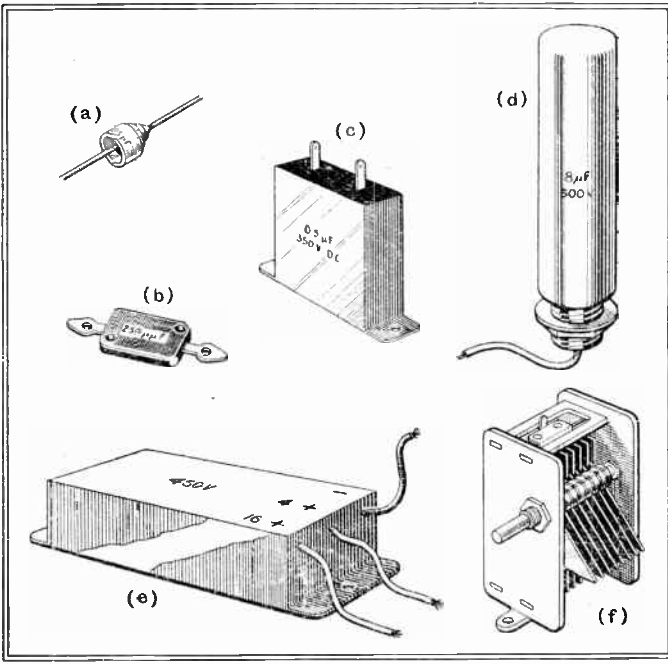

Fig. 16 : Examples of (a) ceramic (b) mica (c) paper (d) wet electrolytic (e) "dry" electrolytic and (f) variable air condensers

In small capacitance sizes, up to about 500 pF, condensers formed by depositing metal on special high-K ceramic material are very satisfactory and stable.

Mica condensers are generally from 50 to 10,000 pF (0·01 μF), and paper condensers 0·01 to 4 μF. The most usual sizes of electrolytics are 8 μF or multiples thereof.

22. The Electrostatic Voltmeter

The attraction between the oppositely charged plates of a condenser reveals itself as an actual physical force which causes them to move towards one another if free to do so. The fact that the attraction is related to the potential difference in volts suggests that it may be used to measure voltage. This is so, and although the electrostatic voltmeter, based on this principle, is not the most commonly used, it has more right to the title of "voltmeter" than most instruments bearing that name. In appearance it resembles a very small variable air condenser, with the moving vanes delicately suspended and kept in the low capacitance position by a hairspring. When a difference of potential is set up across the vanes, the attraction causes the moving vanes to rotate against the restraint of the hairspring until a balance exists. A pointer attached to the moving part shows on a scale the voltage.

Other so-called voltmeters, as we shall soon see, really measure the current driven through a high resistance and thereby arrive at the voltage indirectly through Ohm's Law. Unlike them, the electrostatic voltmeter has the merit of taking no current (apart from a minute and momentary one on switching on), but unfortunately is fairly delicate and impracticable to construct for indicating less than a few hundred volts. It is, however, excellent for reading voltages of several thousand, for which other types are less suitable.

23. Charge and Discharge of Condenser

It is instructive to consider the charging of a condenser rather more precisely than with the crude arrangement shown

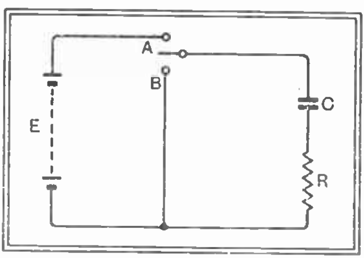

Fig. 17 : Circuit used for obtaining the condenser charge and discharge curves shown in Fig. 18

in Fig. 14. In Fig. 17 the condenser C can be charged by moving the switch to A, connecting it across a battery: and discharged by switching to B. R might represent the resistance of the wires and condenser plates, but as that is generally very small it will be easier to discuss the matter if we assume we

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have added some resistance, enough to bring the total up to, say, 200 ohms. Suppose the condenser is 5 μF and the battery E.M.F. 100 volts.

At the exact moment of switching to A the condenser is as yet uncharged and so there can be no voltage across it; the whole of the 100 volts E.M.F. of the battery must therefore be occupied in driving current through R, and by Ohm's Law that current can easily be found to be 0·5 amp. Reckoning the potential of the negative end of the battery as zero, we have at this stage the positive end of the battery, the switch, the upper plate of the condenser, and (because the condenser being uncharged can have no potential difference across it) the lower plate also, all + 100 volts. The lower end of the resistance is zero, so we have 100 volts drop across the resistance, as already stated. Note that *immediately* the switch is closed the potential of the lower plate of the condenser as well as that of the upper plate jumps from zero to + 100. The use of a condenser to transfer a sudden change of potential to another point, without a conducting path, is very common.

We have already seen that the number of coulombs required to charge a condenser of C farads to V volts is CV. In this case, it is $5 \times 10^{-6} \times 100$ or 0·0005. As a coulomb is an ampere-second, the present charging rate of 0·5 amp., *if maintained*, would fully charge the condenser in 0·001 sec. The condenser voltage would rise steadily as shown by the sloping dotted line in Fig. 18*b*. Directly it starts to do so, however, there are fewer volts available for driving current through R, and so the current becomes less and the condenser charges more slowly. When the condenser is charged to 50 volts, 50 volts are left for the resistance, and the charging rate is 0·25 amp. When the condenser is charged to 80 volts, 20 are left for R, and the current is 0·1 amp. And so on, as shown by the curves in Fig. 18. The upper curve indicates voltage across R if the current scale is multiplied by 200. Note that the voltages across C and R always add up to 100 (as far as 0·004 sec.) and so check up as explained in Sec. 13. *In the time the condenser would take to charge fully at the starting rate it actually reaches 63 per cent. of a full charge.* This is a general rule; and the time in question can be seen by the foregoing explanation to be CR seconds, and is named the *time constant*. The type of curve shown in Fig. 18 is known as *exponential*, and is characteristic of growth and die-away phenomena.

Theoretically the condenser is never completely charged, because the charging depends on there being a difference between the applied E.M.F. and the condenser voltage, to

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drive current through the resistance. For practical purposes the condenser may be as good as fully charged in a very short time. Having allowed 0·004 sec. for this to happen, we move the switch to B. Here the applied voltage is zero, so the voltages across C and R must now add up to zero. We know that the condenser voltage is practically 100, so the voltage across R must be — 100, and the current — 0·5 amp., that is to say, it is flowing in the opposite direction, discharging the condenser. And so we get the curves in Fig. 18 from 0·004 sec. onwards. The upper shaded area represents energy stored in

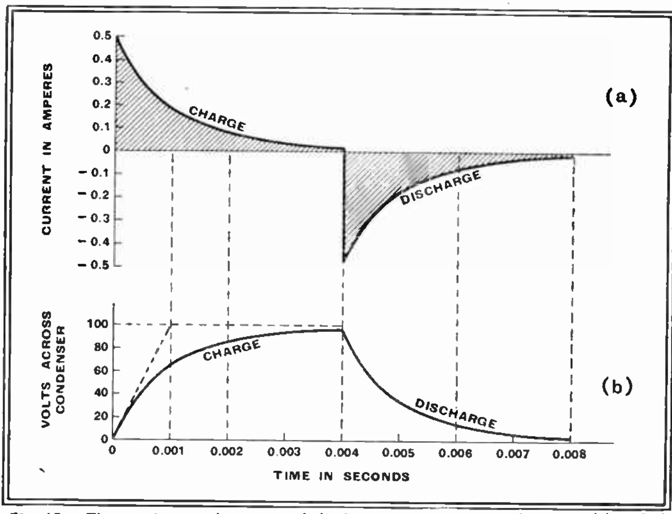

Fig. 18 : The condenser charging and discharging currents are shown at (a) and the resulting voltage across the condenser at (b). The voltage across the resistor is given by (a) if the scale is multiplied by R, i.e., 200

the electric field ; the equal negative area represents this energy being returned as the field collapses.

The significance of this result will be developed in the next chapter. In the meantime it may be noted that the potential we succeeded in conveying to the top end of R via the condenser has been very short-lived. Some idea of its duration can be found by multiplying C by R, giving the time constant.

24. Magnets and Electromagnets

If a piece of paper is laid on a straight " bar " magnet, and iron filings are sprinkled on the paper, they are found to arrange

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themselves in some such pattern as that indicated in Fig. 19*a*. These lines show the paths along which the attraction of the magnet exerts itself, and so are called *lines of magnetic force*. (Compare the lines of electric force in Fig. 14.) As a whole, they map out the *magnetic field*, which is the "sphere of influence," as it were, of the magnet. The field is most concentrated around two regions, called *poles*, at the ends of the bar. The lines may be supposed to be continued internally through the magnet, emerging at the end marked N and returning at S. The total number of lines is called the *flux*.

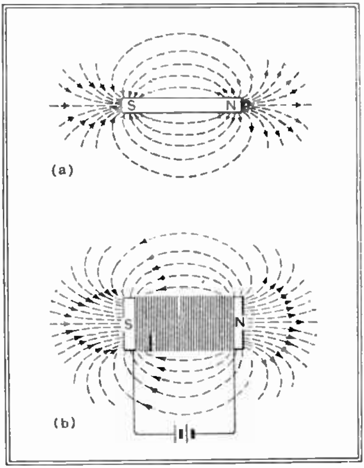


Fig. 19: Conventional representations, in terms of "lines of magnetic force" of the fields round a bar magnet and a coil of wire carrying a current

The same result can be obtained with a previously quite ordinary and unmagnetized piece of iron by passing an electric current through a wire wrapped round the iron; this arrangement, shown in Fig. 20, forms an electro-magnet. It is not even necessary to have the iron; the coil alone, so long as it is carrying current, is interlinked with a magnetic field having the same general pattern as that due to a bar magnet, as can be seen by comparing Fig. 19*a* and *b*. Without the iron core, however, it is considerably weaker. Finally, to reduce the thing to its utter simplicity, if the wire is unwrapped, every inch of it is still found to be surrounded by a magnetic field, though very much less concentrated than in the coil. If the wire is doubled back on itself, so that current is flowing in opposite directions close together, the field is very weak indeed. It goes almost without saying that as the field is caused by the electric current its

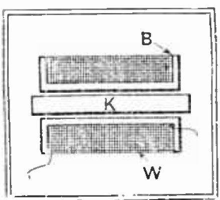


Fig. 20: Section through an electro-magnet. K, iron core; B, bobbin fitting over K; W, winding of insulated copper wire

strength depends on the strength of the current. These facts, which can easily be demonstrated by experiment, illustrate the three things that determine the amount of magnetic flux set up by an electro-magnet, or, more generally, by any electric current flowing anywhere. They are :—

(1) The strength of current.

(2) The shape and dimensions of the current path. Currents flowing in opposite directions tend to cancel out so far as magnetic effects are concerned. But if the wire is wound into the form of a coil, each added turn makes the current flow again in the same direction, adding to the magnetic strength. In fact, with a given size of coil the magnetizing force is proportional to the *ampere-turns* : 10 turns carrying 5 amps. gives the same result as 5,000 turns carrying 10 milliamps. in the same space ; both amount to 50 ampere-turns.

(3) The last factor is the material on which the coil is wound, or with which the wire is surrounded. Certain materials, chiefly iron and steel, have a very large multiplying effect on the magnetic lines, called *permeability*, denoted by the symbol μ (mu). The permeability of most things, such as air, water, wood, brass, etc., is practically 1, but in certain types of iron alloy it may be thousands.

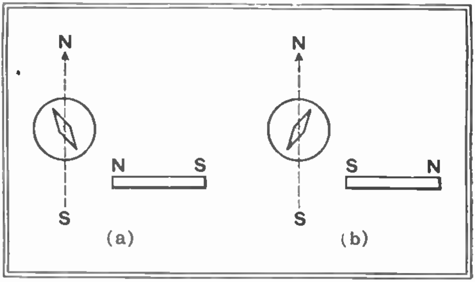
Compare the above three factors with (1) voltage, (2) plate shape and size, and (3) permittivity in the case of *electric* fields.

The magnetic (and electric) lines are, of course, as imaginary as lines of latitude and longitude, and it is a matter of arbitrary agreement that a certain number is said to be produced by a certain current flowing through a circuit of certain dimensions. It is also important to realize that the field occupies the whole space ; the lines indicate the direction and intensity of the field, but are not meant to convey the idea that the field itself is in the form of lines, with no field in the spaces between.

25. Interacting Magnetic Fields

The " N " and " S " in Fig. 19 stand for North-seeking and South-seeking respectively. Everybody knows that a compass needle points to the north. The needle is a magnet, and turns because its own field interacts with the field of the earth, which is another magnet. Put differently, the north magnetic pole of the earth attracts the north-seeking pole of the needle, while the earth's south pole attracts its south-seeking pole. The poles of a magnet are often called, for brevity, just north and south, but strictly, except in referring to the earth itself, this is incorrect. By bringing the two poles of a bar magnet—previously identified by suspending it as a compass—in turn towards a compass needle it is very easy to demonstrate that, as in Fig. 21, like poles repel one another and unlike poles attract. This reminds one, again, of the way electric charges behave.

Fig. 21: Deflection of a compass needle by a magnet, proving that like poles repel and unlike poles attract. The N and S poles of the bar magnet may be found by suspending it like a compass needle and marking as N that pole which turns to the north



Now suppose we hang a coil in such a way that it is free to turn, as suggested in Fig. 22. So long as no current is passing through the coil it will show no tendency to set itself in any

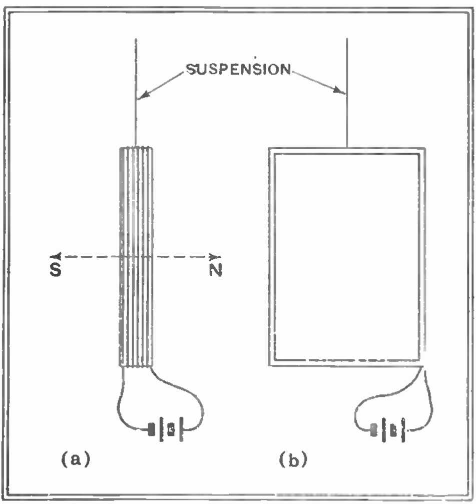


Fig. 22: A coil, free to rotate, sets itself in the orientation indicated, when a current is passed through the winding. (a) is a side view, and (b) an end view

particular direction, but if a battery is connected to it the flow of current will transform the coil into a magnet. Like the compass needle, it will then indicate the north, turning itself so that the plane in which the turns of the coil lie is east and

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west, the axis of the coil pointing north. If now the current is reversed the coil will turn through 180 degrees, showing that what was the N pole of the coil is now, with the current flowing the opposite way, the S.

The earth's field is weak, so the force operating to turn the coil is small. When it is desired to make mechanical use of the magnetic effect of the current in a coil it is better to provide an artificial field of the greatest possible intensity by placing a powerful magnet as close to the coil as possible. This is the basis of all electric motors. In radio, however, we are more interested in other applications of the same principle.

26. Current Measuring Instruments

The tendency of a coil to turn when it is carrying a current depends on the intensity of the magnetic field in which the

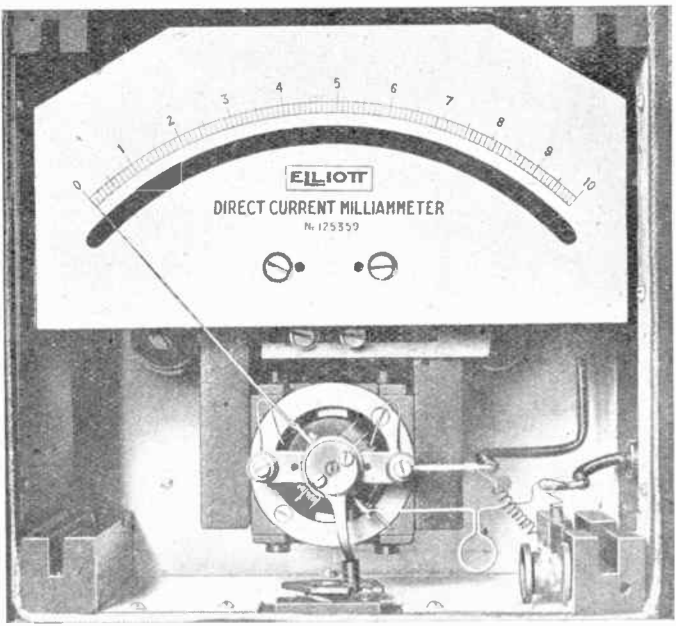

Fig. 23 : Milliammeter with front removed The coil lies behind the pivot between the poles of the magnet. When moved by a current, it sweeps the pointer over the scale

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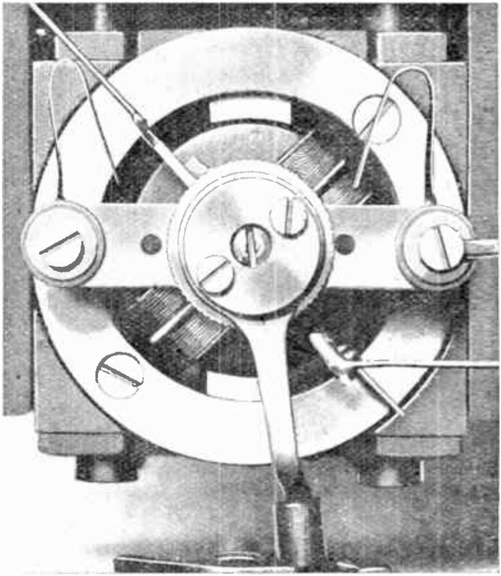

Fig. 24: Enlarged view of pivot of milliammeter. The coil can be seen here; it moves at right-angles to the pointer. Note balancing-weight on the latter

coil lies, and also upon the ampere-turns available to provide the coil's own field. If the intensity of field and the number of turns are fixed, the force with which the coil is rotated depends only on the current passing through it; if the coil, in turning, is compelled to wind up a light spring, the extent of rotation will be a measure of the current causing it.

This is the principle of the *moving-coil* meter. The current to be measured is made to pass through a coil of wire suspended on light bearings between the poles of a permanent magnet (Figs. 23 and 24). The magnetic field set up by the current through the coil is acted upon by the field of the magnet, the two being so arranged that the resulting force tends to rotate the coil. It is free to do so until it has wound up a hairspring enough to exert an equal back force, and carries with it a pointer moving over a scale graduated in current. The instrument is called an *ammeter* or *milliammeter*, according to whether it reads respectively amperes or milliamperes. As the strength of the coil's field is proportional to its ampere-turns, the instrument can be made to cover the desired range of current

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by employing a suitable number of turns. If it is desired to measure a current beyond the range of the instrument it can be done by passing a known fraction of the current through the coil. Suppose the resistance of the coil is 15 ohms and measures a maximum of 5 milliamps. Then if in parallel with the coil is connected a resistance (called a *shunt*) also of 15 ohms, when the meter reads 5 mA the current is really 10 mA because another 5 are going through the shunt. In a multi-range meter a number of shunts are available.

In order that the current in a circuit may not be appreciably altered when the meter is inserted to read it, an ammeter or milliammeter always has a low resistance.

The voltage between two points can be measured by observing

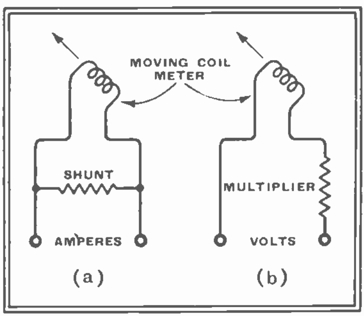

Fig. 25 : Showing how the same moving coil instrument can be used to measure current or voltage

the current that flows when a known resistance is connected between them and then using Ohm's Law. To avoid appreciable alteration in the voltage due to the current so drawn, the resistance is made as high as possible, so the instrument must be a sensitive milliammeter (or even a microammeter). If scaled in milliamps, all readings would have to be multiplied by the value of the resistance (in thousands of ohms). In practice the maker of the instrument does this once and for all, and marks the scale in volts, so justifying calling this current meter a *voltmeter*. The range of a voltmeter of this type can easily be increased by using a higher resistance in series, sometimes called a *multiplier*.

Thus a single moving-coil instrument can be made to cover a large number of ranges of current and voltage, simply by connecting appropriate shunts and multipliers (Fig. 25).

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If a battery is used, of sufficient voltage to give full scale reading, the extent to which the reading is reduced by either shunt or series resistance depends on the value of that resistance. It is therefore possible to graduate the instrument in ohms, and we have an *ohmmeter*.

27. Induction

If two moving-coil meters, A and B, are connected together, so that there is a complete circuit through the two coils, then by taking hold of the pointer of A and turning the coil round with it, the pointer of B will indicate a current. This current lasts as long as A's coil is being moved, and stops when the coil stops. Allowing the coil of A to spring back causes B to deflect in the opposite direction. The strength of current is proportional to the rate at which the coil is moved.

The current is driven by an E.M.F. generated by the movement of the coil in a magnetic field. It need not be a coil ; any conductor moved in a magnetic field has an E.M.F. generated in it, so long as it is not being moved *along* the lines of force. It has to be moved *across* them. The E.M.F., in fact, is proportional to the number of lines of force (or rather the field represented by those lines) that the conductor cuts across per second.

To generate an E.M.F. all that is necessary is some device for moving wires across a magnetic field. It is equally effective if the field is moved across the wires. All the dynamos or generators in power stations do either one or other.

But it is not necessary for either to move. If the magnetic field is produced by an electromagnet, varying the current that energizes the magnet varies the number of lines of force linked with any conductor in the field, even though it be stationary. Thus suppose a number of wires are lying alongside one another. If a current is passed through one of them a magnetic field is set up, and most of its lines of force will embrace the other wires also. So long as the current is steady, nothing will happen. But if the current is varied, an E.M.F. will be generated in *all* the wires, including the one or more responsible for the varying magnetic field. The E.M.F. is said to be *induced* by the varying or moving magnetic field.

28. Self-Inductance

As every circuit carrying current has some magnetic field it follows that an E.M.F. is induced in *any* circuit in which the current is varied. The extent to which an E.M.F. is induced in a circuit (or part of a circuit) when the current in that circuit

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is varied at some standard rate is called its *self-inductance*, or often just "inductance."

We must have a unit with which to specify the amount of inductance in any particular case; it is called the *henry*, and it has been made to line up with the units we already know by defining the self-inductance in henries as being equal to the number of volts induced in a circuit when the current in that circuit is changing at the rate of 1 ampere per second.

Actually it is the change of *flux* that induces the E.M.F., but information about the flux is less likely to be available than that about the current, so inductance is specified in terms of current. What it depends on, then, is the amount of flux produced by a given current, say 1 amp. The more flux, the greater the E.M.F. induced when that current takes one second to grow or die, and so the greater the inductance. The flux produced by a given current depends, as we have seen, on the dimensions of the circuit or part of circuit, and on the permeability. A circuit such as a coil, with a great number of turns of large diameter, filled with high-permeability iron, sets up a very large flux when carrying one amp.; consequently a very high voltage is induced when the current varies at the rate of one amp. per second. In other words, the inductance is large. A single small turn causes only a very small flux when one amp. flows, so its inductance is small. All the same, it may be quite important *if the current is changing at a very high rate*. The maximum current may even be much less than one amp. but if it comes and goes a thousand million times a second its rate of change may well be millions of amps. per second and the induced voltage quite large.

There is one other important question about the induced voltage—what is its direction? There are two possibilities: it might be in such a direction as to assist or to oppose the change of current that caused it. If it were to assist, the current would change more rapidly, inducing a greater voltage, increasing the change of current still more, and there would be no limit to the thing. Obviously that is absurd; and in fact the induced voltage always tends to *oppose* the change of current responsible for it.

29. Growth of Current in Inductive Circuit

What happens when current is switched on in an inductive circuit can be studied with the arrangement shown in Fig. 26 Compare it with Fig. 17. To make the comparison easy, let us assume that the inductance, for which L is the standard

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symbol, is 0·2 henry, and that R (which includes the resistance of the coil) is 200 ohms and E is 100 volts as before.

With the switch as shown, no current is flowing through L and there is no magnetic field. At the instant of switching to A, the full 100 volts is applied across L and R. The current cannot instantly leap to 0·5 amp., which is the amount indicated by Ohm's Law, for that would mean increasing the current at an infinite rate, which would induce an infinite voltage opposing it. So the current must rise gradually, and at the exact moment of closing the circuit it is zero. There is therefore no voltage drop across R, and the battery E.M.F. is opposed solely by the inductive E.M.F., which must be 100 volts. That enables us to work out the rate at which the current will grow. If L

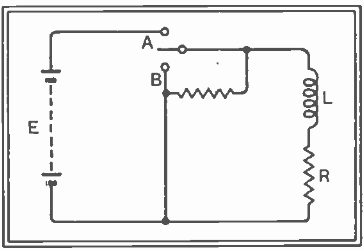

Fig. 26 : Circuit used for studying the rise and fall of current in an inductive circuit

were 1 H, it would have to grow at 100 amps. per sec. to induce 100 volts. As it is only 0·2 H it must grow at 500 amps. per sec.

Drawing a graph (Fig. 27) as we did for the capacitive circuit, the dotted line represents a steady current growth of 500 amps. per sec. If it kept this up, it would reach the full 0·5 amp. in 0·001 sec. But directly the current starts to flow it causes a voltage drop across R. And as the applied voltage remains 100, the induced voltage must diminish. The only way that can happen is for the current to grow less rapidly. By the time it has reached 0·25 amp. there are 50 volts across R, therefore only 50 across L, so the current must be growing at half the rate. The full line shows how it grows; here is another exponential curve—compare Fig. 18—and, as in the capacitive circuit, it theoretically never quite finishes. In the time that the current would take to reach its full Ohm's Law value if it kept up its starting pace, it actually reaches 63 per cent. of it. This time is, as before, called the time constant.

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A little elementary algebra based on the foregoing shows it to be L/R seconds.

The voltage across L is also shown. Comparing with Fig. 18 it is seen that current and voltage have changed places. The voltage across R is of course proportional to the current and therefore its curve is the same shape as the lower one. Added to the upper, it must equal 100 volts as long as 100 volts is applied.

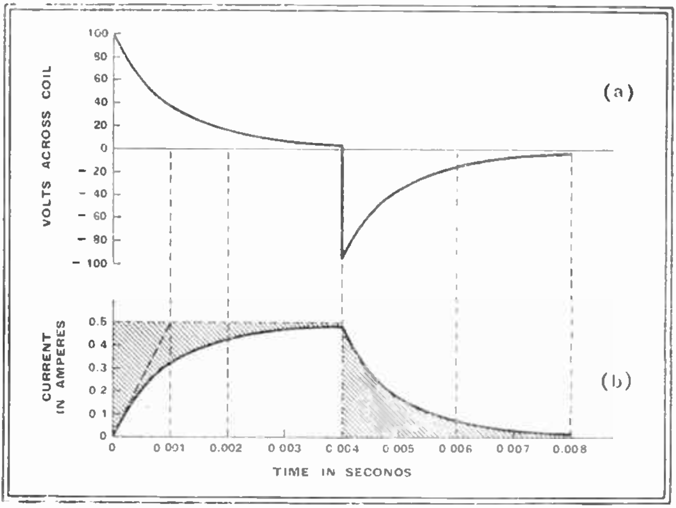

Fig. 27 : The voltage across the coil L in Fig. 26, and the current through it, are shown in relation to time. At the start, the switch is moved to A, and 0·004 seconds later is moved to B

But when the current has reached practically its full value, we flick the switch instantaneously across to B. The low resistance across the contacts is to prevent the circuit from being completely interrupted in this process. At the moment the switch touches B, the magnetic field is still in existence; it cannot be eliminated until the current dies out, and that cannot happen without inducing an E.M.F. that tends to prevent it from dying out. Before the current has had time to change, the voltage drop across R remains at 100, so the current must diminish at such a rate as to induce 100 volts, that is to say, at 500 amps. per sec. As the current wanes, so does the voltage across R, and so must the induced voltage,

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and therefore the current dies away more slowly, as shown in the continuation of Fig. 27.

Comparing with Fig. 18 again the first shaded area represents the energy stored in the magnetic field (without L the current would rise at once to 0·5 amp. and remain steady at that), while the second shaded area represents this same energy being used up in heating R by passing current through it after the battery E.M.F. has been withdrawn.

30. Mutual Inductance

If a coil with the same number of turns as L in Fig. 26 were wound so closely around it that practically all the magnetic field due to the *primary winding* embraced also this *secondary winding*, then an equal voltage would at all times be induced in the secondary. Generally it is impracticable to wind so closely that the whole field links or couples with both coils, and the secondary voltage is less, according to the looseness of magnetic coupling. But if the secondary has *more turns* than the primary, it is possible to induce a greater voltage in the secondary than in the primary. Remember that this voltage depends on the *varying* of the primary current. A device of this kind, for stepping voltages up or down, or for inducing voltages into circuit without any direct connection, is rather inappropriately named a *transformer*. It is represented in circuit diagrams by two (or more) coils drawn closely together. An iron core is shown by a few straight lines drawn between them (Fig. 28).

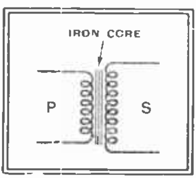

Fig. 28 : Symbol representing an iron-cored step-up transformer

The effect that one coil can exert, through its magnetic field, on another, is called *mutual inductance*, and like self-inductance is measured in henries. The definition is similar too: the mutual inductance in henries between two coils is equal to the number of volts induced in one of them when the current in the other is changing at the rate of 1 amp. per second.

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CHAPTER 4

ALTERNATING CURRENTS

31. Frequency of Alternating Current

IN wireless we are habitually making use of *alternating* currents; that is to say, currents that are constantly reversing their direction of flow. The currents produced by the action of the microphone, and corresponding to the waves set up by speech or music, are of this type. So are the currents used to carry the programme across long distances. And so, too, are those used to light and heat our homes and provide the power (in place of the more expensive and troublesome batteries) that work our wireless sets. The only basic difference between all these is the number of double-reversals the current makes in a second; in a word, the frequency (Sec. 5).

There is no hard-and-fast division between one lot of frequencies and another; but those below 100 cycles per second (c/s) are used for power (the standard in this country is 50 c/s); those from 25 to 10,000 or so are audible, and so are classed as audio frequencies (A.F.), while those above about 20,000 are more or less suitable for carrying signals across space, and are known as radio frequencies (R.F.). Certain of these, notably 550,000 to 1,550,000, are allocated for broadcasting. For classification of radio frequencies see p. 332.

What has been said about currents applies to voltages, too; it requires an alternating voltage to drive an alternating current.

32. The Sine Wave

It is easy to see that an alternating current might reverse abruptly or it might do so gradually and smoothly. In Fig. 18*a* we have an example of the former. There is, in fact, an endless variety of ways in which current or voltage can vary with time, and when graphs are drawn showing how they do so we get a corresponding variety of wave shapes. The steepness of the wave at any point along the time scale shows the rapidity with which the current is changing at that time.

Fortunately for the study of alternating currents, all wave shapes, however complicated, can be shown to be built up of waves of one fundamental shape, called the *sine wave* (adjective: sinusoidal). Fig. 29 is a specimen. Note its smooth, regular alternation, like a pendulum swinging. Waves of more com-

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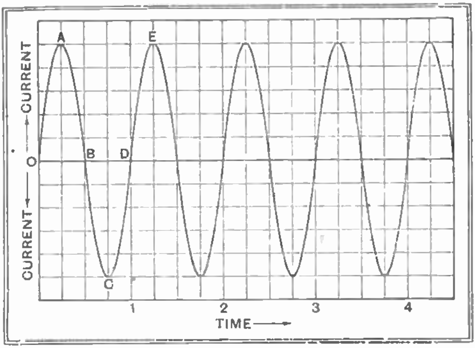

Fig. 29 : A graph showing how a sinusoidal current varies with time.

plicated appearance, such as that of Fig. 18, can be analysed like a number of sine waves of different frequency.

One cycle (Sec. 5) is completed at the moment when the current begins to repeat the same variation. Beginning at O, the end of a cycle is at D ; beginning at A it is at E. Fig. 29 shows $4\frac{1}{2}$ cycles. As we already know, the number of cycles per second is the frequency.

33. R.M.S. and Peak Values

In Fig. 30, G is assumed to be a generator of alternating voltage, driving an alternating current through the resistance R. We have already seen (Sec. 29) that a resistance offers opposition to the flow of current, but is indifferent to changes in that current. Put differently, the sudden application or withdrawal of a voltage produces *instantaneously* a current of the magnitude that Ohm's Law predicts. If therefore the full-line curve of Fig. 31 is taken to represent the variations of the voltage of G with time, the current will follow a curve identical in shape, though different in scale, for the current at any instant will be equal to the voltage divided by the resistance. The current curve is shown dotted.

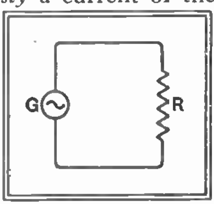

Fig. 30 : An alternating voltage is applied by the generator G (nature unspecified) to the resistance R

The alternating current supplied for

house lighting has, as a general rule, a frequency of 50 cycles per second (50 c/s). In terms of Fig. 29, this means that the time-scale is such that the distance A to E or O to D represents one-fiftieth of a second. Each second thus contains 50 current-pulses in each direction, so that if the temperature of its filament could change quickly enough a lamp connected to such mains would not emit a continuous light, but a series of separate flashes succeeding one another at the rate of 100 per second.

How are electric mains that behave in this fashion to be rated ?—that is, what are we going to mean when we speak of "200-volt 50-cycle mains" seeing that the actual voltage is changing all the time, and is sometimes zero ?

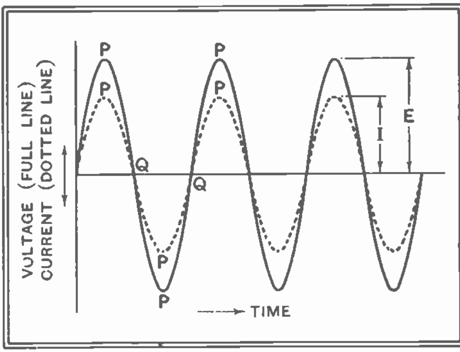

Fig. 31 : Voltage and current relationships in the circuit of Fig. 30. The average power over one complete cycle is ½EI, or half the power developed when E and I both have maximum values

The convention that has been arrived at is based on comparison with direct-current (D.C.) mains. It is obviously going to be a great convenience for everybody if a lamp or a fire intended for a 200-volt D.C. system should be equally suited to alternating mains of the same nominal voltage. This condition will be fulfilled if the *average power* taken by the lamp or fire is the same for both types of current, for then the filament will reach the same temperature and the cost of running will be the same in the two cases.

In Fig. 31 both voltage and current are shown for a resistive circuit. At any instant the power being consumed is given by multiplying voltage by current. At the instant corresponding to P both are at their maximum value, and the power is also at its highest. At Q voltage and current are both zero ; so also

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is the power. The average power must lie somewhere between these extremes.

If by "200-volt mains" we mean a supply whose *peak* voltage (point P) is 200 volts the maximum instantaneous power drawn by a lamp or fire would be the same as the power it would take from 200 volt D.C. mains, but the *average* power would be less. To raise this to the figure for D.C. mains, the peak voltage of the A.C. supply will evidently have to rise well above the rated nominal voltage.

It can be shown mathematically that for a sine waveform the average power is exactly half of that corresponding to the instant P; it is therefore half EI where E is the peak voltage and I the peak current. We need, therefore, to raise the peak voltage sufficiently far above the nominal voltage to double the peak power.

We cannot do this by simply doubling E, because this also doubles I, making the peak power four times as great. To double the power we have to increase the peak voltage $\sqrt{2}$ times, which simultaneously causes the peak current to increase $\sqrt{2}$ times. The increase of power is then to $\sqrt{2} \times \sqrt{2}$ times, or double, its original value, which is what is required.

Alternating mains equivalent to 200-volt D.C. mains must therefore rise to a peak of $200\sqrt{2} = 282{\cdot}8$ volts. Such mains are described as having a R.M.S. (root-mean-square) voltage of 200.

If a fire of 40 ohms resistance is connected to such mains the R.M.S. current will be 200/40 = 5 amps., and the power consumed will be EI = 200 × 5 = 1,000 watts. Although the power is rapidly varying between a peak value of 2,000 watts and zero, the average power consumed and consequently the heat to which it gives rise will be exactly the same as if the same fire were connected to 200-volt D.C. mains.

Using in this way R.M.S. values for voltage and current we can forget entirely the rapid variations taking place, and *so long as our circuits are purely resistive* all calculations dealing with alternating current can be carried out according to the rules already discussed in connection with ordinary direct current.

34. A.C. Meters

If alternating current is passed through a moving coil meter (Sec. 26) the coil is urged first one way and then the other, because the current is reversing in a steady magnetic field. The most one is likely to see is a slight vibration about the zero

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mark. Certainly it will not indicate anything like the R.M.S. value of the current.

If, however, the direction of magnetic field is reversed at the same time as the current, the double reversal makes the force act in the same direction as before, and the series of pushes will cause the needle to take up a position that will indicate the value of current. The obvious way to obtain this reversing magnetic field is to replace the permanent magnet by a coil and pass through it the current being measured.

When the current is small, the field also is very weak, and the deflection too small to be read ; so this principle is seldom used, except in *wattmeters*, in which the main current is passed through one coil and the other—the " volt coil "—is connected across the supply, i.e., in parallel with it.

A more usual type is that in which there is only one coil, which is fixed. Inside are two pieces of iron, one fixed and the other free to move against a hairspring. When either direct or alternating current are passed through the coil, both irons are magnetised in the same polarity, and so repel one another, to a distance depending on the strength of current. These *moving iron* meters are useful when there is plenty of power to spare in the circuit for working them, but tend to use up too much in low-power circuits.

Another method is to make use of the heating effect of the current. When a junction of two different metals is heated a small unidirectional E.M.F. is generated, which can be measured by a moving coil meter. Instruments of this kind, called *thermojunction* or *thermocouple* types, if calibrated on D.C. obviously will read R.M.S. values of A.C. regardless of waveform. They are particularly useful for much higher frequencies than can be measured with instruments in which the current to be measured has to pass through a coil.

The electrostatic instrument (Sec. 22) can be used for alternating voltages.

But perhaps the most popular method of all is to convert the alternating current into direct by means of a *rectifier*—a device that allows current to pass through it in one direction only—so that it can be measured with an ordinary moving coil meter. The great advantage of this is that by adding a rectifier a multi-range D.C. instrument can be used on A.C., too. Most of the general-purpose test meters used in radio are of this kind. Because the moving coil instrument measures the *average* current, which in general is not the same as the R.M.S. current, the scale is marked off in such a way as to take account of the factor necessary to convert one to the other. This is approxi-

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mately 1·11 for sine waves, but is different for other waveforms, so the rectifier type of meter reads the latter incorrectly.

35. Capacitance in A.C. Circuits

The behaviour of a condenser towards an alternating voltage can be seen by considering once more the circuit shown in Fig. 17, but with one slight modification; half the battery is removed from the A branch and connected in the B branch the opposite way round, so that if the switch is moved alternately to A and B at equal time intervals it will provide an alternating voltage of square waveform as shown in Fig. 32.

Suppose also that an A.C. ammeter is connected in series with the condenser. Our circuit is now as in Fig. 33.

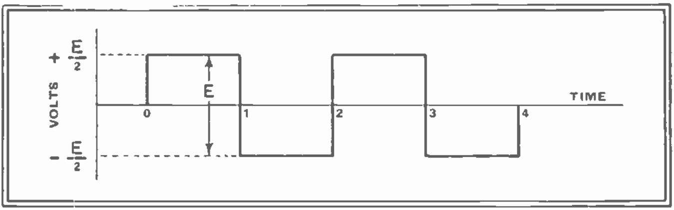

Fig. 32 : Alternating voltage produced by means of the apparatus shown in Fig. 33

If the switch is moved very slowly, so that the time of one cycle is long compared with the time constant, the meter shows a kick when switched to A, and another when switched to B. The average current is small.

By speeding up the switch, increasing the frequency, there is never time for the current to fall to anywhere near zero, and the sluggishness of the meter causes its pointer to give a more or less steady reading.

We shall then have evidence of a current flowing, apparently continuously, in a circuit which is broken by the insulating material between the plates of the condenser. But, as the way in which the current has been built up clearly shows, electrons are flowing *in and out* of the condenser, and not *through* it in the ordinary sense of the word.

During each momentary burst of current the flow is greatest at the beginning and tails off towards the end, as the curve of Fig. 18 shows. The more rapid the vibration of the switch, therefore, the greater is the proportion of the total time during which the current is high, and the greater, in consequence, is the average current read on the meter.

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We conclude that the higher the frequency of the applied voltage, the greater will be the current in a circuit containing capacitance. It is easy to see, too, that, if the capacitance is greater, a greater number of electrons is required to perform the alternating charges and therefore the current is greater.

Instead of taking this rapidly reversing current from a battery and a mechanical switch, it can be drawn from any normal source of alternating current, such as the electric light

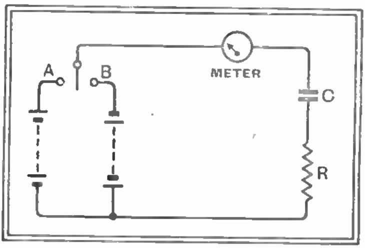

Fig. 33 : By moving the switch backwards and forwards between A and B an alternating voltage can be produced and the resulting current indicated by M shows the behaviour of a condenser in an A.C. circuit

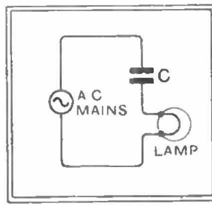

Fig. 34 : If a condenser of large capacitance (2 μF or more) is placed in series with a lamp lighted from A.C. mains, the lamp will light, though with less than normal brilliance

mains. If, as suggested in Fig. 34, a lamp (40-watt is recommended for the experiment) is connected to A.C. mains through a condenser of capacitance some 2μF or more, the lamp will light, and stay alight. But its brilliance will be below normal.

Without the condenser, the alternating current drives electrons to and fro in the lamp filament ; with the condenser in circuit, the elastic opposition of the electrons in the dielectric restricts, to some small extent, the number of electrons that can so move at each change in direction of the voltage.

36. Reactance of a Condenser

As has just been indicated, the obstruction offered by a condenser to the flow of current depends upon its capacitance and upon the frequency of the current, becoming less as either of these rises. If an alternating E.M.F. of R.M.S. voltage E at a frequency f cycles per second is applied to a condenser of capacitance C farads the current flowing through it is $E \times 2\pi fC$ amperes R.M.S., where π is the ratio of the circumference of a circle to its diameter and has a numerical value of 3·1416 or 22/7 approximately. This way of calculating the current is like Ohm's Law except that $1/2\pi fC$ takes the place of R.

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Therefore it is reasonable to measure the quantity of $1/2\pi fC$ in ohms, and to distinguish it from resistance it is called *reactance* and is denoted by the letter X. So our "Ohm's Law" for a circuit consisting of an alternating E.M.F. E of frequency f c/s and a condenser C farads is $I = E/X$, remembering that $X = 1/2\pi fC$.

In the case of the $2\mu F$ condenser of Fig. 34, the reactance to 50-cycle current will be $1/2\pi fC = 1/2\pi.50.2.10^{-6} = 10^6/200\pi = 1590$ ohms.

It is particularly to be noted that the electricity pushed into the condenser at one instant bounces out again the next; the passing of an alternating current through a condenser *does not involve the expenditure of energy.* Such energy as is required to charge the condenser may be regarded as an extremely short-term loan, which is repaid in the next half-cycle, and so the net expenditure taken over any complete cycle is nil. If a resistance were used in place of C in Fig. 34 it would get hot, showing that this method of dimming the lamp diverts some of the unwanted energy to the resistance and there wastes it. Equal dimming by using a condenser wastes no power, as can be shown by the fact that C remains stone cold.

It is for this reason that its opposition to the current is called reactance, to distinguish it from resistance, the passing of current through which always involves the expenditure of energy.

37. Inductance in A.C. Circuits

It will be remembered (Sec. 29) that the effect of inductance is to delay the rise or fall of current in a circuit, this being due to the formation or collapse of a magnetic field. It is easy to see that the same delaying effect is present when the applied voltage attempts to *reverse* the current, as an alternating voltage is constantly doing.

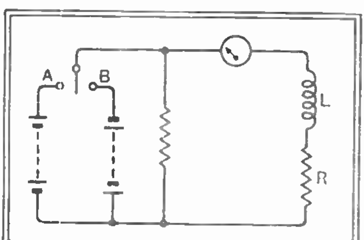

Fig. 35 : Circuit for showing the behaviour of an inductance in an A.C. circuit

Let us apply the battery-made alternating voltage that we used in Fig. 33 to a coil instead of a condenser (Fig. 35). The resistance in parallel with the batteries is merely to avoid complications due to the momentary break in the circuit when the switch is flipped across from one contact to the other.

At a very low frequency of switching the time taken for the growth of current in each direction will be negligible compared with the time of steady flow, and the average current will be practically that which the resistance alone would take from the battery. At a higher speed the current might reach only a fraction of that amount (see Fig. 27*b*) by the time the voltage reversed, and the average value would be less. At a very high frequency the current would be hardly measurable. It is clear that the greater the frequency of reversal the less will be the current.

Compare this with the current through a condenser, which becomes *greater* when the frequency is increased.

When an alternating voltage is applied to a coil of negligible resistance the current that flows will be decided both by the frequency of that voltage and by the inductance of the coil, decreasing as either of these is raised. The current in amps. is actually $E/2\pi fL$, where f is the frequency in cycles per second and L the inductance in henries. The place of R in Ohm's Law is taken by $2\pi fL$, which is called the reactance of the coil.

So there are two kinds of reactance—capacitive and inductive—often denoted by X_C and X_L respectively. X is the general symbol for reactance.

As in the case of the condenser, no power is consumed by driving an alternating current against the opposition that inductive reactance represents, because the energy put into the magnetic field in building it up is restored to the circuit when it collapses. The resistance, r, of the wire with which the coil is wound involves, of course, the usual consumption of energy, being I^2r, where I is the current flowing.

An inductance consists in most cases of a coil of wire. As a tuning coil it is usual to wind the wire on a tubular former of bakelite or paxolin ; some 100 turns of wire on a former of $1\frac{1}{2}$in. diameter provide 170 μH or thereabouts for tuning over the medium wave band. Alternatively it may be of the self-supporting "wave-wound" type. A radio-frequency choke, of inductance perhaps 200,000 μH, will generally be wound of many turns of fine wire on a slotted former. Such a choke will offer a reactance of 1·26 megohms at $f = 1{,}000$ kc/s, while having a reactance of only 6,290 ohms at 5,000 cycles per second. Such a component is called a radio-frequency choke for the rather obvious reason that it opposes, or chokes back, the flow of currents of radio-frequency, while allowing those of audio-frequency a relatively unimpeded passage.

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If it is necessary to offer considerable reactance to currents of quite low frequency, it is evident that a much higher inductance than this is necessary. To obtain high inductance without excessive resistance the coil is wound round a core of iron, or iron alloy, which offers a much easier passage than air to the lines of magnetic force, and so, by increasing the magnetic field, puts up the inductance which is a manifestation of that field (Sec. 24).

38. Condensers in Parallel or Series

We have already seen (Sec. 20) that capacitances in parallel add up just like resistances *in series*.

We can arrive at the same conclusion by simple algebraic reasoning based on the behaviour of condensers to alternating current. Fig. 36 represents an A.C. generator of voltage E

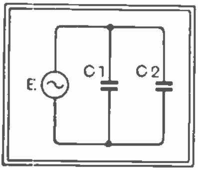

Fig. 36: An A.C. generator applied to two condensers in parallel. Each condenser takes its own charging current without respect to the other. Since condensers are rated by their charging current, the capacitance equal to these two in parallel is given by $C = C_1 + C_2$

connected to two condensers in parallel having capacitances of values C_1 and C_2. The separate currents through the condensers are respectively $E \times 2\pi f C_1$, and $E \times 2\pi f C_2$. The total current is thus $E \times 2\pi f (C_1 + C_2)$, which is equal to the current that would be taken by a single condenser of capacitance equal to the sum of the separate capacitances of C_1 and C_2.

Section 20 also showed that if a single condenser is divided by a plate placed midway between its two plates, the capacitance between the middle plate and either of the others is twice that of the original condenser. In other words, the capacitance of two equal condensers in series is half that of each of them. Let us now consider the more general case of any two capacitances in series.

If X_1 and X_2 in Fig. 37 are respectively the reactances of C_1 and C_2, their combined reactance X is clearly $(X_1 + X_2)$, as in the case of resistances in series. By first writing down the equation "$X = X_1 + X_2$", and then replacing each "X" by its known value, of form $1/2\pi f C$, it is easy to see that $1/C = 1/C_1 + 1/C_2$.

That is, the sum of the reciprocals of the separate capacitances is equal to the reciprocal of the total capacitance. As in the case of resistances (Sec. 17) this equation can be put in a more manageable form— $C = C_1 C_2/(C_1 + C_2)$.

Observe that the rule for capacitances in series is identical in form with that for resistances or reactances in parallel. From

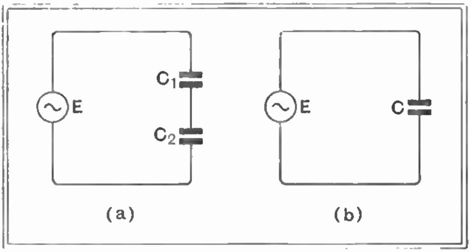

Fig. 37: If the one condenser C in *b* is to take the same current as the two in series at *a*, its capacitance will be given by $1/C = 1/C_1 + 1/C_2$

the way the rule was derived it is evidently not limited to two capacitances only, but applies equally to three, four, or more, all in series.

The rule implies that if two or more condensers are connected in series, the capacitance of the combination is always less than that of the smallest.

39. Inductances in Series or Parallel

The case of inductances in combination is quite straightforward. Several in series have a total inductance equal to the sum of their separate inductances, while if connected in parallel they follow the " reciprocal law ", so that $1/L = 1/L_1 + 1/L_2 + 1/L_3 + \ldots$ . . In either case they combine just as do resistances or reactances. The reader can verify this for himself by adding reactances when they are in series, and adding currents when they are in parallel.

It should be noted that if L_1 and L_2 are placed in series their total inductance will be $(L_1 + L_2)$ only on the condition that the field of neither coil affects the other. If the coils interact so that their mutual inductance is M. the total inductance will be $(L_1 + L_2 + 2M)$ or $(L_1 + L_2 - 2M)$, " 2M " being added or subtracted according to whether the placing of the coils causes their separate magnetic fields to assist or oppose one another.

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CHAPTER 5

A.C. CIRCUITS

40. Phase-Relations between Voltage and Current

IN the previous chapter we have obtained a general picture of how an alternating current can pass through an inductance or capacitance without dissipating electrical energy in the form of heat, as it would do if passed through resistance. This is because the current alternately sets up a magnetic or electric field, so storing up a certain amount of energy, which is then returned to the generator, driven by the collapse of that field. With no resistance, the energy given back is exactly equal to that taken in establishing the field. In practice there always is some resistance; even if there are no actual resistors the turns of wire in a coil (and to a less extent the plates and connecting wires, or leads as they are called, in a condenser) offer some resistance to the flow of current. In order to see how combinations of resistance, capacitance, and inductance make possible a great variety of useful results we have to study the precise relationships of alternating voltage and current in such circuits, and especially time or *phase* relationships.

41. Resistance: E and I in Phase

In Fig. 38 the upper full line represents a sinusoidal alternating E.M.F of R.M.S. value 1 volt. Rather more than one complete cycle is shown, and for convenience in reference each cycle is shown divided in the conventional way into 360 parts, corresponding to the 360 degrees of angle into which a circle is divided.

If this voltage is applied to a 2-ohm resistance the current will be $E/R = 0{\cdot}5$ amp. R.M.S. We have already seen that in a circuit consisting of pure resistance the current adapts itself instantaneously to changes in voltage; we may therefore apply Ohm's Law to each momentary voltage all through the cycle. By doing this we arrive at the dotted curve, which shows the current in the circuit at every instant. At the beginning of the cycle (at 0°) the voltage and current are both zero; at 90°, the end of the first quarter-cycle, both are at their maximum in a positive direction, dropping again to zero halfway through the cycle (at 180°), and rising again to a maximum in the negative direction.

Having thus drawn out voltage and current separately for each instant, we can calculate, by simple multiplication of one

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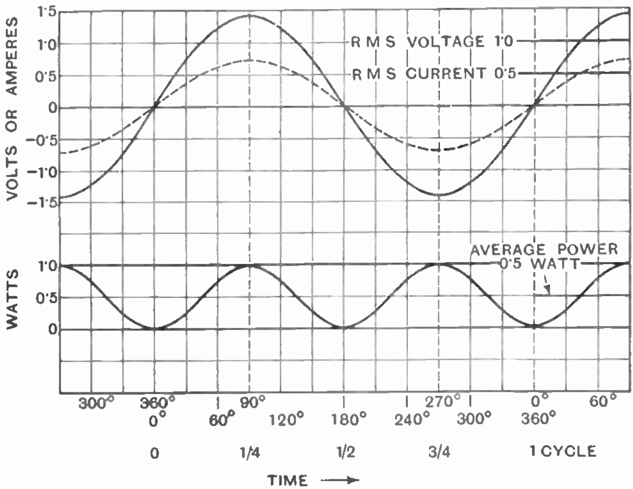

Fig. 38 : Relation of voltage, current, and power in a purely resistive circuit

by the other, the power being consumed. At 0°, for example, E and I are both zero ; so therefore is the power. At 30° E = 0·707, I = 0·353 ; hence the power EI, is 0·25 watt. Proceeding in this way for a number of points distributed over the first 180° of the cycle we find that the power rises to a maximum at 90°, and then falls again to zero, as the lower curve of Fig. 38 shows. In the next half-cycle, 180° to 360°, voltage and current are *both* negative ; their product is, therefore, still positive. A second rise and fall of wattage, exactly equal to that of the first half-cycle, will therefore occur.

In a resistive circuit, then, the power rises and falls once every half-cycle of the applied voltage. But it remains always positive, so that at every individual instant (except at 0° and 180°) power is being consumed in the circuit. R.M.S. voltage and current, and average power, are marked on the curves ; it will be seen that, as already explained, the calculation of average power from R.M.S. voltage and current, or from either of these and the value of resistance in the circuit, is worked out exactly as for direct current.

When current and voltage rise and fall exactly in step, as in the figure we have been discussing, the two are said to be *in phase*. It is evident that in any such case their product will remain positive at every instant. This relationship of

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current and voltage, therefore, cannot apply to purely reactive circuits (inductance or capacitance alone) in which the energy drawn to build up the field is completely restored when the field collapses. In such circuits it is evident that the two must be out of step.

42. Capacitance : I Leading by 90°

In Fig. 39 is repeated the full-line voltage-curve of Fig. 38, but this time there is associated with it a current curve displaced by 90°, or one-quarter of a cycle, towards the left. Since the diagram is read from left to right, this means that the current reaches its maximum a quarter of a cycle sooner than the voltage ; it is therefore said to *lead* the voltage by 90°, and is referred to as a *leading current.*

To calculate the power consumed with this new relationship between them we have, as before, to multiply corresponding pairs of values and plot the result. This leads to the lower full-line curve of this figure, from which it will be seen that the power is positive (i.e., absorbed) for the first quarter-cycle from 0° to 90°, is negative (i.e., refunded) for the next quarter-cycle from 90° to 180°, and so continues alternately positive

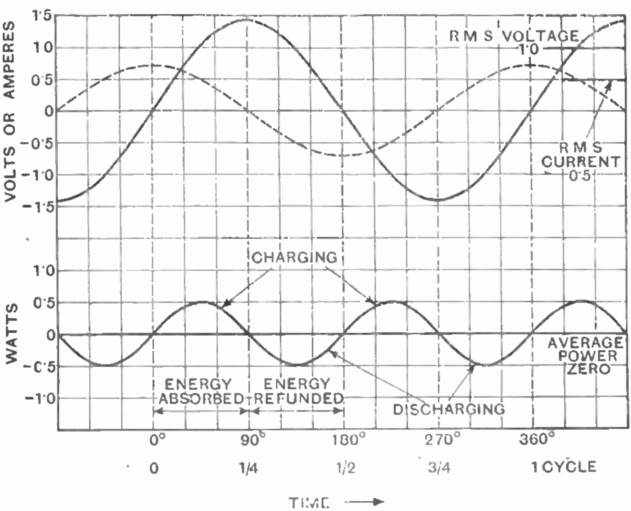

Fig. 39 : Relation of voltage, current and power in a purely capacitive circuit. Note that the average power is zero, and compare Figs. 38 and 40

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and negative. This would correspond satisfactorily with the conditions known to hold when an alternating voltage is applied to a condenser *or* a coil, energy being alternately stored in, and returned from, the electric or magnetic field.

That the curves shown actually represent the case of the condenser can be seen if we remember that at every instant the voltage across it is that of the upper full-line curve. At moments of full charge the voltage across the condenser is at its maximum, but the current is zero, for it is just on the point of changing direction. These instants occur at 90° and 270° in Fig. 39. Immediately after each stop the current is positive if the voltage is running up from negative to positive (270° to 90°) and negative if the voltage is running from positive to negative (90° to 270°). The curves thus show in detail the way in which an alternating voltage drives a current through a condenser.

A common difficulty is "*How*, if the E.M.F. causes the current, can the current lead the E.M.F.? How does the current know what the E.M.F. that produces it is going to be?" This question evidently assumes that if, for example, an alternating voltage began at the moment specified as 360° in Fig. 39 then the current would start at 270°. Actually both would start together, and it would only be after a number of cycles that they would get into the relative phases shown in Fig. 39. The current, then, is not really determined in some occult way by a voltage that is still in the future, but by the present and past voltage. The statement that the current leads the voltage by quarter of a cycle means that after a sufficient number of cycles for both to get into their stride, a current maximum occurs quarter of a cycle earlier than a voltage maximum, and must not be taken to imply that that current maximum is a direct result of the voltage maximum that has not yet taken place.

43. Inductance: I Lagging by 90°

If we displace the current-curve by 90° to the right instead of to the left of that representing the voltage, we arrive at the diagram of Fig. 40. Here, again, the power is alternately positive and negative, making, as before, an average of zero power over the complete cycle. These curves show the relationship between voltage and current that is found when the circuit consists of pure (i.e., resistanceless) inductance. We have already seen that the need for building up the magnetic field round the coil slows the growth of the current, while its collapse

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tends to maintain the current for an instant after the voltage driving it is removed. Examination of the curves of Fig. 40 show that they fulfil just these conditions, the current rising and falling always later than the voltage. At 180°, for example, the current is flowing in a positive direction even though the voltage has dropped to zero. The current is in the direction in which the voltage was urging it a quarter of a cycle earlier.

As in the case of Fig. 39, current and voltage are *out of*

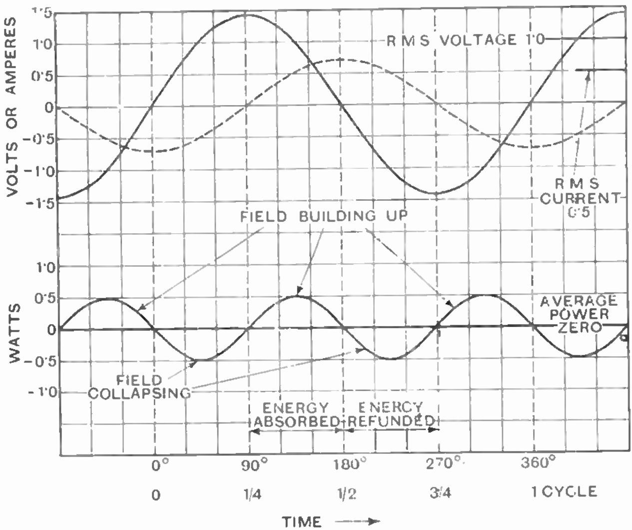

Fig. 40 : Relation of voltage, current and power in a purely inductive circuit. Note that the average power is zero, and compare Figs. 38 and 39

phase, there being a *phase-difference* of 90° between them. This is the necessary condition for a wattless current. In the present case the current is known as a *lagging* current, for the reason that it reaches each maximum after the voltage.

44. Resistance and Capacitance in Series

Suppose a resistance and a capacitance are connected in series, and an alternating E.M.F. is applied across the whole, as in Fig. 41. A current will flow, and, since this current consists of the physical movement of electrons, it will be the same at all parts of the circuit at any one instant. An E.M.F.,

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in phase with the current, will be required to drive the current through the resistance; if the peak value of the current is 0·25 amp., as shown dotted in Fig. 42, this E.M.F. must rise to a maximum of 1 volt, and is shown as a full-line curve marked Er. Similarly, the current must have an E.M.F. to pass it "through" the capacitance; this, however, will be 90° later in phase than the current, as shown by the full-line curve Ec. Its maximum of 2 volts, therefore, does not coincide in time with the maximum of the voltage across the resistance.

The total voltage across the two circuit elements, which is, of course, equal to the E.M.F. of the generator, must at every

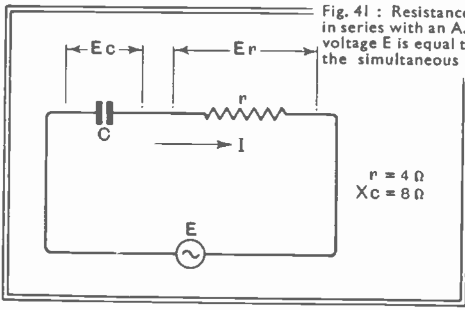

Fig. 41: Resistance and capacitive reactance in series with an A.C. source. The generator-voltage E is equal to the total voltage due to the simultaneous presence of Ec and Er. See Fig. 42

instant be equal to the sum of the two separate voltages (Sec. 13), and can be found by adding the heights of the two curves, point by point over the cycle. (The term "adding", it is to be noted, may mean "subtracting" in the sense that + 1 V and — ½ V add up to + ½ V by subtracting the negative half-volt from the positive volt.) The result of this addition is shown in the bottom curve of Fig. 42.

It will be noticed that this total voltage has a phase between those of the two component voltages from which we have built it up; it is some 63° out of phase with the current. Further, the maximum voltage is not the sum of the two separate peak voltages, because these do not occur at the same instant, but, as it rises to 2·24 V, it is larger than either alone.

We are now in possession of the information that an alternating voltage of 2·24 V drives a current of 0·25 amp. through a resistance of 4 ohms in series with a capacitive reactance of 8 ohms. The *impedance* of the circuit, which includes both resistance and reactance taken together, is defined by clinging to the outward form of Ohm's Law and saying that the impedance Z shall be equal to the voltage divided by the current; i.e., $I = E/Z$ instead of $I = E/R$, as in the simple case of direct current. In our present case $Z = E/I = 2{\cdot}24/0{\cdot}25 = 8{\cdot}94\ \Omega$.

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The two components of the impedance, 8 Ω reactance plus 4 Ω resistance, can obviously not be combined by simple addition to form this value of impedance, but it can be shown that a pure resistance r and a pure reactance X in series make up a total impedance Z which is given by $Z^2 = X^2 + r^2$. In

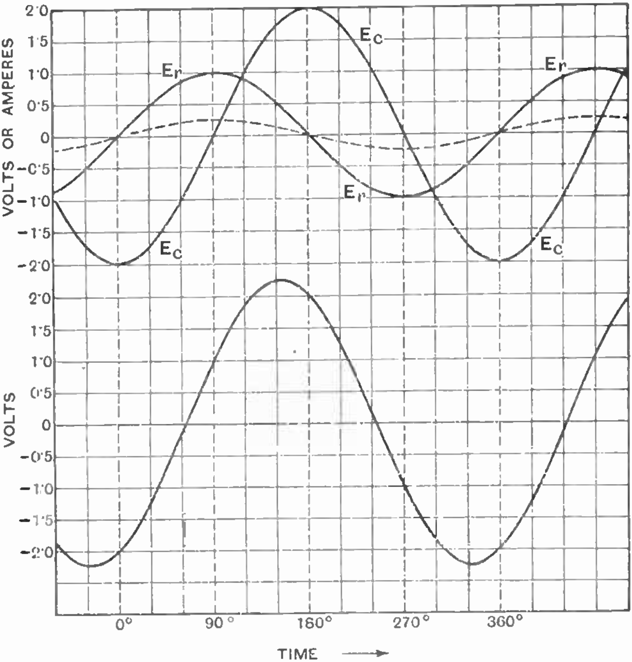

Fig. 42 : In the upper curve, Er represents voltage across the resistance, and Ec voltage across the condenser, of Fig. 41. The lower curve shows the resultant total voltage; that which the generator must have to drive 0·25 amp. through the circuit

our example, $Z^2 = 8^2 + 4^2 = 64 + 16 = 80$, whence $Z = 8{\cdot}94\ \Omega$, as already found. An impedance worked out in this way can always be used in the "Ohm's Law" formula $I = E/Z$ to find the current that will flow when a known voltage is applied.

45. Resistance and Inductance in Series

This combination is comparable with the last. In the circuit shown in Fig. 43 the generator E will drive some current I

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through the circuit. The voltage across r will be Ir, and this will be in phase with the current. Across the inductance the voltage will be IX, where X is the reactance, $2\pi fL$, of the coil at the frequency of the generator. This voltage will be 90° out of phase with the current, and hence also 90° out of phase with the voltage Ir. These two voltages, Ir and IX, must together be equal to E the voltage of the generator, but as their maxima do not occur at the same instant of time owing to their phase difference (see Fig. 42) their combined voltage is not the result of simply adding them together. Their phase-difference being exactly 90°, we can find E by combining them in the roundabout manner now beginning to become familiar: $E = \sqrt{(Ir)^2 + (IX)^2}$. This can also be written $E = I\sqrt{r^2 + X^2}$, showing that the impedance Z of this circuit is $\sqrt{r^2 \times X^2}$.

Comparing this with the case of the condenser, we find that

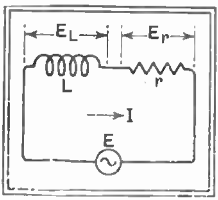

Fig. 43 : Resistance and inductive reactance in series with a source of A.C. voltage. Compare with Fig. 41 and the curve of Fig. 42

the same formula applies in both cases, since both are examples of the combination of voltages differing in phase by 90°. Problems involving an inductance in series with a resistance are therefore treated in exactly the same way as those depending on a resistance in series with a capacitance.

It is particularly to be noticed that inductance must always be associated with resistance in any practical case, even if the resistance is only the D.C. resistance of the wire with which the coil is wound. Although the two are in reality inextricably mixed up, it is satisfactory for purposes of calculation to regard any actual coil as a pure inductance in series with a pure resistance, as in the circuit of Fig. 43.

46. Resistance and Reactance in Parallel

If a capacitance (or an inductance) is connected, in parallel with a resistance, across a source of alternating voltage, each branch will draw its own current independently of the other. These currents will be E/r and E/X, where X is the reactance

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of the coil or condenser (Fig. 44). Since, like the voltages in Figs. 42 and 43, the two are not in phase, they cannot be added directly. That is, the magnitude of the total current is *not* equal to $E(1/r + 1/X)$. So long as the resistance is a pure resistance, and the reactance a pure reactance, so that the two currents are exactly 90° out of phase, the total current is given by $I = E\sqrt{(1/r)^2 + (1/X)^2}$. The impedance of r and X in parallel is, therefore, $\dfrac{1}{\sqrt{(1/r)^2 + (1/X)^2}}$ which may be simplified to $Z = \dfrac{rX}{\sqrt{r^2 + X^2}}$ (compare with the formula for resistances in parallel: Sec. 17).

We thus have:

Resistance and Reactance in Series: $Z = \sqrt{r^2 + X^2}$

Resistance and Reactance in Parallel: $Z = \dfrac{rX}{\sqrt{r^2 + X^2}}$

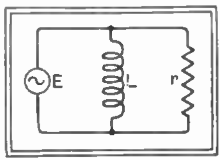

Fig. 44: Inductance and resistance in parallel. The combined impedance is given by
$1/Z = \sqrt{(1/r)^2 + (1/X)^2}$

It is particularly to be noted that this addition of squares only applies to the simple case where the two currents or voltages involved are exactly 90° out of phase; the further combination of one of these results with another reactance or resistance requires considerably more advanced methods than we propose to discuss here. The simple cases dealt with cover, fortunately, most ordinary wireless problems.

47. Power in A.C. Circuits

We have already seen that, in any alternating-current circuit, power is permanently consumed only when a current flows through a resistance. In this case voltage and current are in phase, and the power is equal to the product EI, both being expressed in R.M.S. units. In any purely reactive circuit current and voltage are 90° out of phase, and the power consumed, taken over a complete cycle, is zero. When both resistance and reactance are present together the phase difference lies between 90° and 0° as in Fig. 42, from which we conclude that the power consumed lies between zero and the product EI. It can be calculated by multiplying EI by a factor, always

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less than 1, that depends on the phase angle. But it is usually easier to find the current flowing through the circuit as a whole, and to multiply this by the voltage dropped across the resistive elements, ignoring entirely the voltage lost across the capacitance or inductance.

If, in Fig. 43, $X = 100\ \Omega$ and $r = 100\ \Omega$, the total impedance $Z = \sqrt{100^2 + 100^2} = 141{\cdot}4\ \Omega$. If $E = 200$ V, $I = E/Z = 1{\cdot}414$ amps. The voltage dropped across the coil is $IX = 141{\cdot}4$ V, but as it is known to be at 90° to the current, no power is consumed here. Across the resistance the voltage drop is $Ir = 141{\cdot}4$ V, implying the consumption of $I \times Ir$ or $I^2r = 200$ watts. This is the sole consumption of power in the circuit. The same method of determining the power can be applied to any complex circuit in which we may happen to be interested.

In brief, power in an A.C. circuit can always be reckoned from the formula I^2r, but the alternative formulæ EI and E^2/r can only be used on the strict understanding that E stands for the voltage on the resistance alone. Further, the symbol r means resistance only, and does *not* mean total impedance.

48. Transformers

In Sec. 30 we had a brief introduction to the transformer, which is very widely used in radio and electrical work generally. The main features of its action can be grasped by considering

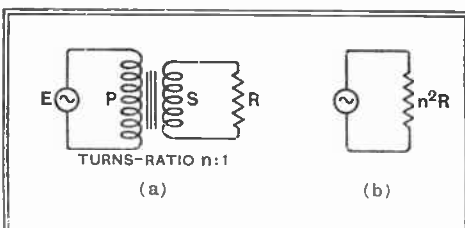

Fig. 45 : Iron-cored transformer drawing current from the generator E and delivering it, at a voltage equal to E/n, to the load R. From the point of view of loading the generator, diagram *b* represents an equivalent circuit (assuming a perfect transformer)

the simple arrangement shown in Fig. 45*a*. Such complications as resistance of the windings are conveniently assumed to be absent. An iron core is used, not only to multiply the magnetic lines of force but also to ensure that practically all of them link both primary and secondary windings. It will be seen later that transformers used for high frequency A.C. often have no iron core and only part of the field due to one winding links with the other ; in other words the coupling is

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loose. It is much simpler, to begin with, to assume that the coupling is 100 per cent.

Firstly, consider what happens when the resistance R is disconnected. There being no resistance in the primary circuit either, the generator E.M.F. E is opposed solely by the back voltage caused by the varying of a magnetic field. The situation is exactly that described in detail in Sec. 43. To build up the field requires a certain number of ampere-turns; with an iron core that number is far smaller than it would be without. So if the number of turns is large the current is small. It is called the *magnetising current*.

How about the secondary coil? As R is disconnected, the circuit is broken and no current can flow; but as we have assumed that the whole of the sinusoidally varying magnetic field associated with the primary links with the secondary, too, an E.M.F. is generated in the secondary. The E.M.F. generated by a given change of field linked with a coil is proportional to the number of turns. The E.M.F. generated in the primary must be equal to E. As the secondary has one *n*th as many, we get E/n volts across its terminals.

For example, suppose P is connected to a 200-volt supply. The magnetising current rises until sufficient alternating magnetic field is developed to generate 200 volts. Suppose 100 ampere-turns are necessary to do this. Then if P has 2000 turns, the magnetising current is $100/2000 = 0{\cdot}05$ amp., and is 90° out of phase with E. Suppose n is 50, so s has $2000/50 = 40$ turns; then the secondary voltage is $200/50 = 4$, which might be used to heat the valves in a radio set. In practice it is not uncommon to have several secondary windings delivering different voltages for different purposes, on the same iron core, all energised by a single primary.

Now suppose R to be connected, and its value to be 0·5 ohm. The secondary voltage, 4, causes to flow a current of 8 amps., which introduces a magnetising force of $8 \times 40 = 320$ ampere-turns. As the secondary current is flowing in a resistive circuit, the current and the resulting field must be in phase with the secondary and primary voltages, not 90° out of phase as is the magnetising current. In order to restore the status quo in the primary circuit it is necessary for the 320 ampere-turns due to the secondary to be cancelled out by an equal number from the primary, in phase opposition (180°) to the primary back-E.M.F. and therefore in phase with E. This is so because only the original 100 amp.-turns 90° out of phase can provide the correct back E.M.F. to balance E. We have therefore an additional primary current—called the *load current*,

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because it is due to connecting the load R—sufficient to give 320 ampere turns, i.e., 320/2000 = 0·16 amp.

The magnetising current, being 90° out of phase with E, represents no power loss. But the load current, 0·16 amp., being in phase, represents 200 × 0·16 = 32 watts. Note that this is equal to what we are drawing out from the secondary —4 volts 8 amps.—which is what we would expect, having assumed a perfect transformer with resistanceless coils. In practice the secondary voltage drops somewhat, according to the amount of current taken; and the difference between input and output watts equals the loss in the transformer itself, due to coil resistance and currents induced in the iron core.

As, disregarding transformer losses, the P and S wattages are equal, it is obvious that a $n : 1$ voltage step-down must be a $1 : n$ current step-up. (In the above example the ratio of S to P load currents is 8 : 0·16 or 50 : 1).

It is also easy to deduce from this that the secondary load resistance, R, draws as much current from the primary as a resistance equal to n^2R. (In the example $n^2R = 50^2 \times 0{\cdot}5 =$ 1250 ohms, through which 200 volts would drive 0·16 amp.). See Fig. 45*b*.

It follows that the turns ratio, $1 : n$, necessary to make a secondary load resistance R_s equivalent to a resistance R_p across the primary must be $\sqrt{R_s/R_p}$.

The total primary current, I, is of course the combination of the load current I_L and the magnetising current I_M, and we have seen in Sec. 46 how to add two currents 90° out of phase with one another: $I = \sqrt{I_L^2 + I_M^2}$. If, as commonly happens, I_M is not more than about one quarter of I_L, it can generally be neglected in the total.

If it is not necessary to insulate the secondary winding from the primary, there is no need for two separate coils. The winding having the smaller number of turns can be abolished, and the connections tapped across the same number of turns forming part of the other winding. This device is called an *auto-transformer*.

The foregoing example is typical of a small "mains" transformer. Although the same principles apply, the emphasis is rather different in A.F. and R.F. types, which will be considered later.

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CHAPTER 6

THE TUNED CIRCUIT

49. Inductance and Capacitance in Series

THE use of inductance and capacitance in combination is one of the outstanding features of any wireless circuit; it will therefore repay us to make a fairly close study of their behaviour. Fig. 46 *a* shows a coil and a condenser connected in series across an A.C. generator of voltage E. As in the case of Fig. 42 (C and *r* in series) we will begin with the current and work backwards to find the voltage necessary to drive it.

The dotted curve of Fig. 47 represents, by the usual sine-curve, the current flowing; we have allotted to it a value of 0·25 amp. peak. Taking the reactance of C as 4 Ω, the voltage across it will be 1 V peak, displaced in phase by 90° from the current. The rise and fall of this voltage with time is given by the full-line curve E_C. As required for a condenser, the current reaches its maximum a quarter of a cycle before the voltage.

The voltage across the coil, of reactance 8 Ω, will be 2 V peak, and its phase will be such that the current reaches each maximum a quarter of a cycle after the voltage. The full-line curve E_L fulfils these conditions.

It will be seen at once that the two voltages E_L and E_C are out of phase by 180°, which means that at every instant they are in opposition. If we find the sum of the two by adding the heights of the curves point by point and plotting the resulting figures we obtain for E (the generator voltage necessary to drive the assumed quarter-ampere through the circuit) the curve at the bottom of the diagram.

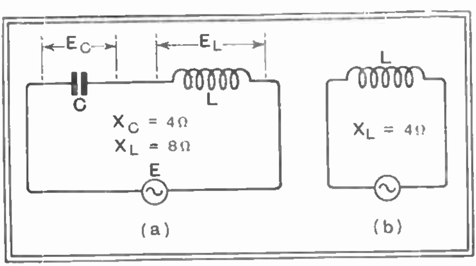

Fig. 46: Diagram *a* represents inductance and capacitance in series across an A.C. source, while diagram *b* shows the equivalent circuit for the values given. See Fig. 47

In obtaining this curve it was necessary to perform a subtraction at each point, since the two component voltages are at every instant in opposition. It is, therefore, scarcely surprising to find that the voltage required for the generator has the phase of the larger of the two voltages and is equal in magnitude to the difference of the two. The peak value of E is 1 volt, and its phase with respect to the current is that of the voltage across the inductance.

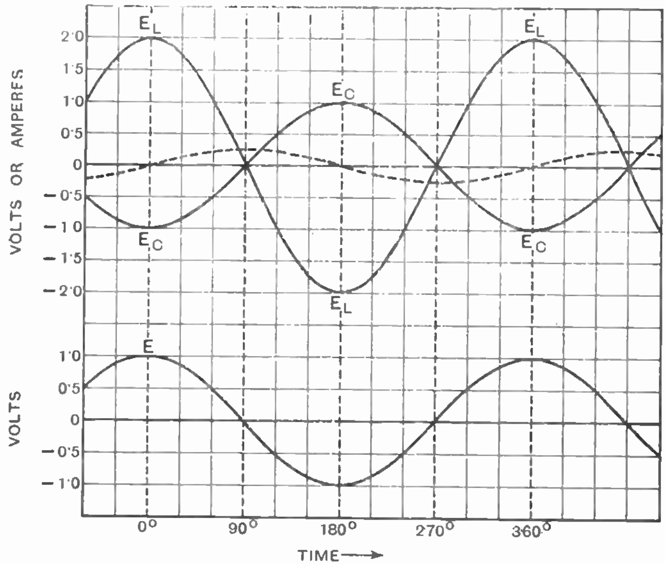

Fig. 47 : E_C and E_L represent the voltages across C and L of Fig. 46 when the current shown by the dotted curve is flowing. These voltages are at every instant in opposition and together make up to the voltage E

The same voltage and current relations that we find for the complete circuit could therefore equally well have been produced by applying 1 volt to a coil of reactance 4 Ω, as in Fig. 46 *b*. The capacitive reactance X_C of 4 Ω has exactly nullified 4 of the original 8 Ω of the inductive reactance X_L, leaving 4 Ω of inductive reactance still effective. We therefore conclude that the total reactance of a series combination of L and C is given by: $X = X_L - X_C$.

If we had made $X_L = 4\ \Omega$ and $X_C = 8\ \Omega$ in the original example, we should have found the circuit equivalent to a

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condenser of reactance 4 Ω. Applying the same rule, the total reactance would now be $(X_L - X_C) = (4 - 8) = -4\ \Omega$. The minus sign is conventionally taken to indicate that the combined reactance is capacitive.

50. L, C and *r* all in Series

Since the combination of a coil and a condenser in series is always equivalent either to a coil alone or to a condenser alone, it follows that the current through such a combination will always be 90° out of phase with the voltage across it. We can therefore combine the whole with a resistance in the same manner as any other reactance. To find the total impedance of the circuit of Fig. 48 *a*, for example, we have first to find the reactance X equivalent to X_L and X_C taken together;

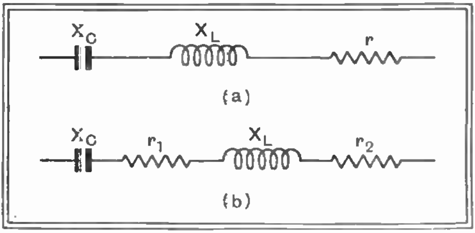

Fig. 48 : Capacitance, inductance and resistance in series. For *a*, $Z^2 = (X_L - X_C)^2 + r^2$. For *b*, $Z^2 = (X_L - X_C)^2 + (r_1 + r_2)^2$

$X = X_L - X_C$. To bring in the resistance we use the formula $Z = \sqrt{X^2 + r^2} = \sqrt{(X_L - X_C)^2 + r^2}$. There is no more complication here than in combining a resistance with a simple reactance.

Faced with a circuit like that of Fig. 48 *b*, we might feel inclined to begin by combining r_1 with X_C and r_2 with X_L, afterwards combining the two results. But a little consideration will show that neither of these pairs would be either a pure resistance or a pure reactance, so that we should have no immediate knowledge of the relative phases of the voltages across them. The final stage of the process would, therefore, be outside the range of the methods we have discussed. We get round the difficulty by first finding the total reactance of the circuit by adding X_L and X_C, then finding the total resistance by adding r_1 and r_2, and finally working out the impedance as for any other simple combination of reactance and resistance. The fact that neither the two reactances nor the two resistances are neighbours in the circuit does not have to be taken into

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consideration, since the same current flows through all in series.

51. The Series-Tuned Circuit

We have already seen that the reactance of a condenser falls (Sec. 35) and that of an inductance rises (Sec. 37) as the frequency of the current supplied to them is increased. It is therefore going to be interesting to study the behaviour of a circuit such as that of Fig. 49 over a range of frequencies. For the values given for the diagram, which are reasonably representative of practical broadcast reception, the reactances of coil and condenser for all frequencies up to 1,800 kilocycles per second are plotted as curves in Fig. 50. The most striking feature of this diagram is that at one particular frequency, about 800 kc/s, the coil and the condenser have equal reactances, each amounting then to about 1,000 ohms.

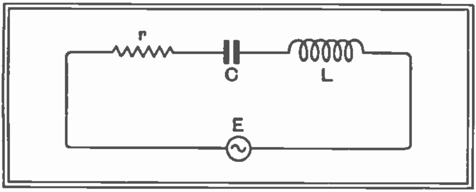

Fig. 49 : Series-tuned circuit : if L=200μH, C = 200$\mu\mu$F, r = 10 ohms, the magnification will be 100 at resonance. (See Fig. 50)

At this frequency the total reactance, being the difference of the two separate reactances, is zero. Alternatively expressed, the voltage developed across the one is equal to the voltage across the other ; and since they are, as always, in opposition, the two voltages cancel out exactly. The circuit of Fig. 49 would, therefore, be unaltered, so far as concerns its behaviour *as a whole* to a voltage of this particular frequency, by the complete removal from it of both L and C. This leaving only r, would result in the flow of a current equal to E/r.

Let us assume a voltage not unlikely in broadcast reception, and see what happens when E = 5 millivolts. The current at 800 kc/s will then be 5/10 = 0·5 milliamp., and this current will flow, not through r only, but through L and C as well. Each of these has a reactance of 1,000 ohms at this frequency ; the potential across each of them will therefore be 0·5 × 1,000 = 500 mV., which is just one hundred times the voltage E of the generator to which the flow of current is due.

That so small a voltage should give rise to two such large voltages elsewhere in the circuit is one of the queer paradoxes of alternating currents that make wireless possible. If the foregoing paragraphs have not made clear the possibility of

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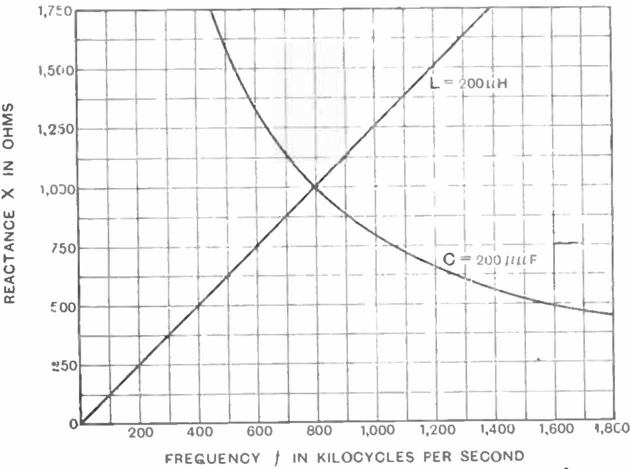

Fig. 50: Reactances of the coil and condenser of Fig. 49 plotted against frequency. Note that 800 kc/s, where the curves intersect, is the frequency of resonance

the apparent absurdity, the curves of Fig. 47, modified to make the two voltages equal, will give the complete picture of the large individual voltages in opposite phase.

52. Magnification

In the particular case we have discussed, the voltage across the coil (or across the condenser) is one hundred times that of the generator. This ratio is called the *magnification* of the circuit, and is generally denoted by the letter Q.

We have just worked out the Q for a particular circuit; now let us try to obtain a formula for the Q of any circuit. It is equal to the voltage across the coil divided by that from the generator, which is the same as the voltage across r. If I is the current flowing through both, then the voltage across L is $2\pi f$LI, or X_L. I; and that across r is Ir. So $Q = 2\pi f L/r$ or X_L/r. At any given frequency, magnification depends solely on L/r, the ratio of the inductance of the coil to the resistance of the circuit.

As we are considering the state of affairs when $X_C = X_L$, it is equally true that $Q = X_C/r$.

If r is made very small, the current round the circuit for the frequency for which the reactances of L and C are equal will be correspondingly large. In the theoretical case of zero

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resistance, the circuit would provide, at that one frequency, a complete short-circuit to the generator. Huge currents would flow, and the voltages on C and L would in consequence be enormous.

To obtain high magnification of a received signal (for which the generator of Fig. 49 stands), it is thus desirable to keep the resistance of the circuit as low as possible.

53. Resonance Curves

To voltages of frequencies other than that for which coil and condenser have equal reactance, the impedance of the circuit as a whole is not equal to r alone, but is increased by the net reactance. At 1,250 kc/s, for example, Fig. 50 shows that the individual reactances are 1,570 and 636 ohms respectively, leaving a total reactance of 934 ohms. Compared with this, the resistance is negligible, so that the current, for the same driving voltage of 5 mV., will be 5/934 mA, or, roughly, 5 microamps. This is approximately one hundredth of the current at 800 kc/s.

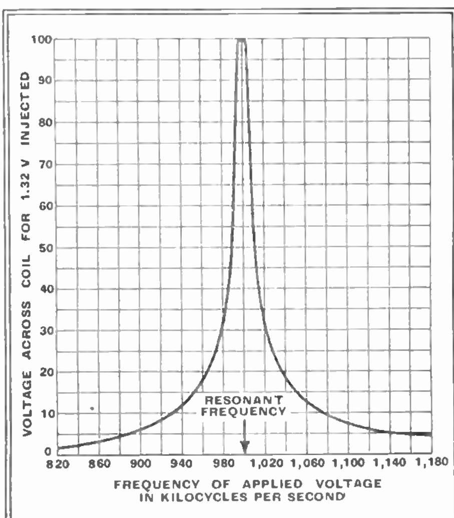

Fig. 51: Voltage plotted against frequency for a series circuit in which L= 180μH, C = 141$\mu\mu$F, r = 15 ohms, V = 1·32 volts

By extending this calculation to a number of different frequencies we could plot the current in the circuit, or the voltage developed across the coil, against frequency. The curve so obtained is called a *resonance curve* ; one is shown in

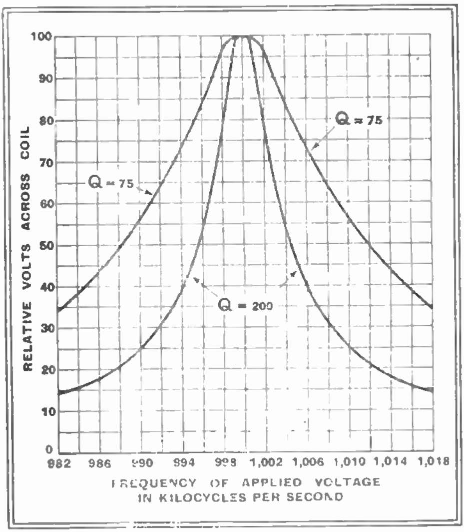

Fig. 52 : Resonance curves of two tuned circuits of different magnifications (Q) at 1,000 kc/s. The greater selectivity of the circuit of higher magnification is very apparent. Note that E (Fig. 49) is 1·32 volts for coil Q = 75, but only 0·5 v. for coil Q = 200

Fig. 51. The vertical scale shows the voltage developed across the coil for an injected voltage of 1·32 volts ; at 1,000 kc/s, the frequency at which $X_L = X_C$, the voltage across the coil rises to 100 volts, from which we conclude that Q = 100/1·32 = 75. Without going into details, a glance at the shape of the curve is enough to show that the response of the circuit is enormously greater to voltages at 1,000 kc/s than to voltages of substantially different frequencies ; the circuit is said to be *tuned* to, or to *resonate* to, 1,000 kc/s.

The principle on which a receiver is tuned is now beginning to be evident ; by adjusting the values of L or C in a circuit such as that under discussion it can be made to resonate to any desired frequency. Any signal voltages received from the aerial at that frequency will receive preferential amplification ;

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and the desired transmitter, distinguished from the rest by the frequency of the wave that it emits, will be heard to the comparative exclusion of the others.

54. Selectivity

We have said " comparative exclusion " because it is found that the *selectivity* of a single tuned circuit is seldom enough to provide sufficient separation between stations, so that two, three, or even more are used, all being tuned together by a single knob. The increase of selectivity obtained by multiplying circuits is very marked indeed ; with a single circuit of the constants of Fig. 51 a station is reduced to one-twentieth of its possible strength by tuning away from it by 120 kc/s (5 V response on Fig. 51 at f = 880 kc/s). Adding a second tuned circuit to select from the signals passed by the first leaves only one-twentieth of this twentieth—i.e., one four-hundredth. A third circuit leaves one-twentieth of this again—that is, one eight-thousandth. This last figure represents a set of about the minimum selectivity acceptable for general reception ; it follows that a receiver requires a minimum of three tuned circuits except in cases where means are provided for increasing the sharpness of tuning beyond that given by the unaided circuit.

The sharpness with which a circuit tunes depends entirely upon its magnification, as comparisons of the two curves of Fig. 52 will show. These are plotted to the same maximum height, thereby helping comparisons of selectivity while obscuring the fact that a circuit of Q = 200 gives a louder signal (more volts at resonance) than one for which Q = 75. In Fig. 53 the curves are redrawn to show the relative response of the two circuits to the same applied voltage ; the more selective circuit is also, as we have seen, the more responsive.

55. Resonant Frequency

At the frequency of resonance the reactance of the coil equals that of the condenser ; consequently we know that for that particular frequency $2\pi f L = 1/2\pi f C$. By a little re-arrangement of this equation, we get the important relationship $f = 1/2\pi\sqrt{LC}$, the resonant frequency f being in cycles per second, while L and C are in henries and farads respectively. This formula allows us to predict the frequency to which any chosen combination of inductance and capacitance will tune.

If we prefer our answer in terms of wavelength, we can replace f in the formula by its equivalent $\frac{3 \times 10^8}{\lambda}$, where λ

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is the wavelength in metres. This leads to the well-known formula $\lambda = 1{\cdot}885\sqrt{LC}$, where the figure 1·885 includes all numerical constants, and is made a convenient number by taking L in *micro*henries and C in *pico*farads.

It will be noticed that if a coil is tuned by a variable condenser (the customary method) it is necessary to quadruple the capacitance in order to double the wavelength or halve the

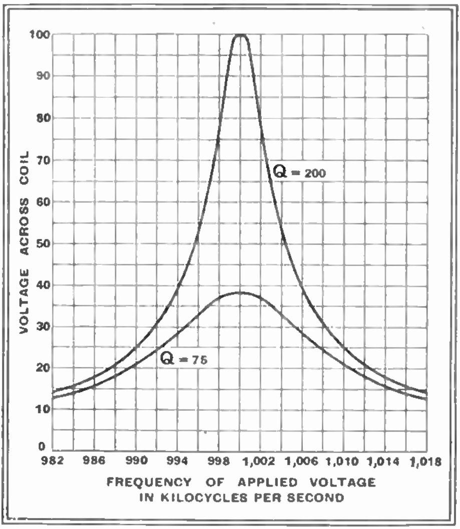

Fig. 53 : Voltages on two different coils for the same injected voltage (E = 0·5 v.). Note greater response, as well as higher selectivity (Fig. 52), of coil of higher Q

frequency. This is so because wavelength is proportional to the square root of the capacitance.

The average tuning condenser has a maximum capacitance of about 530 pF (or $\mu\mu F$); while the minimum capacitance, dependent more on the coil and the valves connected to it than upon the condenser, is generally about 70 pF. This gives a ratio of maximum to minimum capacitance of $530/70 = 7{\cdot}57$. The ratio of maximum to minimum frequency is the square root of this, namely, 2·75. Any band of frequencies with this range of maximum to minimum can be covered with one swing of the condenser, the exact values of the frequencies reached being dependent on the inductance chosen for the coil.

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Suppose we wished to tune from 1,500 kc/s to 1,500/2·75 or 545 kc/s, corresponding to the range of wavelengths 200 to 550 metres. For the highest frequency or lowest wavelength the capacitance will have its minimum value of 70 *p*F; by putting the appropriate values in either the formula for *f* or that for λ we find that L must be made 161 μH.* It should be evident that if we calculate the value of L necessary to give 545 kc/s (550 metres) with a capacitance of 530 *p*F, the same value will again be found.

By using instead a small inductance suitable for the short waves (0·402 μH.) we could cover the range 10 to 27·5 metres (30,000 to 10,900 kc/s, or 30 to 10·9 *mega*cycles per second), while the choice of 2120 μH, (or 2·12 *milli*henries) would enable us to tune from 728 to 2,000 metres.

Observe how convenience is served, large and clumsy numbers dodged, and errors in the placing of a decimal point made less likely by suitable choice of units, replacing "kilo-" by "mega-", or "micro-" by "milli-" whenever the figures suggest it. The preceding paragraph, rewritten in cycles and henries, would be almost impossible to read.

56. The Parallel-Tuned Circuit

The series-tuned circuit rather obviously derives its name from the fact that the voltage driving the current is in series with both coil and condenser, as in Fig. 54 *a*. In its very similar counterpart, the parallel-tuned circuit, the voltage is considered to be applied in parallel with both coil and condenser, as in Fig. 54 *b*. The change in circuit from one to the other results in a kind of interchange in the functions of current and voltage.

* Worked out thus. $f = 1/2\pi\sqrt{LC}$, so that $(2\pi f)^2C = 1/L$, or $L = 1/(2\pi f)^2C$. Now putting in values:

$$\begin{aligned} L &= 1/(2\pi \times 1500 \times 10^3)^2 \times 70 \times 10^{-12} \\ &= 1/88{\cdot}9 \times 10^{12} \times 70 \times 10^{-12} \\ &= 1/6210 = 0{\cdot}000161 \text{ henries} \\ &= 161 \text{ microhenries.} \end{aligned}$$

Starting from the formula $\lambda = 1{\cdot}885\sqrt{LC}$, we get $(\lambda/1{\cdot}885)^2 = LC$, or $L = \frac{1}{C}\left(\frac{\lambda}{1{\cdot}885}\right)^2$. Putting in values,

$$\begin{aligned} L &= \frac{1}{70}\left(\frac{200}{1{\cdot}885}\right)^2 \\ &= 106^2/70 \\ &= 11{,}250/70 \\ &= 161 \text{ microhenries, as before.} \end{aligned}$$

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In *a*, the current is necessarily the same at all parts of the circuit; we elucidated its behaviour by considering the voltages that this current would produce across the various components and added them up to find E, the driving voltage. In *b*, the

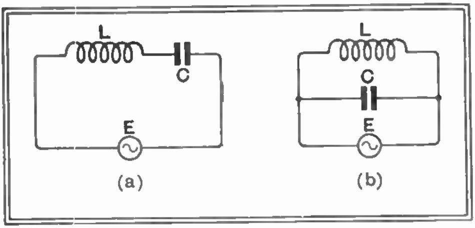

Fig. 54: Series- and parallel-tuned circuits compared. In the series circuit *a* the current through L and C is the same; in *b* the voltage across L and C is the same

position is reversed; here we have the same voltage applied to both the inductive and the capacitive branches, and we have to find the separate currents in the two and add them to find the total current.

With no resistance, the current in the L-branch will be determined by the reactance $2\pi f L$ of the coil; it will be $E/2\pi f L$. In the C-branch, it will similarly be $E/(1/2\pi f C) = E.2\pi f C$. We know already that these two currents will be exactly out of phase with one another, as were the voltages in Fig. 47. The net current taken from the generator will, therefore, be the simple difference of the two individual currents.

The currents become equal, and their difference consequently zero, at the frequency of resonance. As in the series circuit, this occurs when $2\pi f L$ equals $1/2\pi f C$, so that once again the frequency of resonance is given by $f = 1/2\pi\sqrt{LC}$. A coil and condenser of negligible resistance thus tune to the same frequency irrespective of whether they are arranged in series or parallel with the source of voltage that drives the current.

A parallel circuit has a resonance curve in all respects similar to that of the series circuit already discussed, the sharpness of tuning being determined, as before, by the magnification, X/r.

57. Series and Parallel Circuits Compared

In the series circuit, the current flowing produced across L and C two equal voltages, which, although they might individually be quite large, cancelled one another out. Taken together, L and C formed a part of a circuit across which no voltage was developed however large the current flowing; their joint impedance, therefore, was zero. If it were not for the presence of resistance, the series circuit would act as a

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short-circuit to currents of the frequency to which it is tuned; it is therefore often known as an "acceptor" circuit.

In the parallel resonant circuit the voltage E produces through L and C two equal currents, which, although they may individually be quite large, cancel one another out. Taken together, L and C give a circuit through which no current flows however large the voltage applied; their joint impedance, therefore, is infinitely large. The parallel circuit thus acts as a perfect barrier to currents of the frequency to which it is tuned; it is therefore often known as a "rejector" circuit.

58. The Effect of Resistance

But it will be clear that two conditions are necessary for this rejector action to be perfect. Firstly, the currents through the two branches must be equal, which can only happen when $X_L = X_C$; in other words, at the exact frequency of resonance. At other frequencies there will be more current through one branch than through the other.

The second condition for complete cancellation of the two currents is that they shall be out of phase by exactly 180 degrees. We have already seen that in a mixed circuit, containing both L and *r* or C and *r* the phase of the current lies between those appropriate to the individual circuit-elements in the way summarized in Fig. 55 and shown for one particular case in the curves of Fig. 42. It follows that if resistance is present in either the inductive or the capacitive branch of a parallel-tuned circuit,

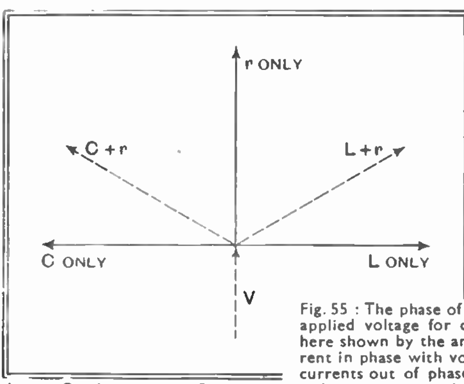

Fig. 55: The phase of the current relative to the applied voltage for different types of circuit is here shown by the angle of a line: *r* gives current in phase with voltage V, while C and L give currents out of phase by 90° in opposite directions. Combinations of C and *r* or L and *r*, give intermediate phases. Note that the currents with C only and L only tend to cancel out in a parallel circuit

the two currents are less than 180 degrees out of phase, and so can never exactly cancel one another. Even at resonance, therefore, there is a small residual current, with the result that the tuned circuit no longer presents a *complete* barrier to currents of the frequency to which it is tuned. Further, it will be clear that the larger the resistance r in series with L, the more the phase of the current passing through that branch of the circuit will depart from 90 degrees lagging, and so the larger will be the uncancelled residue of the capacitive current. Put briefly, a larger r leads to a larger current through the circuit as a whole, *and hence to a decrease in the total impedance of the circuit.*

59. Equivalent Parallel and Series Resistance

This curious result, that making the resistance of a part of a circuit larger reduces the impedance of the whole can be grasped more completely by studying Fig. 56. Circuits *a* and *b* differ only in *a* being series and *b* parallel and the value of R in *b* being as many times greater than X_L as X_L is greater than r in *a*. The actual values taken for example are as stated in the caption.

First of all we assume a sinusoidal current through the series circuit, say 1 amp. peak. A time curve of this, I is drawn in Fig. 56*c* for one and a half cycles. The voltages across L and r are easily worked out, and their curves drawn, in the same way as was done in Fig. 42. By adding the curves of E_L and Er we get the total voltage E required to drive the 1 amp. through the circuit. It is shown by a dotted curve, and it will be noted that the effect of r is to shift the phase of I relative to E (making it very obviously less than 90°) but hardly to alter the amplitude at all. We knew E_L is 10 volts, and now we see that E is practically the same, because the added 1 volt Er is 90° out of phase.

To calculate the parallel circuit we have to start off with the voltage E, and we shall make it the same as E in the series circuit, namely (as near as makes no matter) 10 volts. For comparison the curve of E in Fig. 56*d* is drawn in phase with E in the diagram above. The separate currents, I_L and Ir are calculated, and their curves drawn and added to give the total current, I. It can then be seen that *it is just the same in magnitude and phase as in the series circuit.* If therefore L and r or R were hidden in a box, with only the pair of terminals for connecting the generator accessible, it would be impossible to tell by measuring the number of amps. and the degrees phase lag on applying a voltage E at a fixed frequency whether the box contained 10 ohms inductive reactance in series with 1 ohm resistance or in parallel with 100 ohms resistance. Note *at a fixed frequency;* if the frequency were altered the reactance would alter and the two circuits would no longer be equivalent.

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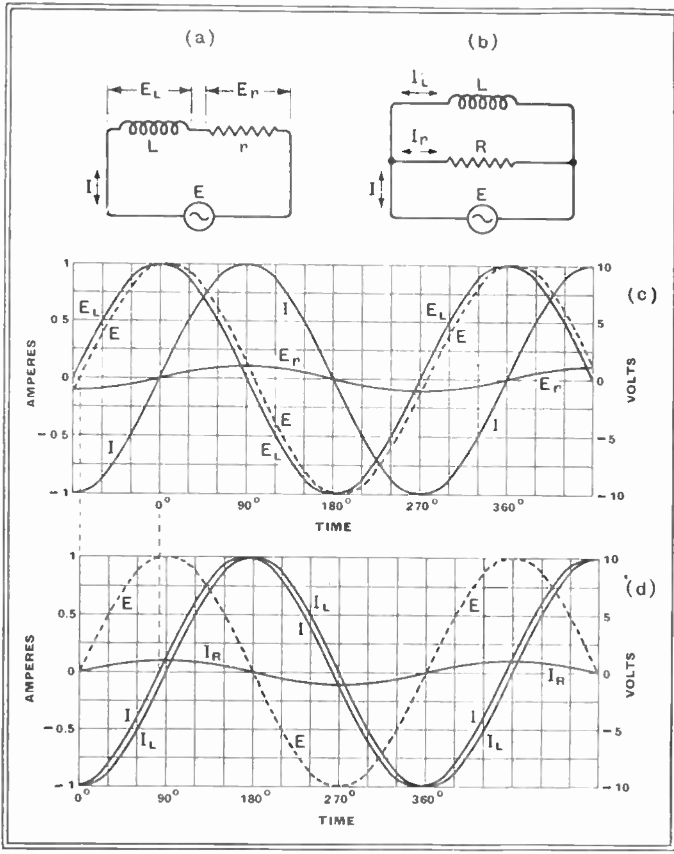

Fig. 56 : In the series circuit a, $X_L = 10\Omega$, $r = 1\Omega$, and I = 1 amp ; and the necessary voltage E is derived by the curves c. In the parallel circuit, *b*, $X_L = 10\Omega$, $R = 100\Omega$, and E = 10 volts ; and the total current I is derived by the curves *d*. Comparing c and *d* it can be seen that the series and parallel circuits are equivalent as wholes though not in their parts

It can be shown mathematically that, so long as r is small relative to X, then at any one frequency the parallel circuit *as a whole* is practically equivalent when $R/X = X/r$, or $R = X^2/r$. This applies to both inductive and capacitive reactances.

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60. Dynamic Resistance

In most tuned circuits r is not more than one tenth of X (in other words, Q is at least 10), so can be reckoned small enough for the series-parallel transfer explained in the preceding Section to be used. Take the tuned circuit shown in Fig. 57*a*, in which r represents the resistance of the tuning coil L. Performing the conversion we get *b* as the equivalent, in which $R = X^2/r$ or $(2\pi fL)^2/r$ or $1/(2\pi fC)^2 r$. We now have a resistance, R, in parallel with a *perfect* rejector circuit L and C, which passes no current at the resonant frequency, and can therefore

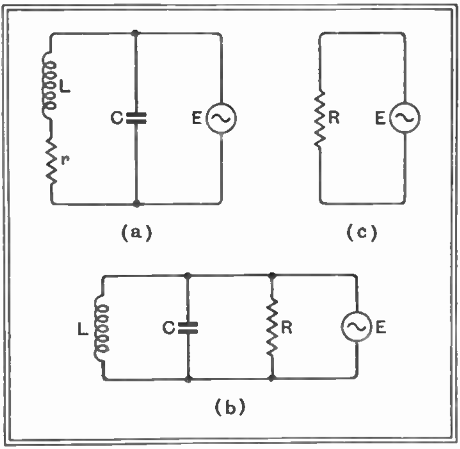

Fig. 57 a : Parallel-tuned circuit with resistance in coil. This is equivalent to *b*, which at the frequency of resonance can be simplified to c

be removed entirely, giving Fig. 57*c*. It is therefore permissible to replace the whole of a tuned circuit by a pure resistance R, it being strictly understood that this simplification is only allowable as long as we restrict ourselves to considering the behaviour of the circuit towards currents of the exact frequency to which it is tuned.

This resistance R, as we have seen, is infinitely large when r, the true resistance of the circuit, is zero, but decreases as r is increased. Since real, physical resistances do not behave in this topsy-turvy way, we have to distinguish R from an ordinary resistance by coining a special name for it; it is generally referred to as the *dynamic resistance* of the tuned circuit. As

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at the resonant frequency $2\pi fL = 1/2\pi fC$, it is possible to simplify the formula $R = (2\pi fL)^2/r$ to $R = L/Cr$. Thus a tuned circuit consisting of an inductance of 160 μH, tuned with a capacitance of 200 *p*F, and with a high-frequency resistance r of 7 ohms has a dynamic resistance $R = (160 \times 10^{-6})/(200 \times 10^{-12} \times 7) = (160 \times 10^{6})/1,400 = 114,000$ ohms. It is evident that if r had been $3\frac{1}{2}$ or 14 ohms R would have come out at 228,000 or 57,000 ohms respectively, so that halving or doubling the resistance r doubles or halves the dynamic resistance.

The relation between r and R is such that the specification of either, in conjunction with the values of L and C, completely determines the behaviour of the circuit at resonance.

If R is known, r can be found from the relation $r = L/CR$. Low values of parallel, or high values of series resistance *damp* the circuit, resulting in flat tuning.

61. Measuring R and r

In the sense that it cannot be measured by ordinary direct-current methods—by finding what current passes through it on connecting across it a 2-volt cell, for example—it is fair to

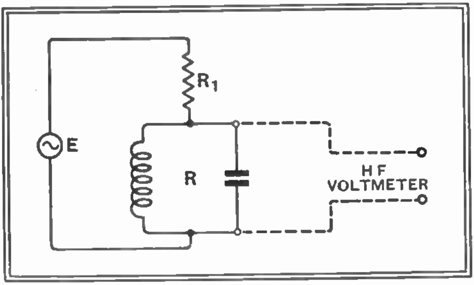

Fig. 58 : Simple method of determining dynamic resistance for a parallel-tuned circuit. If R_1 is adjusted until the R.F. voltmeter reads E/2 volts, we know that $R_1 = R$

describe R as a fictitious resistance. Yet it can quite readily be measured by such means as those outlined in Fig. 58, using for the measurement currents of the frequency to which the circuit is tuned. In spite of the inevitability of resistance in the windings of a coil, r is fictitious to just the same extent as R, for a true value of r cannot be obtained by any direct-current method. Indeed, it may often happen that a change in a coil that will reduce the resistance to direct current—by rewinding it with a thicker wire, for example—has the effect of increasing the high-frequency resistance instead of diminishing it. A true value for r can only be found by making the measurement at

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high frequency, using some such method as that outlined in Fig. 59.

It is possible to calculate the resistance offered to high-frequency currents by the wire with which a coil is wound. This value is always considerably higher than the plain resistance of the wire to ordinary direct current. Each turn of the coil lies in the magnetic field of the other turns, which has the result that there are set up stray currents in addition to the main current, thereby increasing the losses due to the resistance of the wire. Even in a straight wire the resistance at high frequency is greater than for steady currents, the magnetic field setting up the stray currents responsible for this being derived from the main current in the wire itself.

62. Dielectric Losses

By making a measurement of the high-frequency resistance of a tuned circuit on the lines indicated in Fig. 59 we always find a value for r which is appreciably higher than that found by calculation. This indicates that there are sources of

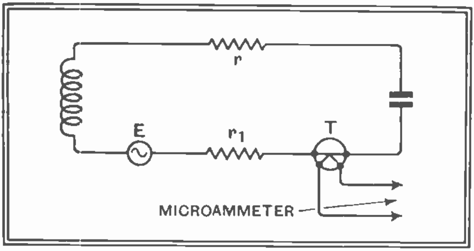

Fig. 59: Simple method of determining series resistance r. If the current is noted when $r_1 = 0$, and then reduced to half this value by introducing and adjusting r_1, the necessary value of r_1 is r. The resistance of the H.F. milliammeter (thermo-junction T) is included in r, and must be allowed for

resistance other than the wire with which the coil is wound. Investigation shows that this additional resistance, which may even be greater than that of the coil, is due to imperfections in the dielectric materials associated with the tuned circuit.

The plates of the tuning condenser, for example, have to be supported in some way; even if the dielectric between the plates is mainly air there is some capacitance between neighbouring portions of the two sets of plates for which the insulating support provides the dielectric. Valve-holders, valve-bases, or terminal blocks, connected across the tuned circuit also introduce capacitance, the dielectric again being the insulating material on which the metal parts are mounted.

To tune to very high frequencies (short waves) the capacitance has to be reduced to a very small amount, with the result that

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most of it may be contributed by such components, which must be selected with care for low-capacitance low-loss properties.

All these dielectrics are imperfect in the sense that they are not "perfect springs". In other words, in the rapid to-and-fro movement of electrons set up in them by the high-frequency voltage across the tuned circuit a certain amount of energy is absorbed and dissipated as heat. We have seen that the absorption of energy is characteristic of resistance; a circuit containing such sources of energy-loss as these is therefore found to have a high-frequency resistance r higher than that calculated for the coil and other metallic paths alone. The total is referred to as the "equivalent series resistance" of the circuit, the value of r so described being that which in conjunction with a perfectly loss-free condenser and coil would give a tuned circuit identical with the actual one at the frequency for which the measurements of resistance were made.

Physically, these dielectric losses behave as though they were a resistance in parallel with the circuit, as in Fig. 57*b*; but just as we converted a series resistance r into an equivalent parallel resistance, using the formula $R = L/Cr$, so we can convert back using $r = L/CR$. For example, suppose that a particular valve-holder is equivalent to 0·45 MΩ parallel resistance at 250 metres. If $L = 100\ \mu H$, then $C = 176\ pF$, and the value of r added to the circuit by connecting the valve-holder across it is $100/(176 \times 0{\cdot}45) = 1{\cdot}26\ \Omega$. But if the inductance of the coil were 200 μH, the added series resistance equivalent to the valve-holder would be 5·04 Ω, four times the preceding value (L doubled implies also C halved). As the true series resistance of the 200 μH coil (i.e., the resistance actually due to the winding itself) will be approximately double that of the 100 μH coil at the same frequency, it follows that the damping effect due to the valve-holder will be twice as great when the larger coil is in use.

A true series loss, such as a high-resistance connection in a switch or at a soldered joint, will add the same series resistance irrespective of the inductance of the coil. The lower the inductance of the coil, and hence the lower its resistance, the greater will be the damping effect of the added resistance. The distinction is important; a fixed series resistance damps a small coil more than a large one, whereas a fixed parallel resistance has a greater effect on a large coil.

The resistance of a coil, or of a tuned circuit, depends very largely upon the frequency. With a coil of some 160 μH, the equivalent series resistance may vary from some 25 ohms at 1500 kc/s to p haps 4 or 5 ohms at 550 kc/s. With decrease of frequency r drops; but C, the capacitance necessary to tune the coil to the required frequency, rises, with the result that the dynamic resistance does not vary so greatly as the figures for r would suggest. In practice, the values for R vary over a range of about two to one over the medium-wave band. The high values for series resistance at the highest frequencies are in the main due to dielectric losses, which, expressed as parallel resistance, are inversely proportional to frequency. A valve-holder that introduces 1½ M Ω parallel resistance at 500 kc/s will introduce ½ M Ω at 1,500 kc/s.

In conclusion, we see that the true representation of a tuned circuit as actually existing in a wireless set should include both series and parallel resistance, making a combination of Figs. 57 *a* and *b*. But, owing to the relationship existing between them, a circuit can be completely described at any one frequency by omitting either and making such an adjustment to the value of the other that it expresses the total loss of the circuit as a whole.

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CHAPTER 7

THE TRIODE VALVE

63. Free Electrons

IN discussing the nature of an electric current (Sec. 9) we saw that it consists of a stream of electrons along a conductor. The conductor is necessary in order to provide a supply of detached electrons, ready to be set in motion by an E.M.F. An E.M.F. applied to an insulator causes no current because the number of free electrons in it is negligible.

To control the current, it is necessary to move the conductor, as is done in a switch or rheostat. This is all right occasionally, but quite out of the question when (as in radio) we want to vary currents millions of times a second The difficulty is the weight of the conductor ; it is mechanically impracticable to move it at such a rate.

This difficulty can be overcome by freeing electrons (which are inconceivably light) from the relatively heavy metal or whatever conductor is used. This can be done by heating a suitable sort of material, called a cathode, causing electrons to " boil off " from it. To prevent the electronic current from being hindered by the surrounding air, the space in which we want it to flow is enclosed in a glass bulb and as much as possible of the air pumped out, giving a vacuum. We are now well on the way to manufacturing a *thermionic valve*, or, as it is called in America, a vacuum tube.

The electrons are too small and too light to feel much of the effects of gravity, and therefore do not tend to move in any particular direction unless urged by an electric field. In the absence of such a field they hover round the cathode, enclosing it in an electronic cloud known as the *space charge*. Because the charge consists of electrons, it is negative, and repels new arrivals back again to the cathode, so preventing any further accumulation of electrons in the vacuous space.

Cathodes are of two types—*directly heated* and *indirectly heated*. The first, more usually known as a *filament*, consists of a fine wire heated by passing a current through it, the electron-emitting surface consisting of a coating directly upon the filament wire itself. This type of valve is chiefly used in battery-driven sets. The indirectly heated cathode is a very narrow tube, usually of nickel, coated with the emitting material and heated by a separate filament, called the *heater*, threaded through it. Since the cathode is insulated from the

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heater, three connections are necessary compared with the two that suffice when the filament serves also as the source of electrons, unless two are joined together or "commoned" as in certain rectifier valves.

In all essentials the two types of cathode work in the same way; in dealing with valves we therefore propose to omit the heater or filament circuit altogether after the first few diagrams, indicating the cathode by a single connection. The operation of a valve depends upon the emission from the cathode; the means by which the cathode is heated to obtain this emission has little significance except in connection with the design of a complete receiver.

64. The Diode Valve

To overcome the stoppage caused by the negative space charge it is necessary to apply a positive potential. This is introduced into the valve by means of a metal plate called the *anode*. The simplest type of valve—the diode—contains only two *electrodes*—cathode and anode.

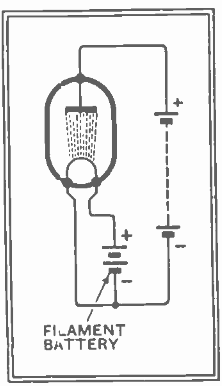

Fig 60: A directly heated (battery) diode valve

When the anode is made positive relative to the cathode, it attracts electrons from the space charge, causing its repulsion to diminish so that more electrons are emitted from the cathode, so that a current can flow, through the valve, round a circuit such as that of Fig. 60. But if the battery is reversed, so that the anode is more negative than the cathode, the electrons are repelled towards their source, and no current flows. The valve will therefore permit current to flow through it *in one direction only*, and it is from this property that its name is derived.

If the anode of a diode is slowly made more and more positive with respect to the cathode, as, for example, by moving upwards the slider of the potentiometer in Fig. 61, the attraction of the anode for the electrons is slowly augmented and the current increases. To each value of anode voltage E_a there corresponds some value of anode current I_a, and if the experiment is made and each pair of readings is recorded in the form of a dot on squared paper a curve like that of Fig. 62 is outlined.

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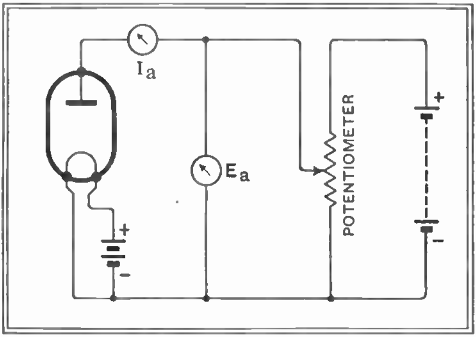

Fig. 61: Circuit for finding relation between anode voltage E_a and anode current I_a of a diode

The shape of the curve shows that the anode collects few electrons at low voltages, being unable to overcome the repelling effect of the space-charge. The greater the positive anode voltage the greater the negative space charge it is able to neutralise; that is to say, the greater the number of electrons that can be on their way between cathode and anode; in other words, the greater the anode current. By the time the point C is reached the voltage is so high that electrons are reaching the anode practically as fast as the cathode can emit them; a further rise in voltage collects only a few more strays, the current remaining almost constant from C to D and beyond. This is called *saturation*.

At B an anode voltage of 100 volts drives through the valve a current of 4 mA; it could therefore be replaced by a resistance of 100/0·004 = 25,000 ohms without altering the current flowing at this voltage. This value is therefore the equivalent D.C. resistance of the valve at this

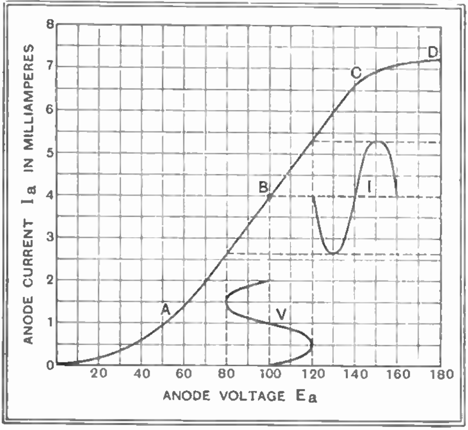

Fig. 62: Characteristic curve of a diode valve. The sine wave curves indicate the effect of superposing an alternating potential of 20 volts peak on the steady potential of 100 v. to which point B corresponds

point. Examination of the curve will show that the equivalent D.C. resistance of the valve *depends upon the voltage applied*; to drive 1 mA, for example, needs 53 volts, which leads to R = 53/0·001 = 53,000 ohms.

This is because the valve has no resistance in the sense in which we have hitherto used that term (Sec. 10). Although the so-called resistance of a valve—really the action of the space charge—is like true resistance in that it restricts the current that flows when a given voltage is applied, it does not cause the current to be exactly proportional to the voltage as it was in Fig. 8.

65. Anode A.C. Resistance (r_a)

One may, however, deduce the resistance of the valve in another way. Over the straight-line portion of the curve, round about B, an *increase* of 30 anode volts brings about an *increase* in anode current of 2 mA. The resistance over this region of the curve would therefore appear to be 30/0·002 = 15,000 ohms. This resistance is effective towards current-*variations* within the range A to C; if, for example, a steady anode voltage of 100 volts were applied (point B) and then an alternating voltage of peak value 20 volts were superposed on this, the resulting alternating current through the valve, as the curves on Fig. 62 show, would be 1·33 mA peak. Based on this, the resistance, as before, comes out to 20/1·33 = 15,000 ohms.* Thus the resistance derived in this way is that offered to an alternating voltage superposed on the steady anode voltage; it is therefore called the *Anode A.C. resistance* of the valve. Its importance in wireless technique is so great that it has had the special symbol r_a allotted to it. It is also, but not so correctly, called the "impedance" of the valve. The number of steps up in current for one step up in voltage is the *slope* of the curve. But r_a is the number of voltage steps for one current step, so it is 1 divided by the slope, or, as it is called, the reciprocal of the slope, at the particular point selected on it. As can be seen from Fig. 62, a steep slope means a low anode A.C. resistance.

The equivalent D.C. resistance of a valve is a quantity seldom used or mentioned; it was discussed here only for the sake of bringing into prominence the strictly A.C. meaning of the valve's impedance.

* By now the reader should have noticed that volts, *milli*amps, and *thousands of* ohms form a self-consistent system to which Ohm's Law applies. This offers a short cut in many wireless calculations.

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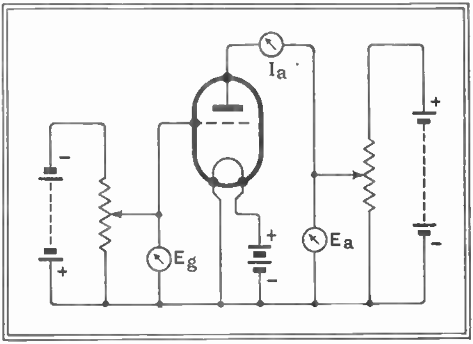
Fig. 63 : Circuit for taking characteristic curves, as in Fig. 64 or 65, of triode valves.



66. The Triode Valve

The diode valve has a very restricted field of use in that it can be used for rectification only ; it will not provide amplification. If a mesh of fine wire is inserted in the valve between cathode and anode in such a way that before they can get to the anode all the electrons emitted from the cathode have to pass through the meshes of this extra electrode a much fuller control of the electron-current becomes possible.

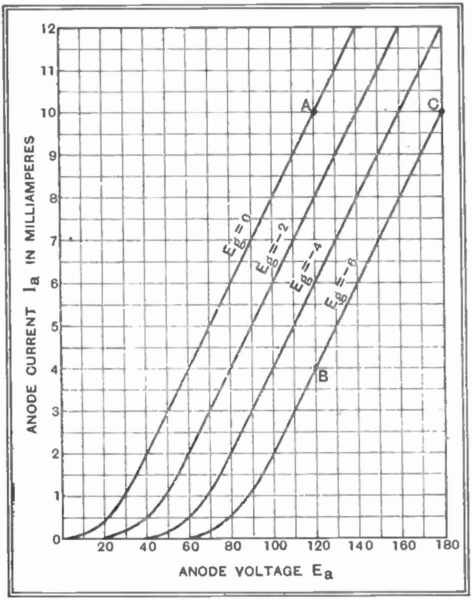


Fig. 64 : Characteristic curves of triode valve, each showing change of anode current with change of anode voltage

We are, in effect, altering the space-charge, which as we have seen is the only thing (short of saturation) affecting the amount of current that a given anode voltage causes to flow.

It is fairly evident that if this new electrode, the *grid*, is made

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positive it will assist the anode to neutralise the space charge, so increasing the anode current; and, being nearer to the cathode, one more grid volt will be more effective than one more anode volt. If, on the other hand, it is made negative it will assist the space charge in repelling electrons back towards the cathode. In Fig. 64 are shown four curves of a three-electrode valve, or *triode*, for comparison with the corresponding curve of the diode (Fig. 62). Each of these curves was taken with a fixed grid voltage which is indicated against each curve. It is to be noticed that this voltage, like all others connected with a valve, is reckoned *from the cathode as zero.* If, therefore, the cathode of a valve is made two volts positive with respect to earth, while the grid is connected back to earth, it is correct to describe the grid as "two volts negative," the words "with respect to the cathode" being understood. In a directly heated valve voltages are reckoned from the *negative end* of the filament.

67. Amplification Factor (μ)

Except for a successive shift to the right as the grid is made more negative, the curves in Fig. 64 are practically identical. This means that while a negative grid voltage reduces the anode current in the way described, this reduction can be counterbalanced by a suitable increase in anode voltage. In the valve of which curves are shown, an anode current of 10 mA can be produced by an anode voltage of 120 if the grid is held at zero potential ($E_g = 0$). This is indicated by the point A. If the grid is now made 6 volts negative the current drops to 4 mA (point B), but can be brought up again to its original value by increasing the anode voltage to 180 V (point C).

A change of 6 volts at the grid can thus be compensated for by a change of 60 volts, or ten times as much, at the anode. For reasons that will presently appear, this ratio of 10 to 1 is called the *amplification factor* of the valve, and is denoted by the Greek letter "μ" (mu).

As in the case of the diode, the anode resistance of the valve, by which is again meant the resistance it offers to a small alternating current when a small alternating voltage is superposed on some steady anode voltage, can be read off from the curves. All four curves of Fig. 64 will give the same value over their upper portions, since they all have the same inclination; over the lower parts, where the steepness varies from point to point, a whole range of values for the anode resistance exists. Over the straight-line portions of the curves this resistance is 10,000 ohms, as can be seen from the fact

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that the anode voltage must change by 10 to alter the anode current by 1 mA.

68. Mutual Conductance or Slope (g_m)

We have already seen that 1 volt on the grid is equivalent to 10 volts on the anode ; a change of 1 volt at the grid will, therefore, also cause a change in the anode current of 1 mA. This can also be read directly from the curves by observing that at $E_a = 100$, the anode current for $E_g = 0$ and $E_g = -2$ are 8 and 6 mA respectively, again a change of 1 mA for each one-volt change on the grid.

The response of the anode current of a valve to changes in voltage at the grid is the main index of the control that the grid exercises over the electron-stream through the valve. It is expressed in terms of *milliamperes* (of anode-current change) *per volt* (of change at the grid), and is called the *mutual conductance* (symbol g_m). It is related to μ and r_a by the simple equation $g_m/1000 = \mu/r_a$, or, if g_m is in *amperes* per volt, $g_m = \mu/r_a$; the derivation of which should be evident if the meanings of the symbols are considered. The magnitude of g_m is more clearly shown by valve-curves in which anode current is plotted, for a fixed anode voltage, against grid voltage. Some data from Fig. 64 are replotted in this form in Fig. 65, where the lines BC represent the anode-current change brought about by a change AB in grid voltage. The ratio BC/AB is very evidently the mutual conductance of the valve in milliamperes per volt. Since this ratio also defines the slope of the curve, it has become quite common to refer to g_m as the "*slope*" of the valve. But it must

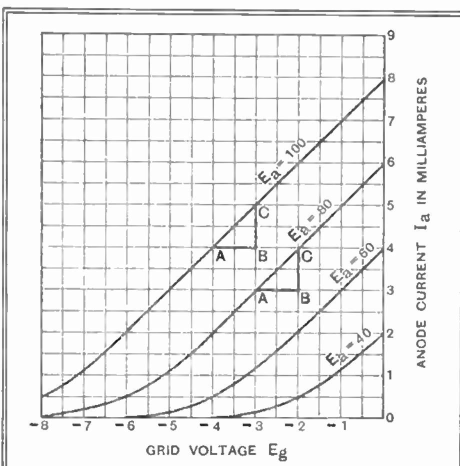

Fig. 65 : Changes of anode current corresponding to variations of grid voltage

be clearly understood that it is the slope of this particular valve characteristic (anode current against grid voltage) that is meant. The slope of the anode-current—anode-volts curve (Fig. 62) is obviously different, and, in fact, is the anode conductance $1/r_a$.

69. Alternating Voltage on the Grid

In Fig. 66 we have a set of E_g — I_a curves for a typical triode of the medium-impedance class. As the slope of the curve shows, its mutual conductance g_m is about $3\frac{1}{2}$ to 4 mA per volt for anode currents greater than about 4 mA, but less for lower currents. Suppose that, as suggested in the inset to that figure, we apply a small alternating voltage V_g to the grid of the valve, what will the anode current do? If the

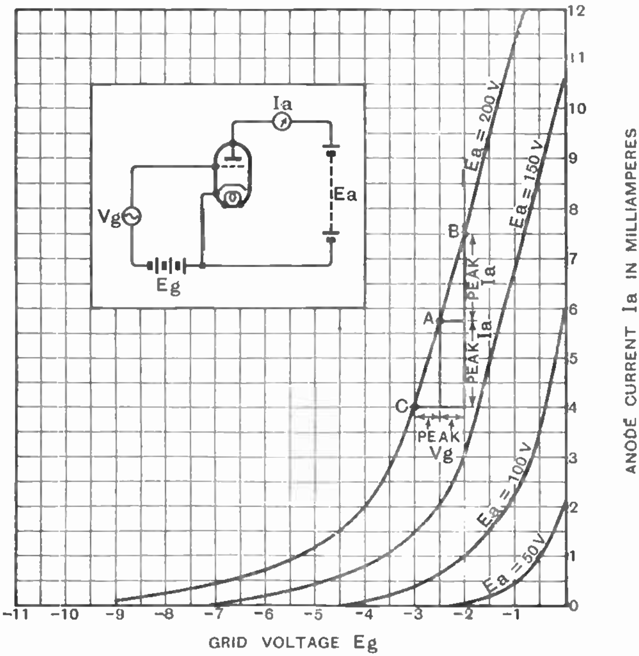

Fig. 66 : Anode-current grid-voltage curves of a medium-impedance indirectly heated triode. The alternating anode current caused by an alternating grid-voltage Vg can be read from the curves

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batteries supplying anode and grid give 200 and $-2\frac{1}{2}$ volts respectively, the anode current will set itself at about $5\frac{3}{4}$ mA—point A on the uppermost curve.

If the alternating voltage applied to the grid has a peak value of 0·5 volt, the total voltage on the grid will swing between -3 and -2 volts, alternate half-cycles adding to or subtracting from the initial (negative) grid voltage. The anode current will swing in sympathy with the changes in grid voltage, the points B and C marking the limits of the swing of both. The current, swinging between $7\frac{1}{2}$ and 4 mA, is reduced by $1\frac{3}{4}$ mA on the negative half-cycle and increased by the same amount on the positive one. The whole is therefore equivalent to the original steady current with an alternating current of $1\frac{3}{4}$ mA peak superposed on it.

70. The Triode as Amplifier

The development of an alternating anode current in response to the signal is not, however, enough. If the object is to operate another valve, it is necessary to convert the alternating current into alternating voltage before it can be effective at its grid. The way to do this is to pass the current through an impedance, of which the simplest sort to consider is a resistance (see inset to Fig. 67). On the other hand, if the alternating current from our valve is sufficient to operate a loud speaker or other device without further amplification, the loud speaker, etc., inevitably has some impedance, so in principle the circuit is the same.

We saw in Sec. 67 than a change of 1 volt at the grid is equivalent, in its effect on anode current, to a change of μ volts at the anode ; that is to say it produces a change in anode current equal to μ/r_a amp. But now the μ volts is applied to r_a in series with R, so the resulting anode current change is $\mu/(R+r_a)$. The change of voltage across R due to this is given by multiplying the two, giving $\mu R/(R+r_a)$ volts. This is the number of volts change across R resulting from 1 volt change of grid voltage, and is therefore the voltage amplification or *gain* of the whole *stage* (valve plus resistance, batteries, etc.). It is often denoted by the letter A. Examining the Fig. 67 curves in the region of A (for example) we see that to keep the current constant at 3·5 mA while the grid voltage is lowered by 1 volt (from -2 to -3) it is necessary to increase the anode voltage by 39 (from 151 to 190). So μ is 39. Keeping E_g steady at $-2\cdot5$ volts, raising I_a by 1 mA (from 3 mA to 4 mA) necessitates a rise in anode volts from 163 to 177·5 (14·5) so r_a is 14,500 Ω. According to the formula at which we have just arrived, then,

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the voltage amplification in this particular case = 39 × 20,000/(20,000 + 14,500) = 22·6.

71. The Load Line

We can tackle the same problem directly from the valve characteristic curves of Fig. 67. It is evident that when the valve is working with a resistance in its anode circuit the voltage

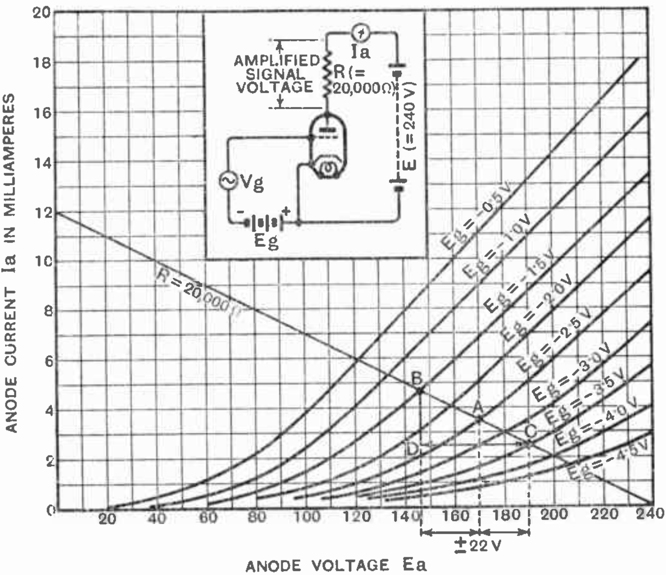

Fig. 67 : E_a—I_a, curves of the valve of Fig. 66. The line "R = 20,000 Ω" drawn across the curves gives, in conjunction with the curves themselves, full data as to the performance of the valve with this value of load resistance and an anode battery of 240 V.

across that resistance decreases with more negative grid voltage, and as the voltage of the anode battery or other source is supposed to be steady, the voltage at the anode must rise by an equal amount. This tends to offset the reduction in anode current that would be caused by the grid voltage change if there were no load resistance R.

The inset to Fig. 67 indicates that the battery supplies 240 V to the anode circuit as a whole. The steady anode current through the anode load resistance R will drop across this resistance some portion of the applied voltage ; at the

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anode itself the voltage will therefore be less than 240 V. For any given value of R we can plot voltage-at-anode against anode-current; if R = 20,000 Ω, there will be lost across it 20 volts for every milliamp. flowing, and the voltage at the anode will be reduced below 240 V by this amount. Thus the anode voltage will be 200 if I_a = 2 mA, 160 if I_a = 4 mA, and so on. Plotting these points gives us the line "R = 20,000 Ω" of Fig. 67.

The reason why it slopes in the opposite direction to the valve curves is that it represents the ratio of *loss* of anode voltage to increase in anode current.

From the way in which this line has been derived it is evident that every possible combination of I_a and E_a is expressed by some point along its length. Each of the valve curves across which it falls indicates the combination of E_a and I_a that are possible for the particular value of grid-bias indicated against that curve. It follows that if we set the bias at $-2\frac{1}{2}$ V with the load resistance in circuit and connected to 240 V as shown, I_a and E_a will be indicated by the point A, since the *working point* has to fulfil the double conditions of lying on both straight line and curve. For any other value of bias the anode current and voltage would equally take the values shown by the intersection of the *load-line* with the corresponding curve.

If the grid-voltage of the valve slowly increased from $-\frac{1}{2}$ V towards $-4\frac{1}{2}$ V the anode current will fall, as in Fig. 66, but the fall will be slower than in that figure since the anode voltage will rise as the current drops, as shown by the intersections of the load-line with the curves for successive values of bias.

In Fig. 67, the point A lies on the curve $E_g = -2\frac{1}{2}$ V. If we superpose on this a signal of 1 volt peak the grid will swing over a range of 2 volts, between $-3\frac{1}{2}$ and $-1\frac{1}{2}$. The limits of resulting anode current and voltage swing are indicated by the straight line BC, the anode voltage swing is 44 volts, a peak voltage of 22 for a grid input peak signal of 1 volt; so the stage gain is 22 times, which agrees quite well with the result calculated in the previous Section.

An important point that ought to be noticed is that *the signal voltage on the anode is opposite in polarity to that causing it at the grid.* When the grid voltage is made more negative the anode voltage becomes more positive, and vice versa.

72. The Effect of Load on Amplification

The formula for voltage amplification that we have used, $A = \mu R/(R + r_a)$, shows at once that by making R so large that r_a is negligible in comparison with it, the amplification

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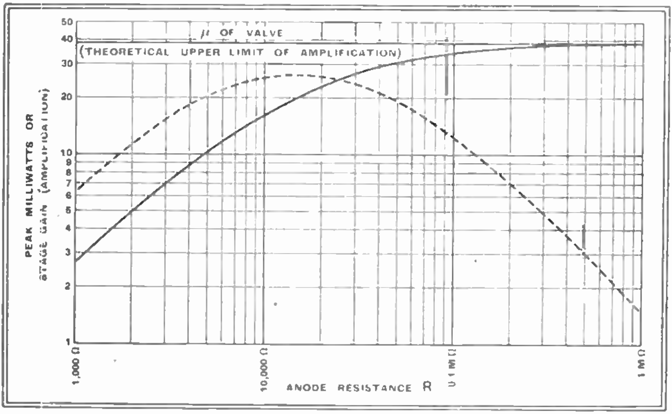

Fig. 68 : The full line shows how the voltage amplification depends on the value of load resistance R. It is calculated for a valve with $\mu = 39$ and $r_a = 14{,}500\,\Omega$. The dotted line shows the peak milliwatts delivered to R for a grid input of 1 volt peak. The maximum power is obtained when $R = r_a$

given by the stage will rise towards a theoretical maximum equal to μ, which supplies the reason for calling this quantity the " amplification factor " of the valve.

The same result can be had graphically by considering Fig. 67. For a higher value of R than 20,000 Ω, the line would be more nearly horizontal ; assuming the working point A retained, it would cut the axis $I_a = 0$ at a higher voltage. Imagining the line pivoted round A till it becomes horizontal (R infinitely high) we arrive at a diagram in which the anode current remains constant while the anode voltage changes, and so leads to μ as the stage-gain.

Conversely, the effect of a lower load can be studied by tipping the load-line towards the vertical ; evidently a lower stage-gain would be produced until, in the limit, the line becomes vertical ($R = 0$) and there is no change in voltage at the anode in response to the signal.

Fig. 68 shows the stage-gain obtained with various anode loads for a valve in which $r_a = 14{,}500\,\Omega$ and $\mu = 39$.* It

* *Logarithmic* scales are used for R and A in order to cover a wide range without compressing the most interesting part of the curve on the extreme left—a useful dodge that will be employed in other parts of the book. It is a scale in which equal distances represent equal *multiplications* instead of equal *additions*. For example, one-third of the whole R scale represents a ten-fold increase no matter where that third is taken.

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should be added that in order to keep these values constant when R is increased it is necessary to increase the voltage of the battery accordingly. If this were not done, if, for example, the battery in Fig. 67 were kept at 240 volts while R was raised to 100,000 Ω, the working point would be shifted from A to the foot of the curves where r_a would be much greater than 14,500 Ω and Fig. 68 would no longer apply. To keep the valve working at point A, where the anode current is 3·5 mA, would call for 350 volts to drive this current through the load resistor, in addition to the 170 volts needed for the valve itself. Any gain above about two-thirds μ, therefore, is obtained at an uneconomical cost in supply voltage.

It is obvious that a curve showing *current* output would rise towards the left-hand side. What is of more interest when the valve feeds some power-operated device such as a loud speaker is how the *power* depends on R. The dotted line in Fig. 68 shows the power in peak milliwatts when a signal of 1 volt peak is applied to the grid, all the conditions being as before. The curve falls at low R because the voltage falls (see full-line curve), and falls at high R because the current falls. It reaches a maximum when R is 14,500 Ω, i.e., is equal to r_a. This is an important general principle, of which more will be seen later in the book.

73. Power in Grid and Anode Circuits

The observant and enquiring reader may have wondered why the grid of the valve has been shown as always negative, and never positive, with respect to the cathode. The reason is bound up with the desire to expend as little power as possible in the grid circuit. If the grid were allowed to run positive it would act as an anode and collect electrons instead of making them all pass through its meshes, and a current would then flow round the grid circuit, absorbing power from the generator. Since this may be, in practice, a tuned circuit of high dynamic resistance, this absorption of power would have markedly ill effects in reducing the voltage across it and in decreasing the effective selectivity.

Provided no current flows in the grid circuit we may, for the moment, regard the valve as absorbing no power in that circuit. This condition will normally be fulfilled if the initial negative voltage, known as *grid bias*, applied by the battery, makes the grid negative enough to prevent the flow of grid current even at the peak of the positive half-cycle of signal voltage. In general, the bias required is equal to, or a volt or so greater than, the peak of the signal that the valve has to accept.

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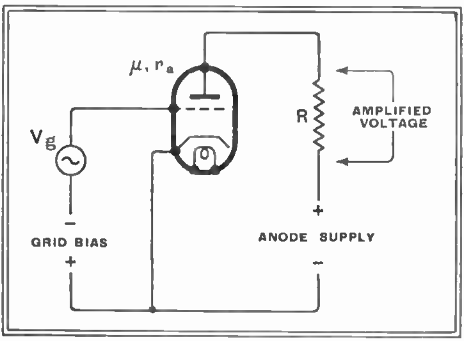
At very high frequencies, such as 30 Mc/s and above, other more subtle effects come into play, that are equivalent to placing a comparatively low resistance across the grid input points, causing it to absorb power.

In spite of the fact that at low and medium frequencies the power consumed in the grid circuit is negligibly small, alternating voltages applied to the grid can release an appreciable amount of A.C. power in the anode circuit. We have just discussed a case in which 1·1 mA peak of A.C. developed 22 volts across a resistance, making (22 × 1·1) = 24 peak or 12 R.M.S. milliwatts of power. This power is, of course, derived from the anode battery, which is continuously supplying 3·5 mA at a total of 240 volts, which is 840 milliwatts.

Behind all the curves and calculations there lies the simple basic fact that the valve is able to convert the D.C. power from the battery into A.C. power in response to a practically wattless A.C. driving-voltage on its grid. It is to this conversion, at bottom, that it owes its ability to amplify.



(a)

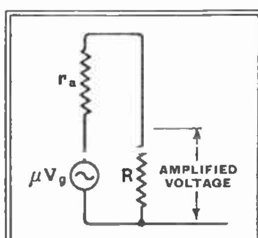
Fig. 69: (a) A stage of amplification, and (b) a diagrammatic representation of the anode circuit only, the signal voltage V_g on the grid being replaced by its equivalent μV_g volts, in series with the valve's own anode-cathode resistance r_a. Only Ohm's Law is needed to show that the voltage on R is $\frac{\mu R}{R + r_a}$ times V_g

74. Six Important Points

For reference, and as a summary of this chapter, we will enumerate the most important points about the triode valve.

(1) One volt at the grid controls the anode current to the same extent as μ volts at the anode; μ is the amplification factor of the valve.

(2) Towards A.C. the valve has an anode resistance r_a depending for its exact value upon the steady voltages applied.



(b)

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(3) The control of anode current by grid voltage is given by the ratio μ/r_a, known as the mutual conductance or slope, g_m.

(4) As a corollary to the above, it follows that a valve can be represented as a resistance r_a in series with a generator the voltage of which is μ times the alternating voltage applied to the grid. This representation (Fig. 69) takes no account whatever of steady voltages and currents, except through their influence in determining μ and r_a.

(5) The voltage amplification given by a valve in conjunction with a load resistance R is $A = \frac{\mu R}{R + r_a}$, as Fig. 69 clearly shows.

(6) Since r_a, μ, and g_m all depend, to a greater or lesser extent, on actual operating voltages, all *detailed* study of a valve's behaviour must be made by drawing load-lines across its actual curves, as in Fig. 67. Deductions such as Fig. 68 can then be made from these.

(7) The more elaborate valves which we shall discuss later are really only improved versions of the triode. In consequence, these six points cover about 90 per cent. of the philosophy of these more complicated structures.

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CHAPTER 8

OSCILLATION

75. Generating an Alternating Current

IN the last few chapters we have considered alternating currents in a variety of circuits, but have taken the generator of such currents for granted, representing it in diagrams by a conventional symbol. The only sort of A.C. generator actually shown (Fig. 33) is incapable of working at the very high frequencies necessary in radio, and produces a square waveform (Fig. 32) which is generally less desirable than the "ideal" sine shape (Fig. 29).

The alternating currents of power frequency—usually 50 c/s in Great Britain and many other countries—are generated by rotating machinery which moves conductors through magnetic fields produced by alternate N and S poles (Sec. 27). Such a method is quite impracticable for frequeneies of millions per second. This is where the curious properties of inductance and capacitance come to the rescue.

76. The Oscillatory Circuit

Fig. 70 shows a simple circuit consisting of inductance and capacitance in parallel, connected between the anode of a valve and its positive source. The valve is not essential in the first stage of the experiment, which could be performed with a battery and switch, but it will be needed later on. With the grid switch as shown, there is no negative bias, and we can assume that the anode current flowing through the coil L is fairly large. It is a steady current, and therefore any voltage drop across L is due to the resistance of the coil only. As none is shown, we assume that it is too small to take into account, and therefore there is no appreciable voltage across L and the condenser is completely uncharged. The current through L has set up a magnetic field which by now has reached a steady

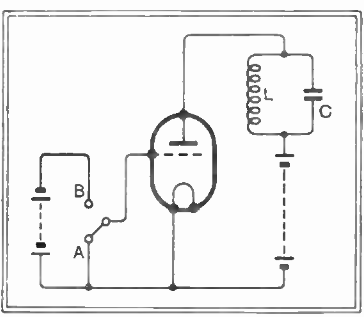

Fig. 70 : Circuit for demonstrating oscillation

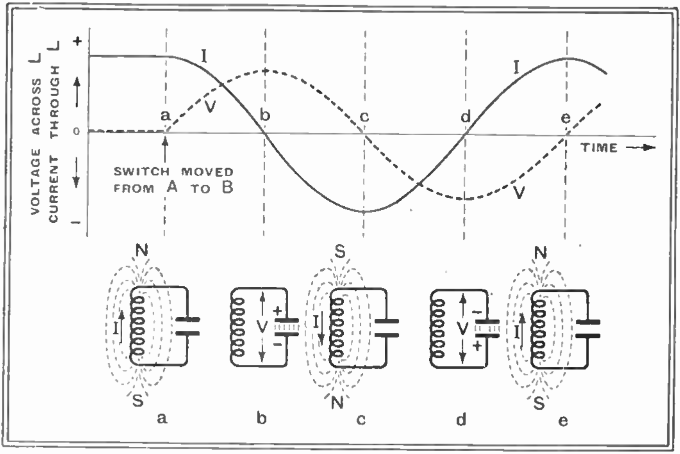

Fig. 71 : Sequence of events in Fig. 70 after the switch has been moved to B. *a* to *e* covers one complete cycle of oscillation

state, and represents a certain amount of energy stored up, depending on the strength of current and on the inductance L. This condition is represented in Fig. 71 by *a* so far as the circuit LC is concerned. The conventional direction of current (opposite to electron flow) is indicated.

Now suppose the switch is moved from A to B, putting such a large negative bias on the grid that anode current is completely cut off. If the current in L were cut off instantaneously, an infinitely high voltage would be self-induced across it (Secs. 28 and 29), but this is impossible owing to C. What actually happens is that directly the current through L starts to decrease, the collapse of magnetic field induces an E.M.F. in such a direction as to oppose the decrease, that is to say, it tends to make the current continue to flow. It can no longer flow through the valve, but there is nothing to stop it from flowing into the condenser C, charging it up. As it becomes charged, the voltage across it rises (Sec. 19), demanding an increased charging voltage from L. The only way in which the voltage can rise is for the current through it to fall at an increasing rate, as shown in Fig. 71 between *a* and *b*.

After a while the current is bound to drop to zero, by which time C has become charged to a certain voltage. This state is shown by the curves of I and V and by the circuit sketch

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at the point *b*. The magnetic field has completely disappeared, its energy having been transferred to C as an electric field.

The voltage across C must be balanced by an equal voltage across L, which can result only from the current through L continuing to fall at the same rate ; that means it must become negative, reversing its direction. We now have the condenser discharging through L. Directly it begins to do so it inevitably loses volts, just as a punctured tyre loses pressure. So the rate of change of current (which causes the voltage across L) gradually slackens, until, when C is completely discharged and there is zero voltage, the current is momentarily steady at a large negative value (point *c*). The whole of the energy stored in the electric field between the plates of C has been returned to L, but in the opposite polarity. Assuming no loss in the to-and-fro movement of current (such as would be caused by resistance in the circuit), it has now reached a negative value equal to the positive value with which it started.

This current now starts to charge C, and the whole process just described is repeated (but in the opposite polarity) through *d* to *e*. This second reversal of polarity has brought us to exactly the same condition as at the starting point *a*. And so the whole thing starts all over again and continues to do so indefinitely, the original store of energy drawn from the battery being forever exchanged between coil and condenser ; and the current will never cease oscillating.

It can be shown mathematically that voltage and current vary sinusoidally (Sec. 32).

77. Frequency of Oscillation

It is obvious that from the moment the valve current was switched off in Fig. 70 the current in L and the current in C (shown by the curve I in Fig. 71) had to be exactly equal at all times. Obviously, too, the voltage across them must always be equal. It follows (Secs. 36 and 37) that the reactances of L and C must be equal, that is to say $2\pi f L = 1/2\pi f C$. Rearranging this we get $f = 1/2\pi\sqrt{LC}$ which is the same formula as we have already had (Sec. 55) for the frequency of resonance. It means that if energy is once imparted to an oscillatory circuit (i.e., L and C joined together as shown) it goes on oscillating at a frequency called the natural frequency of oscillation, which depends entirely on the values of inductance and capacitance. By making these very small—using only a few small turns of wire and small widely-spaced plates—very high frequencies can be generated, running if necessary into hundreds of millions per second.

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78. Damping

The non-stop oscillator just described is as imaginary as perpetual motion. It is impossible to construct an oscillatory circuit entirely without resistance of any kind. Even if it were possible it would be merely a curiosity, without any practical value as a generator, because it would be unable to yield more than the limited amount of energy with which it was originally furnished. After that was exhausted it would stop.

The inevitable resistance of all circuits—using the term resistance in its widest sense : see Sec. 62—dissipates a proportion of the original energy during each cycle, so each is less than its predecessor, as in Fig. 72. This effect of resistance is

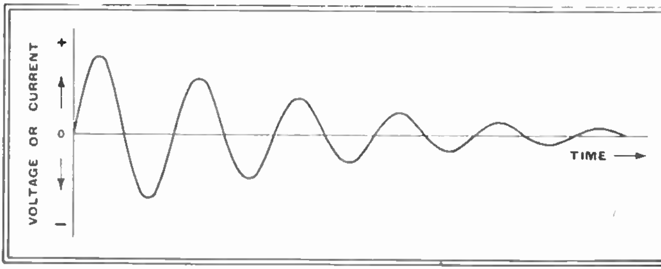

Fig. 72 : The first few cycles of a train of damped oscillations

called *damping* ; a highly damped oscillatory circuit is one in which the oscillations die out quickly. If the total effective series resistance r is equal to $2\sqrt{L/C}$ the circuit is said to be critically damped ; with less resistance the circuit is oscillatory, with more it is non-oscillatory and the voltage or current does not reverse after the initial kick due to an impulse but tends towards the shape shown in Figs. 18 and 27.

79. The Valve-Maintained Oscillator

What we want is a method of keeping an oscillator going by supplying it periodically with energy to make good what has been usefully drawn from it as well as what has unavoidably been lost in its own resistance. We have plenty of energy available from batteries, etc., in a continuous D.C. form. The problem is to apply it at the right moments.

The corresponding mechanical problem is solved in a clock. The pendulum is an oscillatory system that on its own soon comes to a standstill owing to frictional resistance. The driving energy of the mainspring is " D.C." and would be of no use

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if applied direct to the pendulum. The problem is solved by the escapement, which is a device for switching on the pressure from the spring once in each oscillation, the "switch" being controlled by the pendulum itself.

Turning back to Fig. 70 we have the oscillatory circuit LC which is set into oscillation by moving the switch from A to B at the stage marked *a* in Fig. 71. By the time marked *e* a certain amount of the energy will have been lost in resistance, but by operating the switch again from A to B the oscillation will be boosted, at the expense of the anode battery. Obviously the switch must be returned to A at some time between *a* and *e*; a suitable moment is at half-time, *c*, because the rush of

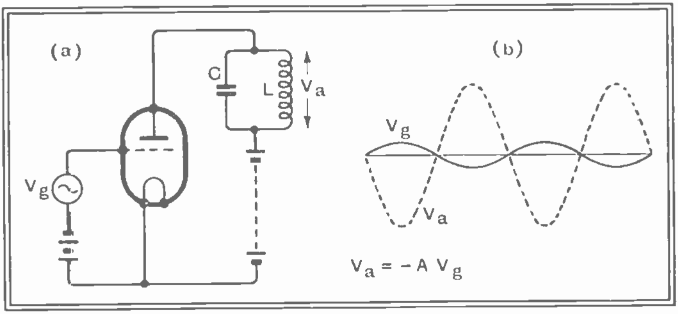

Fig. 73 : Diagram *b* shows the relative amplitude and phase of grid and anode voltages when their frequency is the same as the natural frequency of the circuit LC

current from the battery through the valve helps to reverse the negative current through L. If the energy thus imparted is more than has been lost in resistance, the oscillation grows in amplitude until the loss equals the gain.

What is needed, then, is something (an actual switch is not practicable for high frequencies) to make the grid more negative from *a* to *c* and more positive from *c* to *e*, and so on for each cycle.

At the frequency of oscillation the circuit LC is effectively a high resistance (Sec. 60). So we have all the essentials of a resistance-coupled amplifier (Sec. 70). As the valve amplifies, only a small part of the oscillating anode voltage, if applied to the grid, would release sufficient energy from the anode battery to cancel losses. From *a* to *c* the anode is more positive than its average, and from *c* to *e* more negative, which is just the opposite to what we have just seen to be necessary at the grid. We have already noticed (Sec. 71) that in a stage of valve

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amplification the anode voltage is opposite in polarity to the grid voltage causing it.

Suppose, as in Fig. 73*a*, a small generator of the required frequency is available to apply an alternating voltage V_g to the grid. LC being equivalent to a resistance, an amplified voltage V_a, A times as large, appears across it (Fig. 73*b*). If now a fraction 1/A of this output voltage is turned upside down and applied to the grid there is no need for the generator; the amplifier provides its own input and has become a generator of sustained oscillations, or, briefly, a valve oscillator.

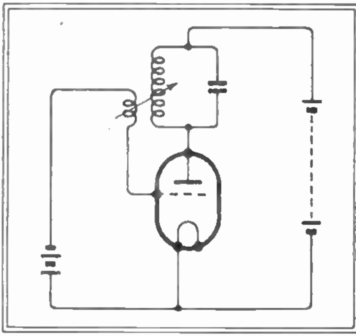

Fig. 74 : The reaction coil valve oscillator circuit

80. Valve Oscillator Circuits

The most obvious practical method of doing what has just been described is to connect a coil in the grid circuit and couple it magnetically to L until it has induced in it sufficient voltage to "oscillate" (Fig. 74). The arrow indicates a variable coupling, obtained by varying the distance between the coils. The grid coil—called a reaction (or retroaction) coil—can be connected either way round in the circuit; one of these ways gives the necessary reversal of sign. The reaction process is sometimes called positive feedback. When the coupling is sufficiently close and in the right direction for oscillation it is not necessary to take any trouble to start it going; the mere movement of electrons that constitutes any current is enough to start a tiny oscillation that very quickly builds up.

If the grid and anode are connected to opposite ends of the oscillatory coil and some point in between

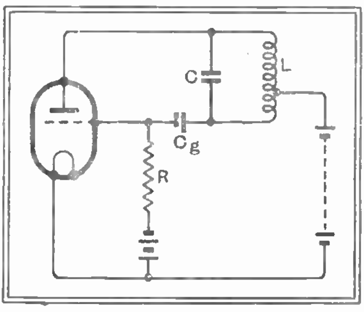

Fig. 75 : The series-fed Hartley circuit, showing parallel-fed grid bias using a grid leak R and condenser Cg.

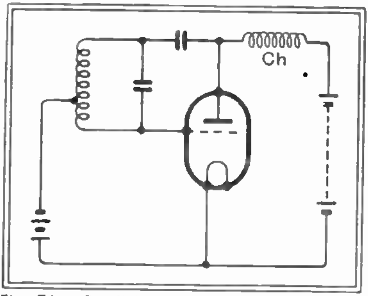

Fig. 76 : Parallel-fed Hartley circuit, with the anode current supplied through a choke Ch. The grid bias is series-fed

is connected through the anode battery to the cathode and is thereby kept at a fixed potential, the alternating grid voltage is always opposite to the alternating anode voltage, and can be adjusted in relative magnitude by choosing the tapping point on L (Fig. 75). Two extra components are required, because if the grid were connected straight to the coil it would receive the full positive voltage of the anode battery which would upset things badly (Sec. 73). C_g is used as a *blocking condenser* to stop this (Sec. 23), and R is used to connect a suitable grid bias. The capacitance of C_g is sufficient for its reactance to be little obstruction at the oscillatory frequency compared with the resistance of R, which is made large. This circuit, known as the Hartley, is a very effective oscillator, useful when the losses are so great that it might be difficult to obtain sufficiently close coupling to get Fig. 74 to oscillate. Fig. 75 shows the *series-fed* variety, called so because the anode feed current flows through the oscillatory circuit. The grid bias is parallel-fed through R. In the "parallel-fed Hartley" (Fig. 76) the anode is connected in this way, but to avoid loss of anode supply volts a choke coil Ch is generally used instead of a high resistance. The grid is shown series-fed, but it is more often parallel-fed in order to allow the coil to be connected to the cathode, which is usually earthed. Moreover, for reasons to be explained in Chapter 10, the combination of C_g, R, and the valve generates its own grid bias and the battery can be left out.

An alternative intermediate-potential point can be obtained on the oscillatory circuit by "tapping" the condenser instead of the coil, in the Colpitts

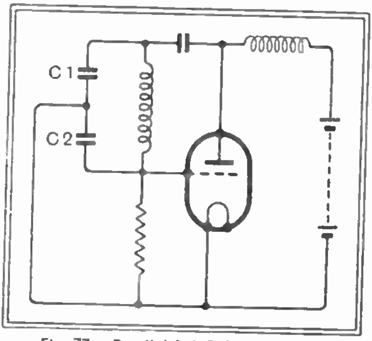

Fig. 77 : Parallel-fed Colpitts circuit

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circuit, of which one variety is shown in Fig. 77. In this the capacitance of the two condensers C_1C_2 together is made equal to that of the one condenser shown in the preceding circuits. The Colpitts is particularly favoured for very high frequency oscillators. C_2 then sometimes consists merely of the small condenser formed by the grid and cathode of the valve itself.

Another method of obtaining the necessary reversal between anode and grid is to take advantage of the fact, shown in Fig. 47, that when a current flows through a coil and condenser in series the voltages across them (neglecting resistance) are 180° out of phase (i.e., in opposite directions). This coil and condenser, L_1 and C_2 in Fig. 78, are proportioned so that the reactance of L_1, and hence the voltage across it, is much less than that of C_2, so that the voltage across it is less and in opposite phase to that across C_2 and L_1 combined, which voltage is the same as that across the oscillatory circuit (D.C. voltages are disregarded). In practice C_2 is usually the capacitance between the anode and grid themselves, while the inductive reactance of L_1 is adjusted by a condenser C_1 across it. The name of this circuit, "tuned-anode tuned-grid", then seems reasonable; but it should not be taken to mean that grid and anode circuits are both tuned to the frequency of oscillation. If the combination L_1 C_1 were so tuned, it would be equivalent to a resistance, and not an inductive reactance as required; while LC must also be slightly off tune so that it has sufficient reactance to shift the phase slightly to make up for the fact that the phase shift in $C_2L_1C_1$ is less than 180°, due to their resistance.

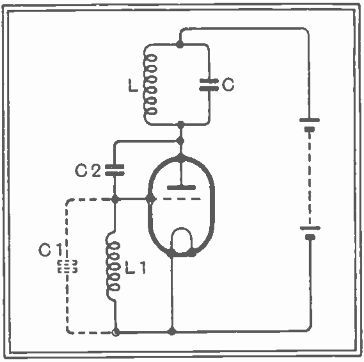

Fig. 78: Tuned-anode-tuned-grid (T.A.T.G.) oscillator circuit. The phase-reversing part of the circuit is C_2 (normally the valve itself) and L_1

The oscillator circuits shown in Figs. 74 to 78 look considerably different from one another at first glance, and can be modified to look even more different without affecting their basic similarity, which is that they are all valve amplifiers in

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which the grid " drive " or " excitation " is obtained by taking a part of the resulting anode alternating voltage and applying it in the opposite polarity, instead of having to go to some independent source for it.

81. Uses of Oscillators

In the outline of the essentials of any wireless system (Sec. 7) one of them was a source of high frequency oscillations, forming the main part of the *transmitter*. We shall consider that application of valve oscillators in the next chapter.

In Chapter 11 we shall approach very near the valve oscillator in considering simple types of *receivers*.

In Chapter 16 it is shown that a valve oscillator is a necessary part of the type of receiver in general use to-day.

In the radio service depot and the laboratory, valve oscillators are among the most important items of equipment, providing " signals " for testing receivers and other apparatus. It is easy to generate any frequency from less than 1 cycle per second up to about 200,000,000 by choosing suitable coils and condensers. Special types of valve and oscillatory circuit have been developed to oscillate up to 5,000 million c/s and even more.

82. Amplitude of Oscillation

It may be asked : what happens if the valve feeds back to its grid a larger amount of power than that being used up ? The oscillation will grow in amplitude ; the voltage at the grid will increase ; that will be amplified by the valve, increasing the output at the anode ; and so on. What stops it growing, like a snowball ?

Part of the answer lies in Sec. 73, in which it was pointed out that if the grid is allowed to become positive with respect to the cathode, it causes current to flow in that circuit and absorb power. When the amplitude of oscillation has grown to the point at which this loss of power due to grid current balances the surplus fed-back power, it will remain steady at that level.

If the grid bias is increased in an effort to obtain a greater amplitude of oscillation before grid current steps in to restrict it, another limiting influence is felt.

Look at Fig. 79 showing some typical valve characteristic curves of the same type as those in Fig. 67. Suppose the anode voltage is 300 and the grid bias -5 volts. Then any greater amplitude than 5 volts (approximately) causes grid current. Suppose also, for example, that the anode oscillatory circuit is equivalent at its resonant frequency to a resistance of 8,000 ohms

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(dynamic resistance) represented by the "load line" shown.* Then if the grid voltage has an amplitude of 5 volts, the anode voltage swings from 210 to 380 volts; i.e., it has an amplitude of 85 volts. If one-seventeenth of this (5 volts) is fed back to the grid in the correct phase, the conditions for oscillation are fulfilled. Any tendency for a slight overcoupling to the grid causes it to go positive, grid current flows, and by transformer

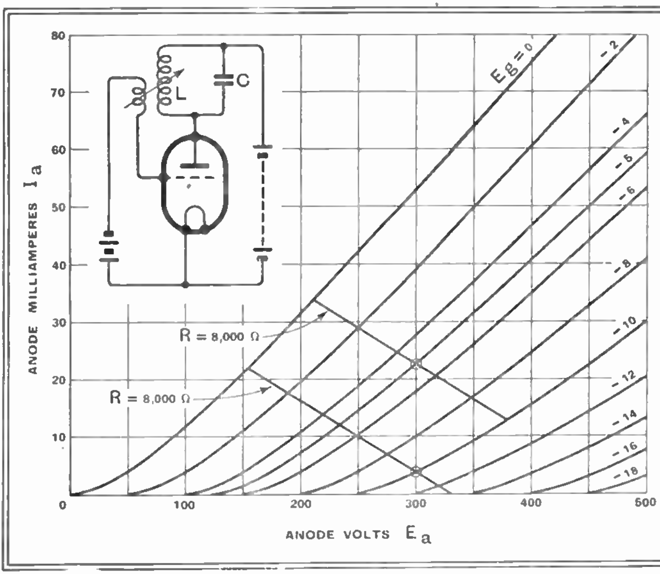

Fig. 79 : The oscillatory circuit LC is equivalent at resonance to a high resistance R and the voltage across it can be derived from a load line on the valve characteristic curves. From this follows the proportion to be fed back to the grid to sustain oscillation

action (Sec. 48) this is equivalent to a resistance load in parallel with R, so reducing the anode voltage swing and hence the voltage fed back to the grid. And so a brake is put on any tendency for the amplitude of oscillation to increase indefinitely.

Now suppose the grid bias is increased, say to −10 volts. The voltage amplitude at the grid can increase to about 10 volts before grid current flows. It is obvious from the lower load

* Note that as the 8,000 ohms applies only to A.C. of the resonant frequency, and not the D.C., the voltage actually reaching the anode is practically the same as that of the supply.

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line in Fig. 79, centred on $E_g = -10$, that the negative half-cycle of grid voltage cuts off the anode current altogether, so that it is distorted. The whole conception of dynamic resistance was based on a sinusoidal current (Sec. 59), and the simple method of calculation employed above breaks down. It should be fairly clear, however, that if during a substantial part of each cycle of oscillation the resistance of the valve, r, is abnormally high, or even infinity, due to the oscillatory voltage reaching to and beyond the " bottom bend " of the characteristic curves, the average amplification of the valve is reduced, and the tendency to oscillate also reduced. So here is another factor limiting the amplitude of oscillation.

83. Distortion of Oscillation

The previous section shows that if a valve oscillates fiercely the positive half-cycles of grid voltage cause grid current to flow, which imposes a heavy load on those parts of the cycle, and they are consequently clipped at the peaks. The corresponding anode current peaks are, therefore, similarly distorted. The negative half-cycles may cause anode current to be cut off, so they are clipped too. The result in either or both cases is a distorted wave, which for some purposes is undesirable. Obviously, to obtain as nearly as possible a pure sine wave from a valve oscillator it is desirable so to adjust the back-coupling that a very slight clipping is sufficient to stop the amplitude of oscillation from growing any more.

It should be clear from Fig. 79 that the greater the dynamic resistance, R, of the anode circuit (load line more horizontal), the greater the amplitude that is possible before either grid current or bottom-bend limiting set in, and the less, in proportion, is a given amount of clipping at the peaks.

We have seen (Sec. 60) that the dynamic resistance is equal to X^2/r, where X is the reactance of either coil or condenser at resonance ; while (Sec. 52) the magnification $Q = X/r$. Combining these we have $R = QX$. So to obtain a pure waveform the tuned circuit should have a high Q. This, incidentally, tends to increase the amplification, so that the amount of feedback needed to maintain oscillation is comparatively small.

There is a more important effect of high Q in minimising distortion. Sec. 58 showed how the smaller the series resistance, r, is in relation to the reactance X, the smaller is the external current fed to a parallel resonant circuit in relation to the current circulating round it. And we have seen that the ratio X/r is Q. The same thing can be seen in another way by considering the circuit, Fig. 70, with which we started to explain

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oscillation in Sec. 76. Here we started a sinusoidal oscillation by means of a square waveform produced by switching grid bias on and off. If the circuit has a high Q, the oscillatory current will build up to a value many times greater—to be precise, Q times greater—than the valve anode current of that frequency. The valve anode current therefore is such a small proportion of the whole that it does not matter very much whether it is distorted or not.

For example, suppose Q is 100, and R is 20,000 ohms. Then, as $R = QX$, X is 200 ohms. The impedance to alternating current of the resonant frequency from the valve is thus 100 times greater than the impedance of the coil or of the condenser. As the voltage across them is the same, the valve current is only 1 per cent. of the oscillatory current circulating internally between coil and condenser, and any distortion it may have hardly counts. The oscillatory current itself is always sinusoidal in normal circumstances.

To obtain as nearly as possible a pure sine waveform from an oscillator, then, use a high-Q circuit and the least back-coupling that will do.

84. Stability of Frequency

The frequency of an oscillator, as we have seen, is adjusted by varying the inductance and/or capacitance forming the oscillatory circuit. For most purposes it is desirable that when the adjustment has been made the frequency shall remain perfectly constant. This means that the inductance and capacitance (and, to a less extent, the resistance) shall remain constant. For this purpose inductance includes not only that which is intentionally provided by the tuning coil, but also the inductance of its leads and the effect of any coils or pieces of metal (equivalent to single short-circuited turns) inductively coupled to it. Similarly capacitance includes that of the coil, the wiring, valve electrodes, and valve and coil holders. Resistance is varied in many ways, such as a change in the supply voltages of the oscillator valve, causing r_g to alter.

Inductance and capacitance depend mainly on dimensions, which expand with rising temperature. If the tuning components are shut up in a box along with the oscillator valve, the temperature may rise considerably, and the frequency will drift for some time after switching on. Frequency stability therefore is aided by keeping the valve, and any resistors which develop heat, well away from the tuning components; by arranging effective ventilation; and by designing components so that expansion in one dimension offsets the effect of that

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in another. Fixed condensers are obtainable employing a special ceramic dielectric material whose variations with temperature oppose those of other condensers in the circuit ; this reminds one of temperature compensation in watches and clocks.

There are also other methods in which changes of frequency bring into effect mechanical or electrical compensating devices, analogous to the governor in an engine.

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CHAPTER 9

THE TRANSMITTER

85. Essentials of a Transmitter

THE last few chapters provide the material for tackling in greater detail what was indicated in barest outline in Fig. 4. Considering the transmitting system first, we have :

(1) A radio-frequency generator.
(2) A microphone, which may be regarded as a low-frequency generator.
(3) A modulator, combining the products of (1) and (2).
(4) An aerial, radiating this combination.

86. The R.F. Generator

In the valve oscillator we have an R.F. generator capable of working at any frequency up to many millions of cycles per second. It is easily controlled, both in frequency and output. For short-range communication, or even long ranges in special circumstances, ordinary receiving valves can be used. But to broadcast over a large area, or consistently over a long range, it is necessary to generate a large amount of power.

The difference between a receiving valve generating a few milliwatts (or at the most a few watts) of R.F. power, and a transmitting valve generating 100 kilowatts, is in the construction rather than the basic principles ; and as this book deals with the latter there is no need to go into the subject of transmitting valves as such. Obviously they are larger, especially as regards the filament, which has to supply the emission to pass a heavy anode current.

The supply voltage must be large, and batteries are uneconomical except in small portable sets. The usual methods of generating the D.C. power are described in Chapter 19.

87. The Need for High Efficiency

The higher the power the transmitter is required to produce, the more important it is that the percentage of power wasted shall be small. If, in generating 10 kilowatts of R.F. power, 30 kW are wasted, the total power to be supplied is 40 kW, and the efficiency (ratio of useful power to the total power employed) is 25 per cent. By increasing the efficiency to 75 per cent. it is possible to obtain 30 kW. of R.F. for the same expenditure, or, alternatively, to reduce the power employed

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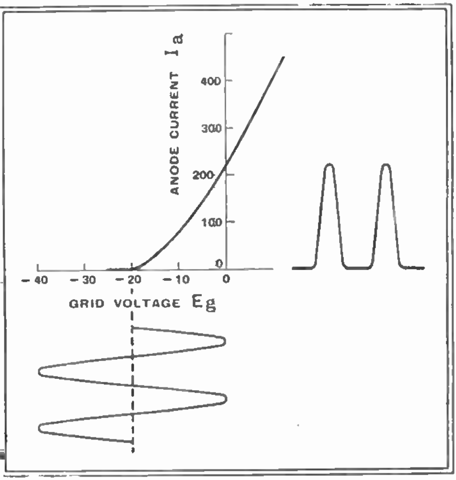
o 13·3 kW for the same output. In the latter case it would lmost certainly be possible to reduce the size and cost of the alves and other equipment as well.

Referring back to Fig. 79, suppose the upper load line, based n a grid bias of −5 volts and an anode voltage of 300, represents he swing of an oscillator, working under practically distortion-ess or so-called "Class A" conditions. The average anode oltage applied is 300, and the average current 23 mA, so the ower is 6·9 watts.

The useful output is obtained by multiplying R.M.S. alter-nating voltage by R.M.S. current. The swing is 170 volts ; herefore peak value 170/2 = 85 volts ; R.M.S. value 5/$\sqrt{2}$ = 60 volts. Current swing 22 mA ; R.M.S. value ·7 mA ; power 0·462 watts. This example points the way o a short cut for working out the A.C. power :

Watts = (Voltage swing × Current swing)/8.

The efficiency is therefore 0·462/6·9 = 0·067, or 6·7 per ent. This is very poor indeed, because under the conditions llustrated there is a considerable anode current flowing all the ime and a considerable voltage between anode and cathode of the valve. The power represented by this is dissipated as heat at the anode, and in the larger transmitting valves it is difficult to carry away all this heat. Many normally run bright red hot.



ig. 80 : Illustrating "Class B" operation with the aid of a alve anode current/grid voltage characteristic curve. The orking grid bias barely cuts off anode current, and only e positive half-cycles of grid signal are effective in reproducing themselves in the form of anode current

88. High-Efficiency Oscillators : "Class B"

Obviously the aim is to pass as little current through the valve as possible, consistent with keeping the oscillation going ; *and to pass that current only* *at times when the voltage across the valve is low.*

Actually we started off our study of oscillators on the right lines because in Fig. 71 the valve was totally cut off by a prohibitive grid bias during half the cycle, from *a* to *c*; and was allowed to pass current only during the half-cycle from *c* to *e* when the voltage across the tuned circuit was in opposition to that of the supply, and therefore the anode voltage low. That is just what we want, and is obtained by biasing the valve until it just fails to pass current (i.e., the working point is on the " bottom bend ").

These conditions are shown in Fig. 80, representing a valve with a negative bias of 20 volts and the feedback adjusted to give a peak grid voltage of the same amount, shown by the sine waves below the grid voltage scale. The resulting anode current has half of each cycle suppressed, shown by the waveform on the right; but, as we have seen, this current is normally very small in comparison with the sinusoidal oscillatory current.

89. " Class C "

The process can be carried farther, and the efficiency further improved, by increasing the negative bias to well beyond cut-off. The grid swing has to be increased too, and th

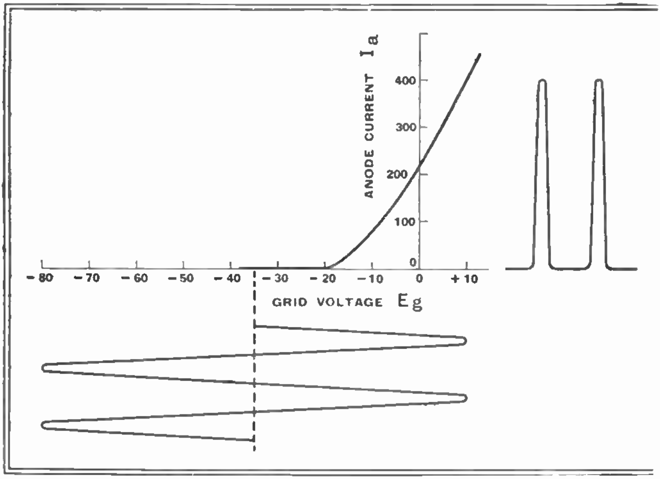


Fig. 81 : " Class C " operation. Comparing this with Fig. 80 it is seen that the worki grid bias (—35 volts in this example) is considerably more than enough to cut off ano current, which flows during only part of the positive half cycles of signal

conditions are referred to as "Class C" and are illustrated in Fig. 81. Note that the period during which anode current flows is less than half of each cycle, and is confined near the point *d* in Fig. 71, at which the anode voltage is a minimum, and therefore the power wasted in the valve is also very small.

Note too that it is advantageous to drive the grid positive, because, although there is a certain loss due to grid current, it occurs during only a small portion of each cycle, whereas the power developed in the anode circuit is much increased. Normally the valve is operated so that the positive grid voltage is nearly equal to the minimum anode voltage. Working efficiencies of the order of 75 per cent. are practicable, but beyond a certain point the power begins to drop off steeply. Moreover there may be difficulty in providing the very large grid swing.

Under Class C conditions the oscillator has either to be given a start or else begun with reduced bias, because full working bias cuts off anode current and prevents feedback. If worked as a self-oscillator (instead of merely a driven amplifier) the latter method is used. Sec. 209 will show a method of generating grid bias proportional to the amplitude of oscillation; an ideal arrangement for Class C, because it ensures the most powerful starting conditions.

90. Constant Frequency

The radio transmitter is one of the applications of the oscillator to which the requirement, noted in Sec. 84, that the frequency shall remain constant, applies with particular force. To avoid confusion it is internationally agreed that transmitter frequencies must keep within very narrow limits; and for special purposes— such as working more

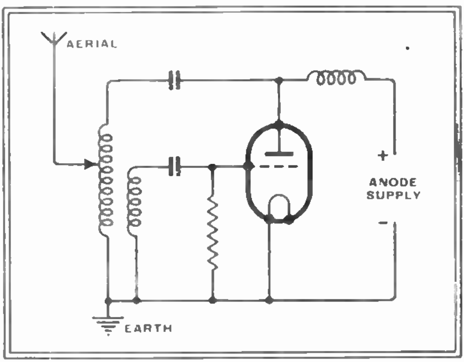


Fig. 82 : Direct coupling of aerial to the tuned circuit of a self-oscillating transmitter. The other end is usually connected to earth. In this arrangement, any changes in the aerial alter the frequency of oscillation

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than one transmitter on the same frequency—it is now common practice to keep within less than one part in a million.

In early days, the tuned oscillatory circuit (sometimes called a tank circuit, because of its capability for storing oscillatory energy) was coupled straight to the aerial, either inductively or by a direct connection or " tapping ", as in Fig. 82. Aerials will be discussed in detail in Chapter 20 ; in the meantime they can be considered as opened-out condensers (or, sometimes, coils) necessary for effectively radiating waves over distances. They have both capacitance and inductance, which depend on their height above ground and other objects ; so on a windy day are likely to vary erratically. Now, the frequency of an oscillator arranged as in Fig. 82 depends on the inductance and capacitance of the tuned circuit, of which the aerial forms part. The frequency stability of such a system is not nearly good enough for present-day needs.

The policy is to confine any circuits or components that can affect the frequency to rigid and compact structures—compact, so that their electric or magnetic fields do not spread out where persons or things might move and upset the field distribution and hence the capacitance or inductance. We shall see (Sec. 136) that it is possible to prevent such variations by shutting up the components in a metal box or *screen*.

Another weakness of the simple Fig. 82 system is that the conditions for getting most R.F. power out of a valve are just about the opposite of those for frequency stability.

91. The Master Oscillator Power Amplifier System

The way out of this dilemma is to separate the function of oscillating at the required frequency from that of feeding R.F power to the aerial, instead of trying to make one valve do both The oscillation frequency is generated by a valve oscillator in which all practicable precautions are taken to avoid frequency drift, and in which no attempt is made to obtain an extremely high power efficiency. This *master oscillator* is coupled, by some means that does not seriously affect the frequency stability, to one or more stages of amplification, the last of which is coupled to the aerial. Being designed to feed as much power to it as possible, this stage is called the *power amplifier* The whole is therefore called a master-oscillator power-amplifier system ; for short, M.O.P.A.

92. Crystal Control

To obtain extreme frequency stability even in a M.O.P.A requires very good design and considerable expense. Where i

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is not necessary to cover a range of frequency, but is enough to be able to work on a number of "spot" frequencies, an easy way of guaranteeing correct frequency to within very narrow limits quite simply and cheaply is to exploit the remarkable properties of certain crystals, notably quartz. Such crystals are capable of vibrating mechanically at very high frequencies and in doing so develop an alternating E.M.F. between two opposite faces. Conversely, an alternating E.M.F. causes them to vibrate.

The subject is a large and specialised one, but for most purposes the crystal can be regarded as equivalent to a "tank" circuit of remarkably high inductance and Q. But unlike the

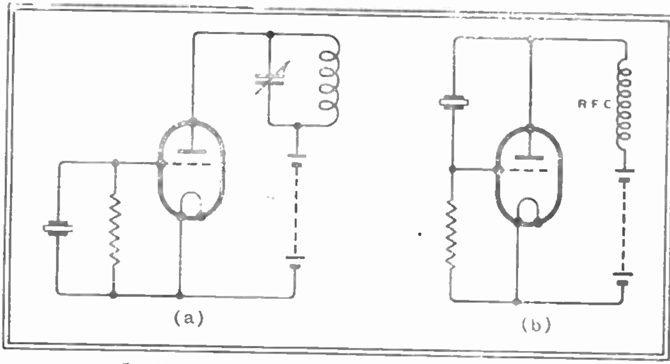

Fig. 83 : Two types of crystal-controlled oscillator circuit

coil-and-condenser tank circuit, there is practically nothing that can be altered about it by fair wear and tear. There is a small change of frequency with temperature ; and where requirements are stringent the crystal is kept in a temperature-controlled box.

The crystal is mounted between two flat metal plates, generally with a small air gap intervening. Fig. 83 shows two of several crystal oscillator circuits.

Imagining the crystal to be a tuned circuit, *a* is seen to be a T.A.T.G. circuit (compare Fig. 78), and *b* is a Colpitts circuit (compare Fig. 77). In *a* the tuning of the anode circuit is adjusted to a point at which oscillation takes place.

Cutting quartz crystals for oscillators is a scientific art, and unless competently done it is generally found that oscillation is possible at a number of frequencies, which may be sufficiently close together to make it uncertain which is the right one.

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93. Telegraph Transmitters : Keying

We have now reached the point—in theory at any rate—at which we can send out into space a continuous stream of waves of constant frequency and (if desired) high power. A transmitter sending out a continuous wave of this type can convey no more information by it than can a stationary lighthouse sending out a steady beam of light. Lighthouses are accustomed to announce their identity to mariners by periodic interruption of their light, according to an arranged " call sign " of long and short flashes. In the same way a radio transmitter can convey messages by a periodic interruption of its wave, breaking it up into short and long bursts that represent the dots and dashes of the Morse code.

To do this it is necessary to connect a Morse key—which is simply a form of switch that makes a contact when pressed and breaks it when let go—in such a way that it starts and stops the radiation from the transmitter.

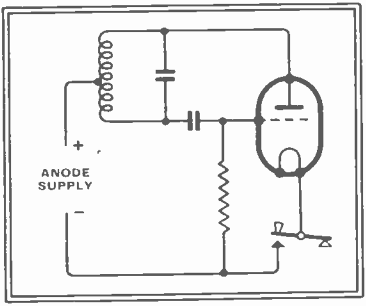

Fig. 84 : The essentials of a convenient method of keying transmitters

It might appear that almost any part of the circuit would do. In practice there are quite a number of things that have to be considered. The key should not be asked to break large amounts of power, because the resulting sparking would soon burn the contacts away. It cannot break a very high voltage, which would just spark across ; and there is safety to consider, too. Therefore the anode circuit is practically ruled out, and so is the aerial circuit. The filament circuit is no good, because of the time taken to heat and cool. And wherever the keying is done it must not affect the frequency of oscillation, so it is better not to keep stopping and starting the master oscillator.

At one time, to avoid keying high-power circuits, the key was made to shift the frequency slightly instead of stopping the transmitter altogether ; but there are now so many stations working that an intolerable amount of interference would be caused by the " spacing " wave.

A popular method of keying is shown in Fig. 84, in which the grid circuit is interrupted, but kept connected to the negative end of the high voltage anode supply. Any tendency for current to continue to flow through the valve causes the cathode to be charged positively (because it loses electrons) and so the grid becomes relatively more negative and prohibits the current.

94. Radiotelephony and Broadcasting : Modulation

To convey speech and music, something more elaborate than simple interruption of the transmission is required. It must be, as it were, moulded into the shape of the sounds. This process is called *modulation*, and the raw material that is modulated is called the *carrier wave*. Keying is an extreme and special case of modulation, in which the strength of the carrier wave is made to vary suddenly at intervals between zero and full power.

Suppose, however, that it is desired to transmit a pure note of 1,000 c/s, available as an electric current derived, in the first place, from the microphone before which the note is being sounded (how, we shall see shortly). This current will have the form of a sine wave. Meanwhile, the transmitter is busy generating a radio wave, which will have the same form but a much higher frequency.

If the wave corresponds to 300 metres, or a frequency of 1,000,000 c/s, each audio cycle will extend over a thousand radio cycles. To enable the musical note to be conveyed by the carrier wave these two oscillations have to be combined to make a single whole.

In Fig. 85 *a* is depicted the wave-form of a radio-frequency carrier wave, while *b* shows the musical note which we wish to combine with it. At first sight it might seem that it would be sufficient to add the two currents together and allow them both to flow in the aerial. Such a mode of combining them results in the wave-form shown in full-line in Fig. 85*c*. Examination of this figure will show that the two currents, although they are flowing simultaneously in the same circuit, are still independent, the whole consisting of the original radio-frequency current oscillating round a zero voltage which moves slowly up and down at the frequency of *b*. The dotted curve shows the new zero voltage. Successive peaks of the radio-frequency voltage are still exactly alike, as they were in *a*.

In view of the known fact that an aerial will not radiate an audio-frequency voltage to any appreciable extent, it is clear that if an attempt were made to send out *c* as a signal, the radio-frequency component would set up its usual wave, as at *a*, in which the audio-frequency component would not be represented.

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It is clear, therefore, that simple *addition* of the currents will not provide us with a resultant current of a suitable type for radiation ; we will therefore try *multiplication*.

Let us suppose that the amplitude of the radio-frequency current in the aerial depends upon the D.C. voltage used to drive some part of the transmitter. The height of curve *a* might then be 1 amp. if a 500-volt battery were used, but might rise to 1·5 or drop to 0·5 amps. if the battery were suitably

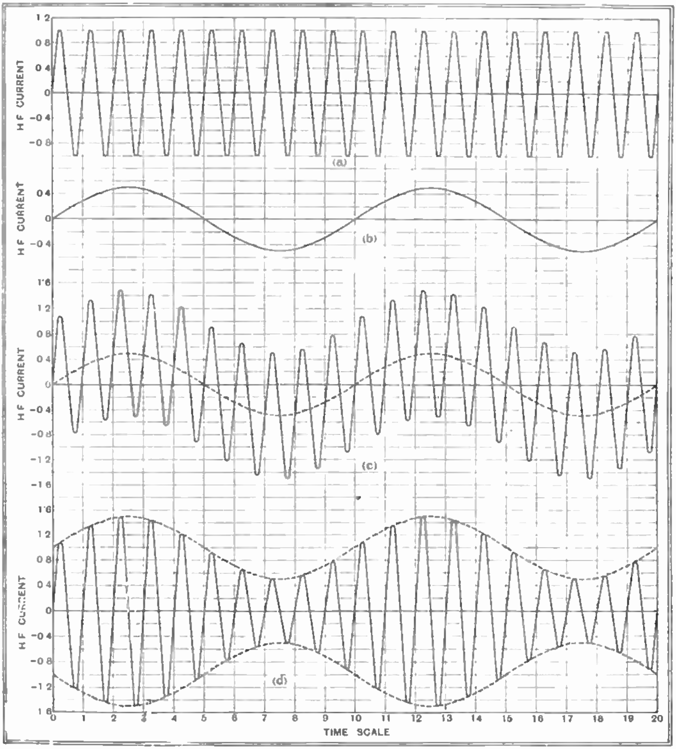

Fig. 85: Diagrams *a* and *b* show a radio-frequency current and a musical note to be combined with it for transmission. Mere addition of the currents results in *c*, in which the currents remain separate so that only the R.F. component would be radiated. Diagram *d* shows the modulated carrier resulting from multiplying the curves as described in the text : it is radiated complete from the aerial

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increased or decreased in voltage. We might now introduce the audio-frequency voltage we desire to transmit *in series* with this imaginary battery; then the total voltage reaching the transmitter would swing about its mean value, the audio-voltage alternately adding to and subtracting from the battery voltage. In consequence the amplitude of the radio-frequency output from the transmitter would also rise and fall, this rise and fall being strictly in time with the audio-frequency voltage we wish to transmit.

The result of this more elaborate means of combining the two curves, which amounts to multiplication of the one by the other, is shown at *d* in Fig. 85, where it will be seen that the amplitude of the radio-frequency current is now actually changing at audio-frequency. Except as an impress on the total amplitude of swing the audio-frequency current has disappeared; it is now represented by the *envelope* (dotted) of the curve as a whole.

A curve such as *d* represents a *modulated* high-frequency current or voltage. It is fairly evident that if this is allowed to flow in an aerial the radiated wave will follow, in its rise and fall, the rise and fall of the current, since the whole is now a radio-frequency phenomenon.

The observant reader will have noticed one important inaccuracy in the diagram; it does not bring out clearly enough the enormous difference in frequency between the carrier and the modulation. If, as suggested, *b* shows a 1,000-cycle (1 kc/s) note, *a* represents a 10-kc/s carrier, having a wavelength of 30,000 metres. To show a 1,000-kc/s (300-metre) carrier in its correct relationship to *b* there should be 100 complete radio-frequency cycles in the place of every one shown. A little imagination must therefore be applied to Fig. 85 before it can give a correct impression of a normal broadcast wave.

Even so, *d* represents nothing more exciting than a tuning-note; for music or speech the form of *b* is extremely complex, and this complexity is faithfully represented in the envelope of the modulated carrier *d*. Nevertheless, the diagram gives a very fair mental picture of the modulated carrier which flows, as a current, in the transmitting aerial, and is radiated outwards through space as a wireless wave.

In Fig. 85 and elsewhere "HF" (high-frequency) is used to refer to the frequency of the carrier wave, following a general custom. But whereas 5,000 c/s is "high frequency" compared with, say, 50 c/s, it is not suitable for a carrier wave. It is less confusing, therefore, to use the term "radio frequency" (RF) in place of HF to mean a frequency that is suitable for radiating;

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and "audio frequency" (AF) in place of LF to mean a frequency that is audible. As there are a number of these confusing equivalent terms in use, a list of them is given at the end of the book.

95. Depth of Modulation

Since our receiver will be so designed that the carrier wave itself, in the intervals of modulation (between items in the programme), gives rise to no sound in the loudspeaker, it is evident that a curve such as that of Fig. 85 *d* represents a note of some definite loudness, the loudness depending on the amount by which the radio-frequency peaks rise and fall above and below their mean value. The amount of this rise and fall in proportion to the normal amplitude of the carrier wave is spoken of as the *depth of modulation.*

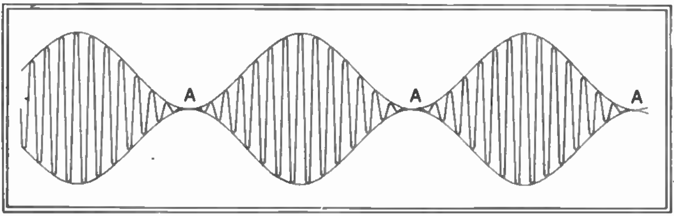

Fig. 86 : Carrier-wave modulated to a depth of 100%. At its minima (points A) the R.F. current just drops to zero : any attempt at still deeper modulation results in a series of separate bursts of current, and the envelope no longer has the form of the modulating wave

For distortionless transmission the increase and decrease in carrier amplitude that correspond to positive and negative half-cycles of the modulating voltage must be equal. It is evident that the maximum possible decrease in carrier-amplitude is found when the modulation reduces the carrier so far that it just, and only just, ceases at the exact moment of minimum amplitude, as shown in Fig. 86. At its maximum it will then rise to double its steady value. Any attempt to make the maximum higher than this will result in the carrier actually ceasing for an appreciable period at each minimum ; over this interval the envelope of the carrier-amplitude can no longer represent the envelope of the modulating voltage, and there will be distortion.

When the carrier has its maximum swing, from zero to double its mean value, it is said to be modulated *to a depth of* 100 *per cent*. In general, the maximum rise in amplitude, expressed as a percentage of the mean, is taken as the measure

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of modulation depth. Thus a rise from 1 volt to 1·5 volt corresponds to 50 per cent. modulation, a rise to 1·4 volt to 40 per cent., and so for other values.

In transmitting a musical programme, variations in loudness of the received music are produced by variations in modulation-depth, these producing corresponding changes in the audio-frequency output from the receiver.

96. Methods of Modulation

In showing, by means of Fig. 85, what has to be done to modulate a carrier wave, it was suggested that one possible way of doing it is to connect a source of audio-frequency voltage

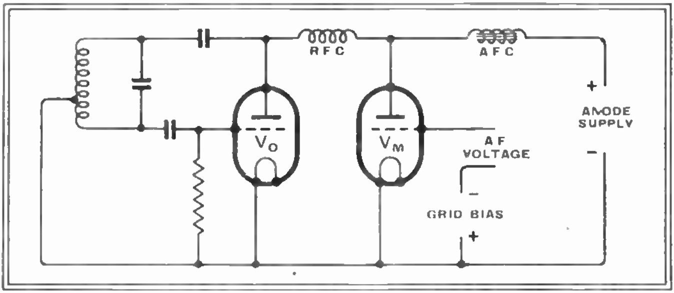

Fig. 87 : Anode modulation, sometimes called " choke control " modulation because the audio voltage is developed across a choke, A.F.C., which forms the anode load of the modulator valve V_M and causes the anode voltage delivered to the oscillator valve V_O to vary in accordance with the audio voltage. V_M acts simply as a high-power amplifier. R.F.C. is a R.F. choke for keeping R.F. current out of the modulator circuit

in series with the source of anode voltage. See Fig. 87, where the A.F. voltage is amplified by a modulator valve V_M which develops the amplified voltage at its anode, across the A.F. choke, and therefore in series with the anode supply to the oscillator valve V_O. A choke is preferred to a resistor, because of the large power to be carried. Suppose, for example, that the oscillator valve is normally fed with 5,000 volts. To reduce the amplitude of oscillation to the verge of zero it would probably be necessary to reduce this to a few hundred volts ; and to double the amplitude it might be necessary to raise it to nearly 10,000. So to effect 100 per cent. modulation the peak amplitude of the audio voltage would have to be nearly 5,000. This source of audio voltage would also have to handle the full anode current of the oscillator, so the amount of audio-frequency power to be supplied is comparable with the power

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to be supplied to the oscillator. This is quite serious in a high-power transmitter, involving large and expensive modulator valves as well as a considerable addition to the power consumption.

Another drawback of this simple arrangement is that the frequency of oscillation depends slightly on the anode voltage, and consequently varies during the cycle of modulation. This rules the method out from most modern practice, in which the frequency of the carrier wave must be kept constant within extremely narrow limits. We fall back, therefore, on the M.O.P.A. system and modulate the amplifier stage by connecting the source of A.F. voltage in series with its anode supply,

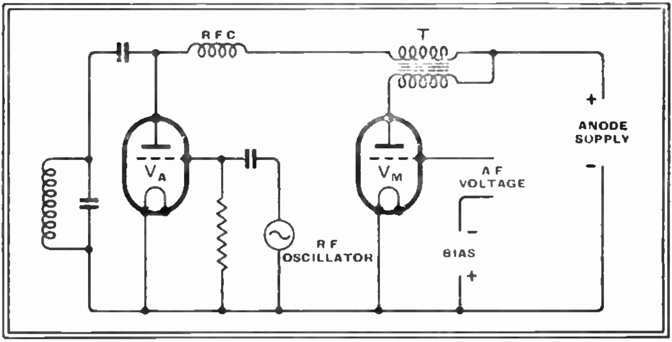

Fig. 88 : Modulating the output from a Class C. R.F. amplifier V_A driven by a separate oscillator, to avoid modulation of frequency as well as amplitude. Note the alternative method of coupling, by transformer T instead of choke as in Fig. 87

taking care that the modulation process has no effect on the frequency of the oscillator.

Fig. 88 is a skeleton circuit of such a system, which is very commonly used. It also shows an alternative method of bringing the A.F. voltage to bear on the anode circuit of the R.F. valve, namely, by transformer instead of direct coupling. This has certain practical advantages. The necessity for providing a large amount of A.F. power remains, however.

So various methods have been tried for overcoming this. One scheme is to generate a low power carrier wave, modulate it by means of correspondingly low-power equipment, and then amplify the modulated wave.

Another line of attack is to apply the modulating voltage to the grid, which can exert a large controlling influence by means of relatively small voltage changes, without involving appreciable

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current flow. If applied to the oscillator grid, we get the undesirable frequency variations ; and in any case the amplitude of oscillation does not depend in a straightforward manner on the grid voltage, so serious distortion results. By modulating at the grid of a Class C amplifier, a reasonably distortionless result is possible by careful adjustment, but only if the R.F. output is restricted by avoiding grid current. Even if this restriction is not observed, the power-efficiency of the amplifier is still much lower than with no modulation, so the saving in power over the anode modulation system is largely an illusion.

There are a number of other methods of modulation, some of which are of practical value, especially in low-power transmitters. But the anode method, of which there are many variations, is the one generally favoured where freedom from distortion is important.

97. Frequency Modulation

Instead of keeping the frequency constant and varying the amplitude, one can do *vice versa*. Fig. 89 shows the difference ;

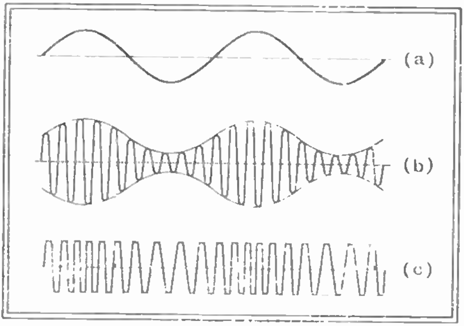

Fig. 89 : Curve *a* represents 2 cycles of audio frequency used to modulate a radio-frequency carrier wave. If amplitude modulation is used, the result is represented by *b* ; if frequency modulation, by *c*

a represents a small sample—two cycles—of the audio-frequency programme, and *b* is the corresponding sample of radio-frequency carrier wave after it has been amplitude-modulated by *a*. If, instead, it were frequency-modulated it would be as at *c*, in which the amplitude remains constant but the frequency of the waves increases and decreases as the A.F. voltage rises and falls. In fact, with a suitable vertical scale of cycles per second *a* would represent the frequency of the carrier wave.

An advantage of frequency modulation, or "F.M.", is that under certain conditions reception is much less likely to be disturbed by interference or noise. These conditions involve the use of very high frequency carrier waves; about fifty times as high as the "medium" broadcast frequencies (550–1,500 kc/s).

The ordinary type of receiver used for amplitude modulation does not respond properly to F.M., but a rough idea of the principle of F.M. reception can be gained by looking at a resonance curve, say Fig. 52. If the mean frequency of the carrier wave were 990 kc/s, then, as the frequency rose and fell due to modulation, the voltage across the tuned circuit would rise and fall, as it would if an amplitude-modulated wave were applied.

F.M. has come into considerable use for broadcasting in America, and for army communications, and is likely to have an important future.

98. Microphones

So far we have learnt something about two of the boxes at the transmitting end of Fig. 4—the R.F. generator and the modulator.

The device by which sound, consisting of air waves, is made to set up electric currents or voltages of identical waveform, which in turn are used to control the modulator, is one that most people handle more or less frequently, in telephones. It is the *microphone*; and the type used for most radio purposes other than broadcasting, and in telephones everywhere, is essentially the same as the original invention by Hughes nearly 70 years ago.

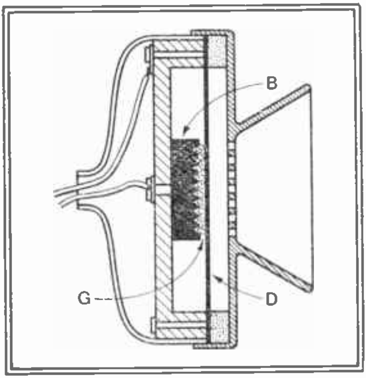

Fig. 90: Section of a typical carbon microphone

Fig. 90 shows the principle. The sounds are directed on to a thin carbon diaphragm D, behind which is fixed a carbon button B, generally with a grooved or honeycombed inner surface. The space between the two contains carbon granules which are prevented from falling out by some such device as a swansdown ring around the edge of the button. Connections are taken from the diaphragm and the button. Details vary (for example the diaphragm may be of metal, carrying a carbon button) but the general idea is the same: as the diaphragm vibrates under the alternating pressure of the air waves of sound, it varies the pressure applied to the numerous carbon contacts, and so varies the resistance between the terminals.

A carbon microphone is, in fact, a resistance that varies to correspond with the sound waves, and so the current delivered by the battery in Fig. 91 varies in the same way. As this current passes through the primary of a transformer, a varying voltage is produced at the secondary terminals. Sound waves of normal intensity are very feeble, and even with a step-up transformer the result is generally only a few volts at most: so for modulating a powerful transmitter several stages of amplification may be necessary.

While such a microphone can be designed to give a reasonably large output over the range (or band) of frequencies corresponding to the essentials of intelligible speech, it does so by virtue of a certain amount of mechanical resonance, and many of the frequencies needed to give full naturalness to speech and music are not fairly represented. If the resonance is reduced in order to give more uniform response to all the audio frequencies, then the output becomes less, and more amplification is needed. Compare electrical resonance, as in Fig. 53. The carbon microphone also has other defects when considered for broadcasting and not just for communication. So various other systems have been tried; but generally speaking it is still true that the higher the quality—i.e., the more faithfully the electrical output represents the sound—the lower the output.

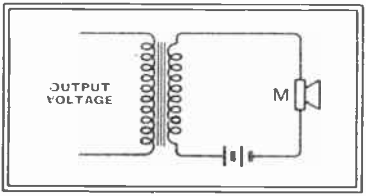

Fig. 91 : How a microphone (M) is connected, to give an A.F. voltage

The condenser or electrostatic microphone consists of a diaphragm, generally metal, with a metal plate very close behind. When vibrated by sound, the capacitance varies, and the resulting charging and discharging currents from a battery, passed through a high resistance, set up A.F. voltages. Its disadvantages are very low output and high impedance.

The crystal or piezo-electric microphone depends on the same properties as are applied in crystal control (Sec. 92); the

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varying pressures directly give rise to A.F. voltages. Crystals of Rochelle salt are particularly effective.

The type most commonly used for broadcasting and the better class of " public address " is the electromagnetic. There are many varieties : in some an iron diaphragm varies the magnetic flux through a coil, so generating voltages ; in others the diaphragm bears a moving coil in which the voltages are generated by flux from a stationary magnet ; while in the most favoured types a metal ribbon is used instead of the coil.

99. Coupling Transmitter to Aerial

The last main component of a transmitter shown in Fig. 4 is the aerial. This is the subject of Chapter 20 ; but one aspect that we can consider at this point is the way in which the transmitter is connected to the aerial, because the power efficiency is greatly affected thereby.

At the end of Sec. 72 we saw that the maximum output is obtained from a valve when the load resistance is equal to the internal resistance of the valve. See Fig. 68. In practice, maximum output may not be the only thing to consider. If by a small sacrifice in output a very considerable increase in efficiency is obtained, then it may be advisable to adopt that policy, which probably means working with a different load resistance. Usually the valve manufacturer specifies the *optimum load resistance* for his valves under certain conditions.

What is the load resistance of a transmitter valve ? If it is a master oscillator, the load is its anode tuning circuit, combined with the input of the power amplifier. The load of the P.A. or of an oscillator feeding the aerial direct, is its anode tuning circuit combined with the aerial.

Any power used up in the tuning circuit itself is wasted ; ideally that circuit would have no loss and would merely be an infinite resistance in parallel with the true load—the aerial. Leaving detailed discussion of aerials as loads to Chapter 20, we can assume in the meantime that their impedance varies greatly with frequency. If by some stroke of luck the aerial impedance at the working frequency were to equal the optimum load resistance for the transmitting valve we could connect the aerial and earth terminals to opposite ends of the anode tuning coil. In general, however, the effective impedance between these terminals will not be equal to the optimum for the valve. A *matching* device is necessary.

In Sec. 48 we saw that a transformer is just such a device, because by a suitable choice of turns ratio any resistance can be made equivalent to any other resistance. The rule was that

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to make R_s equivalent to R_p, a transformer is interposed, R_s being connected across the secondary winding, which has $\sqrt{R_s/R_p}$ turns for every primary turn. For example, if our tuning coil has 40 turns, and the optimum load for the valve is 3,000 ohms, and at the working frequency the aerial is equivalent to a resistance between aerial and earth terminals of 120 ohms, then it should be coupled to the tuning coil by a secondary winding of $40 \times \sqrt{120/3{,}000} = 8$ turns, as in Fig. 92*a*.

Alternatively, an auto-transformer (Sec. 48) may be used (*b*). There are other ways of matching such as *c* (which like *a*, insulates the aerial from the anode coil, which may be at

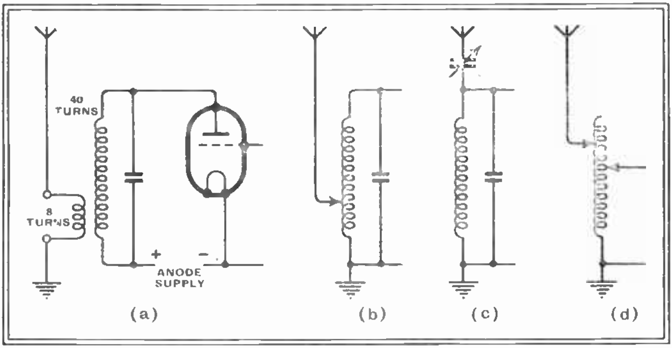

Fig. 92 : Some of many varieties of method in coupling aerial to transmitter. *a* is the inductive coupling or R.F. transformer method, worked out in the text for a particular example ; *b* is the aerial tapping or autotransformer ; *c* is the series condenser method ; and at *d* the valve is tapped down to reduce the effective load resistance

anode potential) or *d* (which is used when the optimum load is *less* than the aerial impedance).

Although it is easy with this simple theory to work out the right number of turns (and rather less easy to calculate the right value of capacitance in Fig. 92*c*) there are complications that make it a more formidable task. For example, we assumed that *all* the magnetic lines link both primary and secondary coils, whereas this is far from being so in R.F. transformers. In practice, the right tapping or coupling is calculated as nearly as possible, and provision made for it to be varied until the best results are obtained. Fig. 68 shows that it is necessary to depart quite a lot from the optimum before there is a serious loss in output.

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Another complication is that an aerial nearly always has reactance as well as resistance, which affects the tuning and makes it necessary to use less or more tuning capacitance than if the aerial were not coupled to the tuning coil ; less, if the reactance effectively in parallel with the coil is capacitive, and greater if it is inductive.

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CHAPTER 10

DETECTION

100. The Need for Detection

TURNING now to the reception end in Fig. 4 we see a number of boxes representing sections of a receiver. Although all of these sections are included in most receivers, they are not all absolutely essential. So we begin with the detector, because it is the one thing indispensable, for reasons now to be explained.

At the receiving aerial, the modulated carrier-wave sets up currents which, apart from any distortion they may have suffered in their journey through space, are an exact duplicate in miniature of the currents in the aerial of the transmitter. By some simple circuit, such as that of Fig. 93 (compare Fig. 92), they can be collected and caused to flow, in magnified form, round a tuned circuit. The function of even the simplest receiver is to convert these

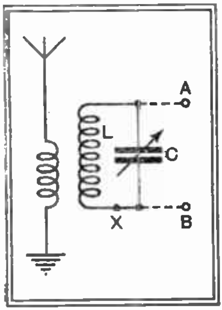

Fig. 93 : Showing how the modulated carrier is "collected" by the aeriel in the form of currents of the same wave-form and passed to the tuned circuit LC as the first stage in reception

electric currents into sound so that the programme may be enjoyed.

If earphones (which consist essentially of an electro-magnet, close to which is a thin iron diaphragm vibrated by the signals passed through the magnet coil) were connected to the circuit, either by inserting them at X to allow the circulating current to flow through them or by joining them across A and B so that the voltage on C would drive a current through them, no sound would be heard. The reason for this can readily be appreciated by considering Fig. 94 *a*, which repeats the diagram of the modulated carrier. Any two consecutive half-cycles of the current are approximately equal (in a practical case, much more nearly equal than in the diagram) and so neutralize each other so far as the earphone diaphragm is concerned, it being understood that this cannot possibly vibrate at anything like such a high frequency as that of the carrier. The average current through the phones, even measured over an interval as short as a ten-thousandth of a second, is therefore zero.

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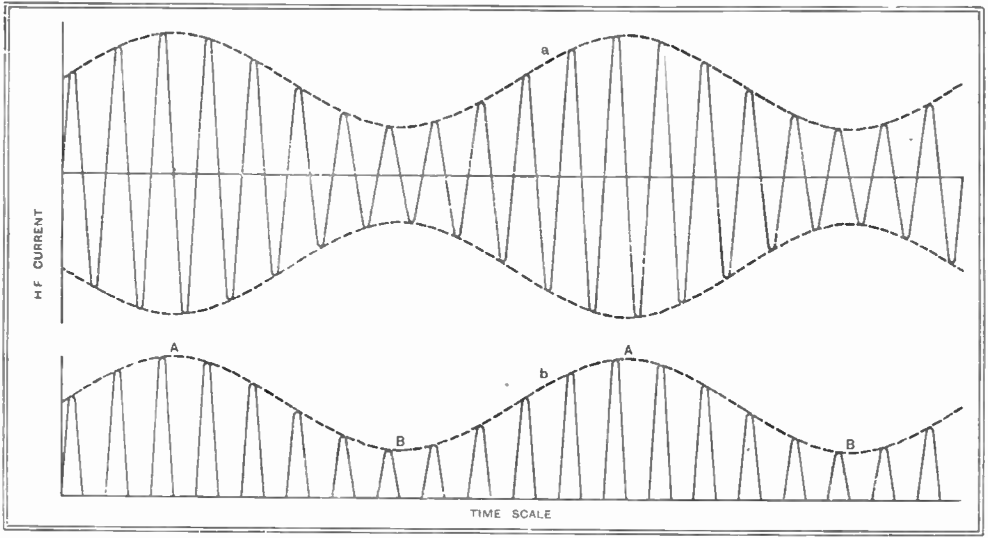

Fig. 94 : Modulated R.F. current (diagram *a*). Over any period of time appreciably greater than an R F. cycle, the average current is zero, and so is inaudible in earphones. The same current, rectified, is shown in diagram *b*. The average current now rises and falls at modulation-frequency and is heard in earphones

But if we could find some means of wiping out one-half of the wave, so that it took on the form shown at *b*, we should have a current to which the phones could respond, for the average current would then be greater at A than at B. While unable to follow the carrier-frequency alternations individually, the phone diaphragm would then rise and fall at the rate of their variation in amplitude. Since these variations are due to the audio-frequency note modulating the carrier, it is this note, which we want, that would be heard.

The process of suppressing half of a complete wave, thus converting alternations of current into a series of pulses of unidirectional current, is called by the general term *rectification*. The particular case of rectifying a modulated carrier in such a way as to reveal the modulation is known in this country as *detection*, and in America as *demodulation*. It can be performed by any device which conducts current, or responds to a voltage, in one direction only; or, less perfectly, by any device which has a lower resistance to currents, or a greater response to voltages, in one direction than in the other.

101. Types of Detectors

A perfect detector would be one that had no resistance to current flowing in one direction, and an infinite resistance to current in the opposite direction. It would, in fact, be equivalent to a switch, completing the circuit during all the positive half-cycles in Fig. 94*a*, thus enabling them to be passed completely, as in *b*; and interrupting it during all the negative half-cycles, suppressing them completely, as also shown in *b*. For very low frequency alternating currents, such as 50 c/s, it is possible to construct such a switch (known as a vibrator rectifier); but any device involving mechanical motion is hardly practicable at radio frequencies.

There are a number of minerals that have a lower resistance to current in one direction than the other, and they were at one time very largely employed as detectors in *crystal* receivers.

More recently the manufactured copper oxide rectifiers used extensively for low frequency rectification have been adapted for radio frequencies and are sold under the trade name "Westector". But the effect most widely used is the one already met in Sec. 64—the one-way movement of electrons in a vacuum—of which the thermionic valve is the embodiment.

102. Detector Characteristics

To understand the various available types of detectors it is necessary to make use of characteristic curves, in which the

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relation is shown between applied voltage and the resultin current. These have already been used in connection with valves (e.g., Fig. 64), but it may be as well to make sure tha their significance is grasped. In Fig. 95, the current flowin in a conductor is plotted against applied voltage. In ar ordinary conductor having a definite unaltering resistance, th graph is a straight line; for example, AOB, which, by Ohm' Law, can be seen (Sec. 11) to represent 2,000 ohms. The lin COD represents 500 ohms.

The steeper the slope, the lower the resistance, as we sav in Sec. 65. The line FOG represents zero resistance, and HOJ infinite resistance; so FOJ would be the graph of

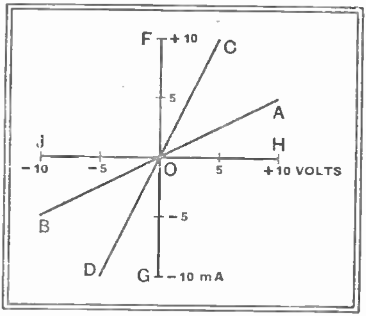

Fig. 95: Characteristic curves of various "linear" resistances. The line COD, by Ohm's Law, is seen to represent 500 ohms. If the resistance to negative voltages is different from that to positive, as shown for example by the line COB, the result is a partial rectifier

perfect rectifier. COB would be an example of a partia rectifier—a resistance of 500 ohms to positive currents anc 2,000 ohms to negative currents. Now apply an alternating voltage to a circuit consisting of any of the foregoing imaginary resistances in series with a fixed resistance of 1,000 ohms (Fig. 96*a*). The voltage across the 1,000 ohms—call it the output voltage—is equal to the applied voltage only if R is zero. Both voltages are then indicated by the full line in Fig. 96*b*. If R were 500 ohms (COD), the output would be represented by the dotted line CD, and if 2,000 ohms by the dotted line AB.

103. Load Resistance and Output

If now a rectifier is substituted for R, the symbol being as shown in Fig. 96*c*, the positive and negative output voltages

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are unequal. For example, the perfect rectifier (FOJ) gives the output FJ in *b*; and the partial rectifier gives CB. The *average* voltage, taken over each whole cycle, is, of course, nil when R is not a rectifier, because the negative half-cycle cancels out the positive. Using the perfect rectifier, the average of the positive half is 63 per cent. of the maximum, while the negative is, of course, nil; so the average rectified D.C. taken over a whole cycle is 32 per cent. of the peak alternating voltage. Using the rectifier with the graph COB, the negative half-cycle (B in Fig. 96*b*) cancels out half of the positive half (C), which is two-thirds of the peak; so the resulting D.C averages over a whole cycle is less than 11 per cent. of the peak A.C. One could present this information in the form shown in Fig. 97. Here the D.C. volts, obtained by averaging the output across the 1,000 ohms, is plotted against the peak alternating voltage.

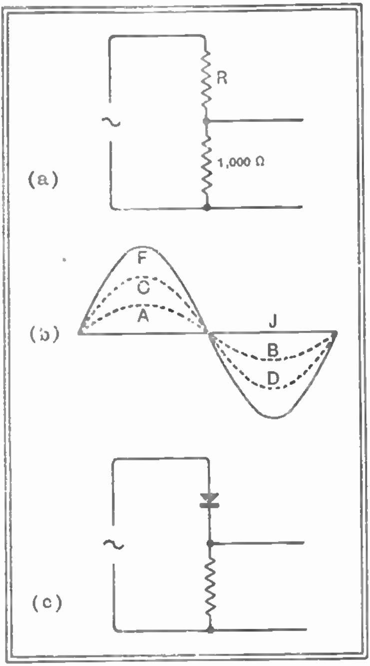

Fig. 96: The results of applying a sine wave voltage to resistances represented by Fig. 95, in the circuit shown here at *a*, are indicated by *b*, which is lettered to correspond with the various characteristics in Fig. 95. When R is a rectifier, it is represented by the symbol shown in *c*

The characteristics of the rectifiers represented by FOJ (perfect) and COB are shown. The resistance characteristics—those shown in Fig. 95—of all real rectifiers are curved; there is no abrupt change from low to high resistance. Look at Fig. 62. If the applied A.C. voltage is small, the rectifier is generally much less perfect than if it is large. So, instead of being straight lines, the graph in Fig. 97 is also curved: it might be like line COB to start with, rising nearly to the steeper slope of the perfect rectifier at large voltages. The effects of this will be con-

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sidered in detail shortly. In the meantime we should remind ourselves that we are interested not in a single cycle, but in vast numbers of them per second; so many that even over a very small space of time it is not possible to draw them individually. Fig. 98 shows them as a shading, and after

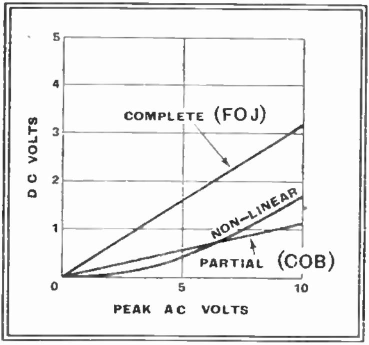
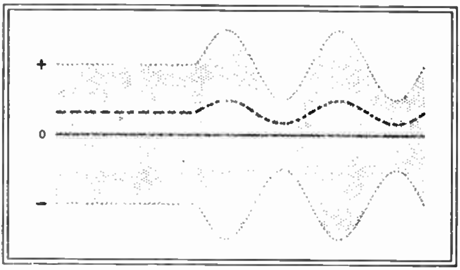


Fig. 97 : The average D.C. outputs given by applying A.C. to various rectifiers are shown in these characteristic curves, lettered to correspond with Figs. 95 and 96

a preliminary period in which they maintain a constant amplitude, representing an unmodulated carrier wave, they are modulated at an audio-frequency (compare Fig. 94*a*). The D.C. output obtained from them by a perfect rectifier would

Fig. 98 : The dotted line indicates the rectified output obtained from the R.F. input, shown first unmodulated, and then modulated 50 per cent. The shading represents cycles at too high a frequency to be distinguished individually



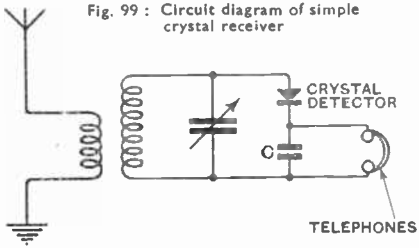
be about one-third of the peak voltage, and is represented by a dotted line. Where the amplitude of the carrier fluctuates, the output fluctuates in proportion, being now an audio frequency voltage capable of operating phones or (if strong enough) a loudspeaker.

104. The Simplest Complete Receiver

What is necessary, then, to form a complete radio receiver, is shown in Fig. 99. Here the modulated R.F. voltages, represented by the shading in Fig. 98, are developed across the tuned circuit; and the crystal rectifier, by virtue of its differing resistances to positive and negative half-cycles, passes through the phones a balance of current in one direction, which unidirectional current is at least approximately proportional to the amplitude from moment to moment of the waves radiated from the distant transmitter, and therefore is a copy of those set up by sound waves impinging on the microphone. And so the sound is reproduced.

In the simple crystal set of Fig. 99, the purpose of every

Fig. 99: Circuit diagram of simple crystal receiver



component part, except the condenser C, has been indicated. In arriving at the steady or slowly varying voltage (or current) represented by the dotted line in Fig. 98 we have spoken of "averaging" the unequal positive and negative R.F. voltages resulting from the action of the rectifier. The condenser does this averaging or smoothing out. When the applied voltage is near its peak in the forward direction the current through the phones cannot rise so rapidly (Sec. 29) and the spare current is taken in by the condenser, which consequently increases its charge. During the intervals between one positive half-cycle and the next, when the current coming through the rectifier is nil or even negative, the condenser keeps things going by parting with some of its charge. It acts in much the same way as the bag in bagpipes, keeping them playing while the player draws his breath.

The circuit of Fig. 99 is that of a simple receiver, but it contains the kernel that every receiver must have. The two

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essentials are tuning, to select the required signal; and detection, to extract the A.F. currents from the received carrier In addition, of course, phones or loudspeaker are needed to produce air waves from these currents.

We may add more tuned circuits to increase selectivity, and amplifiers operating on the signal either before or after detection, or both, to render the set more sensitive. But all these are mere elaborations; the crystal set described contains the essentials of tuning plus detection upon which every set, however ambitious, ultimately depends.

105. The Diode Detector

The thermionic diode valve (Sec. 64) has one strong point in its favour. Although it may not be perfect in the forward direction, for it always has some resistance, it is practically perfect in the reverse direction; that is to say, it passes negligible reverse current, even when quite large voltages are applied.

This fact enables the loss due to the forward resistance to be almost entirely avoided, and is bound up with choice of the resistance across which the output is derived—1,000 ohms in Fig. 96. In the simple set of Fig. 99, the aim is to obtain the maximum rectified *power*. The sound given by the phones depends on the strength of magnetic field produced in them, which in turn depends on the magnetising ampere-turns, which require both current and voltage (to force the current through a sufficient number of turns). But in most sets, and practically all those employing valves, the rectified waves are not used directly in the sound-producing device but are amplified by a valve. Valves, as normally used, are voltage-operated devices; and therefore every effort is made to extract the maximum rectified voltage even if the current is thereby reduced.

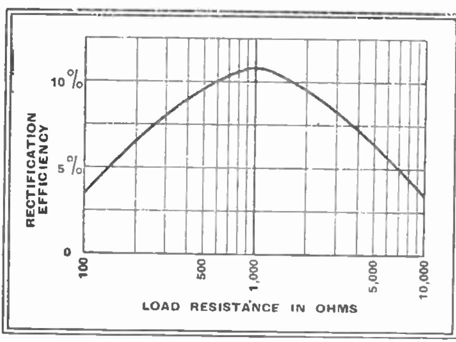

Fig. 100: Relationship between load resistance and rectification efficiency for the circuit Fig. 96

Referring back to Fig. 96, and assuming the rectifier to have 500 ohms forward resistance and 2,000 ohms backward, a load resistance of

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1,000 ohms would have developed across it an average D.C. voltage equal to 10·6 per cent. of the A.C. peak, as already explained. Performing the same calculation for other values of load resistance, we get the result shown in Fig. 100, showing that with this imaginary rectifier a load resistance of 1,000 ohms gives the greatest voltage output,* one-third of that for the perfect rectifier.

If the backward resistance were infinitely large, then none of the forward rectified current would be neutralised by backward current ; so even if the forward current were very small (due to a high forward resistance) the output voltage could, theoretically at least, be made to approach that given by a perfect rectifier, merely by choosing a sufficiently high load resistance. In practice, however, there are limits to the resistance that can be used. With the diode it is usual to make the load resistance 0·1M Ω to 1M Ω or even more, and the output approaches the theoretical maximum quite closely, provided that the applied voltage is not too small. In fact, by making use of the reservoir condenser (C in Fig. 99) it is possible to approach a rectification efficiency of 100 per cent., instead of the 32 per cent. that is the maximum without it ; and so do nearly three times better than our "perfect" rectifier. This point is important enough to justify closer consideration.

106. Action of Reservoir Condenser

Suppose, in order to obtain the utmost rectified voltage, we made the load resistance infinitely great. The reservoir condenser would then be in series with the diode, with no resistor across it (Fig. 101). To simplify consideration we shall apply a square wave, as in Sec. 35, instead of a sine wave ; and make its amplitude 100 volts. This applied voltage is shown as Fig. 102*a*. We shall also assume that its frequency is 500 kc/s, so that each *half* cycle occupies exactly one millionth of a second. At the start the

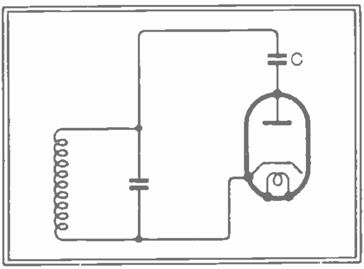

Fig. 101 : Diode detector with infinite load resistance

* In general, the load resistance that gives the greatest rectified voltage output is $\sqrt{R_F R_B}$, where R_F and R_B are respectively the forward and backward resistances of the rectifier. It is interesting to compare this with the amplifier (Fig. 68).

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condenser is uncharged, and therefore has no voltage across it. It can acquire a charge only by current flowing through the rectifier, which offers a certain amount of resistance, and so when the first positive half-cycle arrives its full 100 volts at first appears as a voltage across that resistance, as shown by the full line in Fig. 102*b*.

The current soon causes the voltage across the condenser to rise towards 100 volts, as shown by the dotted line. It is clear that the voltages across rectifier and condenser, being in series, must add up to give 100 volts so long as that is the applied voltage; and examination of Fig. 102*b* shows this to be so. The greater the resistance of the rectifier and the capacitance of the condenser, the longer the condenser takes to charge up,

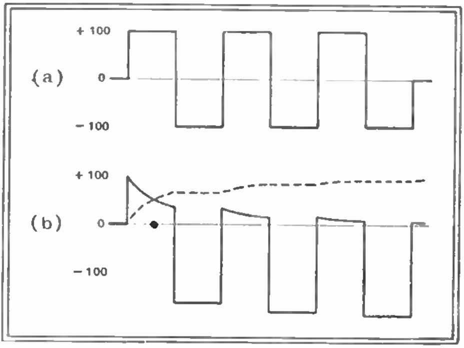

Fig.102: Analysis of application of a square alternating voltage *a* to the circuit Fig. 101. The dotted line in *b* represents the voltage across C; the full line, that across the diode

just as would a very large balloon inflated through a very narrow tube. As a matter of fact, if the capacitance in microfarads is multiplied by the resistance is megohms, the answer (known as the *time constant*) is the number of seconds required for the condenser voltage to reach 63 per cent. of the applied voltage (Sec. 23). Suppose, then, that the rectifier resistance (assumed constant) is 0·01 M Ω and the capacitance is 0·0001 μF. Then the time constant is 0·000001 second, or one millionth of a second. In this case that happens to be the time occupied by one half-cycle of the 500 kc/s applied voltage. So at the end of the first positive half-cycle the voltage across the condenser is 63, while the voltage across the rectifier has dropped to 100 − 63 = 37. Then comes the negative half-cycle. The diode ceases to conduct, and while the condenser therefore

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cannot charge up any more it likewise has no conducting path through which to discharge, and so remains at 63 volts until the second positive half-cycle. Meanwhile, the condenser voltage 63, together with the voltage across the rectifier, must be equal to the new applied voltage, -100. The voltage across the rectifier must therefore be -163.

The net voltage applied to the condenser when the second positive half-cycle arrives is $100-63 = 37$ volts, so at the end of this half-cycle the voltage across the condenser will increase by 63 per cent. of 37, or 23 volts, which, added to the 63 it already possessed, makes 86. The charge thus gradually approaches the peak signal voltage in successive cycles, while the average voltage across the rectifier falls by the same amount, as shown in Fig. 102*b*. The rectified output voltage is, therefore, 100 per cent. of the peak applied R.F. voltage. A similar result is obtained with a sine wave signal.

107. Choice of Component Values

This, of course, is excellent so far as an unmodulated carrier is concerned. When it is modulated, the amplitude alternately increases and decreases. The increases build up the condenser voltage still further ; but the decreases are powerless to reduce it, for the condenser has nothing to discharge through. In order to be able to follow the modulation it is necessary to provide such a path, which may be in parallel with either the condenser or the diode, so long as in the latter case there is a conducting path through the tuned circuit to complete the discharge route.

The resistance must be much higher than the diode forward resistance, or there will be loss of detector efficiency. Incidentally, there will also be damping of the tuned circuit. In practice one may reckon, as a close approximation, that the damping effect of a diode detector with a load resistance R is equivalent to connecting a resistance $R/2$ directly across the tuned circuit.* It is for these two reasons that R is generally

* This is quite easily proved. Denote the resistance which, if connected across the tuned circuit in place of the diode and its load resistance R and reservoir condenser C, would be equivalent to them by R'. Then assuming the resistance of the diode is negligible in comparison with R, C will charge up to the peak voltage $\sqrt{2}E$, E being the R.M.S. voltage developed across the tuned circuit. As R will be the only component dissipating power, the number of watts of which is equal to the square of the voltage across it divided by R (Sec. 14), so is $2E^2/R$. But from the way in which we have defined R', this wattage must also equal E^2/R'. Therefore $R' = R/2$.

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made not less than about 0·5 M Ω. But in order to follow modulation which may be as rapid as 10,000 c/s it is necessary for the time constant to be not greater than about one cycle of modulation, that is to say, 1/10,000th sec. The value of C is thereby fixed. If R is 0·5 M Ω, and C × R = 1/10,000, then C must be 0·0002 μF (or 200 pF), which is in fact quite a likely figure. If the product CR is increased, the higher modulation frequencies cannot be followed and they are consequently reproduced less strongly than they should be. If it is reduced, then the detector efficiency falls off and the damping on the tuned circuit is increased.

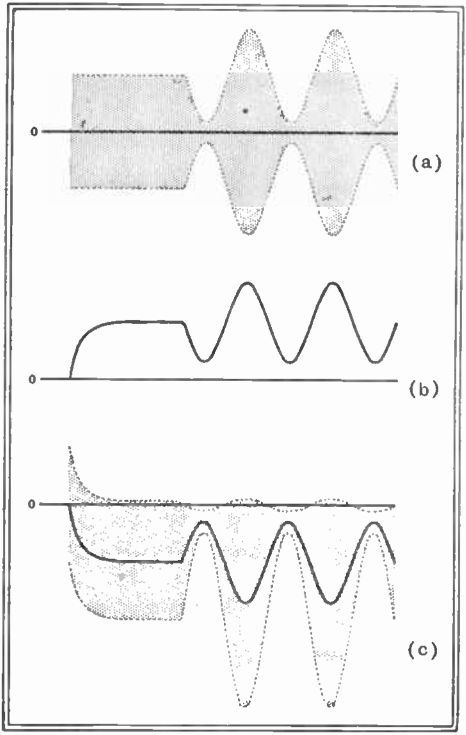

Fig. 103: When a radio-frequency wave a (shown first unmodulated and then modulated) is applied to a diode detector having a time constant intermediate between the periods of the modulation frequency and the radio frequency, the voltage across condenser is as at b, and across diode as at c, where the full line is the average of the R.F.

Fig. 102 shows the process in slow motion, as it were. Speeding up the carrier frequency until it appears a mere blur, so that the variations in its amplitude due to modulation can be shown as in Fig. 103*a*, the charging up of the condenser to the R.F. peaks appears as at *b*, and the voltage across the rectifier itself as at *c*. Here the *average* is indicated by a heavy line and is the inverse of the curve *b*.

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108. Varieties of Diode Detector Circuit

Two variations of the diode detector circuit have already been mentioned, and are shown in Fig. 104 *a* and *b*. The latter is necessary if the tuned circuit has a D.C. voltage between it and the cathode of the diode, as often happens in valve circuits (e.g., Fig. 122). As explained, the time constant of the condenser and resistor are such that the condenser voltage cannot follow the radio frequency to any great extent, but only the much slower audio-frequency variations. Consequently most of the R.F. voltage delivered by the tuned circuit appears in the output along with the A.F. As will be seen later, this

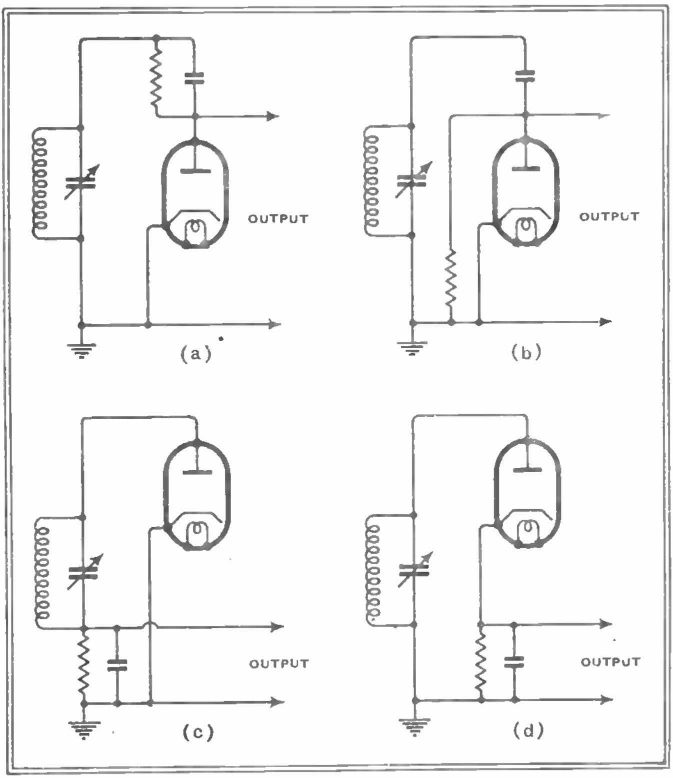

Fig 104 : Various methods of connecting a diode detector to a tuned circuit

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is a nuisance, and may be difficult to get rid of. So *c* and *d* are to be preferred where possible. Unfortunately *c* is not possible in those many cases in which it is essential for the tuning condenser to be earthed or at least at some steady potential. The circuit *d* avoids this difficulty, but runs into another, for the cathode is now at A.F. potential and the capacitance of the filament battery or heater transformer to earth may be objectionable, introducing hum. Sometimes, especially if followed by little amplification, *d* is satisfactory.

109. The Grid Detector

Mention of amplification brings us to the question of how to connect the diode detector to a stage of A.F. amplification

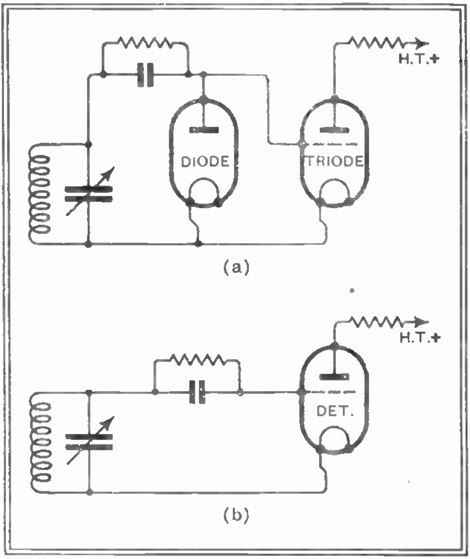

Fig. 105 : Skeleton diagram of diode detector followed by triode as amplifier of the detected signals. Compare this with a " grid detector " (diagram *b*) in which cathode and grid behave as a diode detector, the valve then amplifying the detected signals

(which may or may not be the valve that directly operates the loud speaker). When the detector circuit of Fig. 104*a* is connected straight to an amplifier, as shown in Fig. 105*a*, the anode-cathode path of the diode is in parallel with the grid-

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cathode path of the triode. Since both the anode of the diode and the grid of the triode consist simply of an electrode close to an emitting cathode, there would seem to be no need to have them both present. Experiment confirms this supposition; there is no change in the performance of the system if the diode is removed from its socket.

The simplified result is the well-known *grid detector* of Fig. 105*b*, in which the grid and cathode, acting as a diode, rectify just as in Fig. 104*a*. The A.F. voltages appearing at the grid as a result of this then control the electron stream through the triode and so produce an amplified voltage at the anode in the way described in Sec. 70.

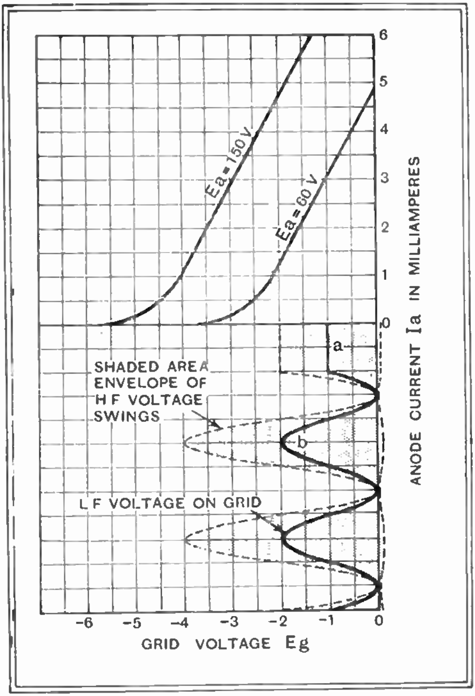

Fig. 105: Curves showing reason for overload of grid detector. Even though the rectified voltage *b* on the grid may not overrun the straight part of the characteristic, the R.F. voltage accompanying it may cause anode-bend detection

In principle this is quite simple, but the two-fold function of the grid introduces a certain difficulty. If the triode were being used purely as an A.F. amplifier, the A.F. input to the grid could be allowed to swing over the full range within which the characteristic is reasonably straight. For example, in Fig. 106, with anode voltage 60, it would be quite safe to swing from 0 to −2 volts, as shown by the heavy line, causing the anode current to vary proportionately between 4·8 and

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1·2 mA. But we can see from Fig. 102*b* (and Fig. 103) that in order to cause an average voltage of −100 at the rectifier (the grid, in the case of a triode) it is necessary for the R.F. voltage to swing between 0 and −200. Similarly in Fig. 106 it is necessary for the R.F. to swing between 0 and −4, shown shaded. Such a swing over-runs the anode current characteristic and the negative peaks of grid voltage would fail to cause a *proportionate* reduction in anode current (for it cannot become less than zero). As the A.F. output is simply the average of the rectified R.F., its negative peaks also would be flattened off and distorted.

One answer is to keep the voltage of the applied R.F. signal down so low that it never drives the anode current, even momentarily, round the bend in the valve characteristic. This means that the A.F. output is restricted to one half of that which could be obtained if the R.F. were absent and the valve were acting purely as an A.F. amplifier.

110. Disadvantages of the Grid Detector

We shall see presently that restricting the input to the detector to not more than a volt or two tends to introduce another form of distortion, and also limits the usefulness of the detector for auxiliary duties (Chapter 18). To escape this dilemma, one method is to raise the anode voltage of the grid detector, as suggested by the 150-volt curve in Fig. 106. It can be seen that such an adjustment would accommodate the R.F. right up to the peaks of modulation, and so preserve the A.F. outline from distortion. This policy was at one time fairly popular under the name of *power grid detection*, but it may be noticed that to handle even such a modest A.F. as 1 volt peak the anode current runs up to an alarming figure, especially at times when there is little or no carrier wave.

The apparent economy of Fig. 105*b* has, therefore, proved illusory, especially as in many modern receivers it is desired to have a rectified voltage of 20 or more. So the diode has returned to favour, because there is no upper limit to the R.F. voltage that can be applied and satisfactorily rectified, and the R.F., when it has served its purpose, can be removed before the derived A.F. is passed on to the next stage.

111. Elimination of R.F.

There are two ways of doing this. One of them is to take the output from across the condenser as in Fig. 104 *c* and *d*; because, as already pointed out, the condenser prevents the

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voltage across it from changing very much at such a high frequency as that of the carrier wave. The other, which enables Fig. 104 *a* or *b* to be employed, is to interpose a simple filter, such as that shown in Fig. 107, in which L is a R.F. choke coil and C a condenser so chosen that L has a much

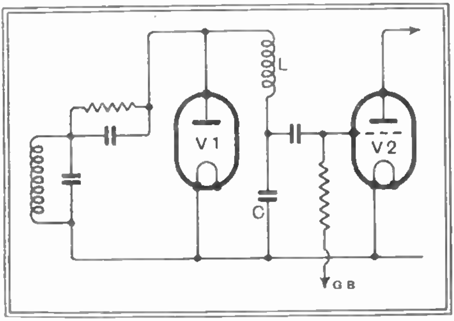

Fig. 107: Detection by diode V_1 followed by pure A.F. amplification by the triode V_2

higher, and C a much lower, reactance to R.F. than to A.F. To rid the Fig. 104 *c* or *d* outputs of the last traces of R.F. such a filter is sometimes used with them too, but as the requirements are not so stringent it is common for L to be replaced by a resistor of 10,000 to 50,000 ohms.

112. Elimination of D.C.

If the explanation of Fig. 103*c* has been carefully followed it will be evident that in addition to the R.F., represented by the shading (which has now, we hope, been eradicated), and the A.F., represented by the waviness of the heavy line, there is also a negative D.C. component, represented by the lowering of the mean level of the heavy line below the zero mark by an amount approximately equal to the peak R.F. voltage. It is only the A.F. that we want to amplify; so to cut out the D.C. a blocking condenser is connected between the diode and the grid of the amplifier, as shown in Fig. 107.

To enable this grid to be set to the most suitable bias for amplification (Sec. 73) it is connected to a source of bias voltage through a resistance, which must be kept high—generally about 1 M Ω—to avoid loading the detector unduly or introducing a particular form of distortion to be described shortly.

113. The Anode Bend Detector

As several causes of distortion have already been mentioned, it is about time to give some details of them. The first can be illustrated by dealing with another type of detector. We have seen that the grid detector is really a diode rectifier, formed by grid and cathode working at about zero volts, where grid current starts, followed by a triode amplifier working on a straight ("linear") part of its characteristic. An alternative

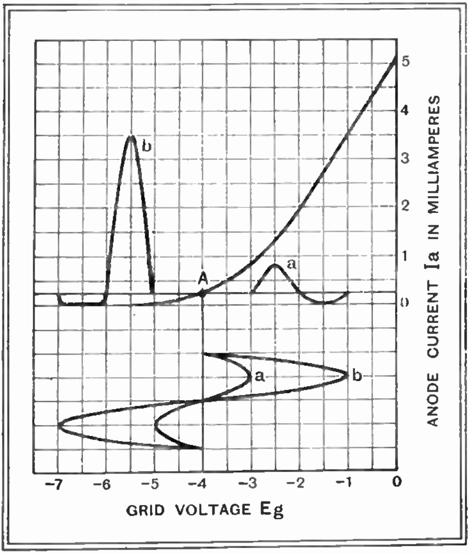

Fig. 108 : Illustrating rectification of both weak and strong signals by an anode-bend detector

is to bias the grid negative so that grid current never occurs and to use the anode characteristic as the rectifier, which can be done by biasing it to a point at which anode current is reduced nearly to zero.

And here it is that we depart from the ideal and semi-ideal rectifiers of Fig. 95. Their characteristics, though they might be imperfect in not giving zero forward resistance and infinite backward resistance, at least change over abruptly from one to the other. But real rectifiers change more or less gradually, the characteristics having a curved portion as shown in Fig. 108

which represents a typical triode. For amplification, such a valve would be worked on the most linear part of the characteristic by means of biasing it about $-1\frac{1}{2}$ volts. The problem now is to find the most suitable bias for detection. Anode current appears to start at a bias of -5 volts, so this might be assumed to be a suitable working point.

But if the scale of anode current were expanded it would be found that a small current is flowing, so that if a small signal voltage were applied, swinging the grid between, say, $-5{\cdot}5$ and $-4{\cdot}5$, the " forward " current (grid going less negative) would be small and the " backward " current would not be zero ; in other words the rectification efficiency would be low. A point such as A is more likely to be chosen, for although the backward current is appreciable its effect is outweighed by the much greater forward current.

It can easily be seen, however, that detection of very small signals is very poor indeed, for any small section of the characteristic curve shows little difference between the current changes produced by the positive and negative halves of the signal cycle. Even the signal marked *a* in the lower portion of Fig. 108, representing 1 volt amplitude in each direction on the grid, gives only a very partial rectification, as shown by the current curve *a* ; whereas the 3-volt signal *b* gives a rectified current much more than 3 times as great. The rectified current, in other words, is not proportional to the amplitude of the carrier wave.

114. Effects of Curved Characteristics

The effects of this are two-fold : firstly, a receiver employing a detector of the nature described is insensitive to weak signals : secondly, the modulation of even fairly strong signals is distorted. In Fig. 97 we plotted the rectified output against the R.F. input voltage, and for the linear detectors whose characteristics are shown in Fig. 76 the results are also linear. But we have just seen that in the Fig. 108 type of detector, known for obvious reasons as the *anode bend* detector, the rectified output increases more rapidly than the applied R.F., and so we would get a rising curve.

The process is shown in greater detail in Fig. 109, which repeats the valve characteristic of Fig. 108, but instead of only a single cycle of R.F., the shading represents too many R.F. cycles to be shown individually, modulated at A.F. The unmodulated carrier wave, a short section of which is given at the top, has an amplitude of 2 volts, and swings the anode current (represented on the right) between 0 and 2 mA. around

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the initial current, which is 0·25 mA. The balance in favour of the positive half-cycles is, therefore, 1·5 mA. (peak values), and this represents the rectified output, and is indicated by a heavy line.

When modulated 25 per cent. the rectified output (positive swings less negative swings) gives a reasonably faithful reproduction of the modulation shape. But when the modulation depth is increased to 75 per cent. the rectified output increases

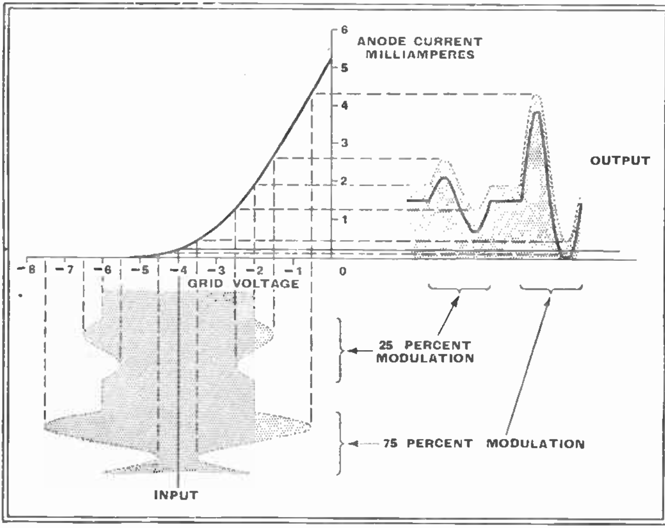

Fig. 109: Extension of Fig. 108 to show application of R.F. carrier wave (modulated to differing depths). The full line represents the modulation-frequency rectified output, and can be seen to be not proportional to the original modulation of the input

considerably more for the positive peaks of modulation than it decreases for the negative. Note that at the latter the R.F. output is nearly equal in the positive and negative directions (above and below 0·25 mA) and the rectified output is, therefore, practically zero. Still greater modulation would, therefore, considerably increase the A.F. positive peaks, but could not reduce the negative which are already zero, so the distortion would be still more noticeable, and would result in falsification of the reproduced sound.

It is fairly easy to see that distortion occurs at a less modulation depth with small carrier wave amplitudes than with large

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ones. If an attempt is made to obtain distortionless reproduction up to fairly large modulation depths by increasing the carrier amplitude, it can be seen from Fig. 108 that there is a danger of running into grid current at the positive peaks of modulation, which would introduce still more serious distortion. Added to these drawbacks is another, described in Sec. 120; and so the anode bend detector is now seldom used except for special purposes.

115. A.C. Loading Distortion

The diode characteristics are not free from starting curvature, but are generally a considerable improvement on the anode bend controlled by grid voltage. Moreover, there is no restriction on increasing the carrier amplitude, and by doing

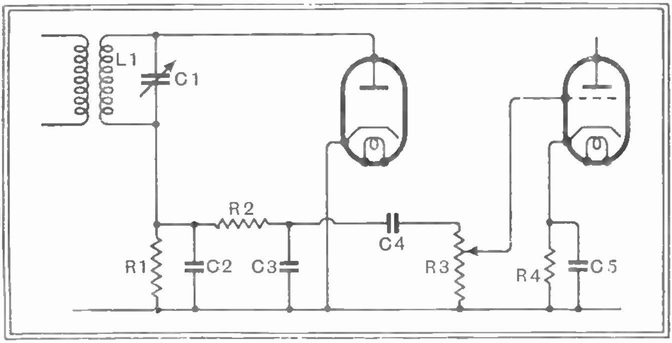

Fig. 110: Typical diode detector circuit followed by A.F. amplifier, showing that the A.C. load resistance is lower than the D.C., due to the path via C_4

so it is possible to reduce the distortion to negligible proportions even with 90 per cent. modulation (which is the deepest the transmitter itself is likely to be able to handle with tolerable distortion). As we have seen, this is not true of the grid detector, which causes distortion when the signal amplitude is either too small or too large.

The diode detector is, therefore, the most satisfactory, as correct operation of it is mainly a matter of arranging that the input is not too small. There is, however, one rather serious, but avoidable, source of distortion with the diode. Fig. 107 shows the blocking condenser which is used to prevent the D.C. component of the rectified output from reaching the amplifier valve. It is necessary to use a resistor on the far side

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of this condenser, for connecting to the appropriate grid bias source. This has also been found to be the most suitable point at which to introduce volume control, because distortion is kept down to a minimum by arranging for the signal to be as large as possible at the detector and as small as possible at the amplifier which follows it.

So the most usual circuit is as shown in Fig. 110, in which L_1C_1 are the final R.F. tuned circuit, R_1 the diode load resistance and C_2 its condenser, R_2 and C_3 a R.F. filter, C_4 the blocking condenser, and R_3 the volume control. The purpose of R_4 and C_5 is to provide bias for the amplifier valve, as explained in Sec. 209. Note that in this circuit the D.C. load of the

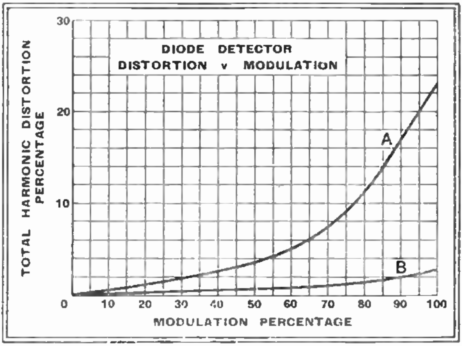

Fig. 111 : Showing how the harmonic distortion in a detector circuit where the A.C. load resistance is two-thirds of the D.C. (curve A) is much greater than if A.C. and D.C. load resistances are equal (curve B)

diode is R_1, whereas the A.C. load resistance is R_1 in parallel with $R_2 + R_3$—the reactance of C_4 being assumed negligible in comparison. The A.C. load resistance is, therefore, lower than the D.C., and in these circumstances it is found that deeply modulated signals are distorted. The smaller the ratio of A.C. load resistance to D.C., the lower the percentage modulation at which distortion sets in.

Fig. 111 (due to Langford Smith's "Radio Designer's Handbook") shows the great increase in distortion resulting from making the A.C. resistance one-third less than the D.C.; as, for example, making R_1 in Fig. 110 0·5 MΩ and $R_2 + R_3$ 1MΩ. This is, therefore, a point that must be carefully watched in the design, especially as other A.C. loads may be connected in parallel (see Chapter 18).

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CHAPTER II

THE SINGLE-VALVE RECEIVER: REACTION

116. The Circuit

BY now we have covered enough ground to be able to discuss the behaviour of a simple type of receiver. This will take us away, for the first time, from the fairway of simple theory, and we shall find ourselves making acquaintance with some of the incidental complications that arise when we have to deal with real circuits in place of circuits idealized to bring out their fundamental properties.

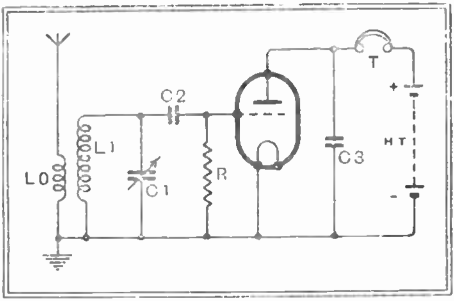

Fig. 112: Circuit of single-valve set in which a triode valve is used as grid detector and amplifier of the detected signals

Fig. 112 shows the circuit of a single-valve receiver. The outline of its working is simple enough. The currents induced in the aerial by the received wave flow through the primary winding L_0, to which is inductively coupled the secondary winding L_1, this being tuned to the frequency of the desired signal by adjustment of the variable condenser C_1. The signal voltage developed across the tuned circuit is applied, through the grid-condenser C_2, between grid and cathode of the triode valve which, since the resistance R (the *gridleak*) is returned to cathode, will behave as grid detector (Sec. 109). The detected and amplified signals in the anode circuit are passed through the phones T and so made audible to the listener.

The function of C_3 is the subject of Sec. 121. The letters "HT" against the anode supply battery stand for "high tension," which in the early days of valve technique was the name given to this battery to distinguish it from the filament battery or "LT." Although the name is undesirable, because

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in electrical engineering its use is definitely allocated to much higher voltages than those in receiving sets, it has gained too strong a hold to be quickly abolished.

117. The Radio-Frequency Transformer

We have already considered the function of a radio-frequency transformer coupling a valve transmitter to its aerial ; now we have one coupling the receiving aerial to the receiving valve. Although the same basic principles hold, the problems are not exactly the same in practice. In the transmitter, any power used up in the dynamic resistance of the tuning circuit is wasted, because it never reaches the aerial. The dynamic resistance should therefore be much greater than the load resistance

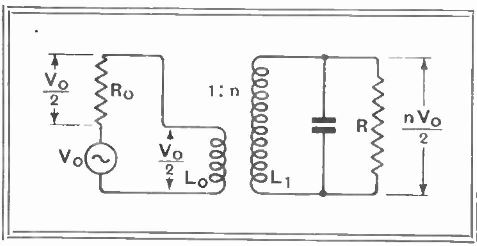

Fig. 113 : R represents the dynamic resistance of the tuned circuit, combined with the resistance of valve, grid leak, etc. Ro represents the resistance, at the working frequency, of the aerial. Maximum voltage across R is obtained when n is so chosen as to make the apparent resistance of the primary equal to that of the aerial. For this, $n=\sqrt{R/R_o}$, and the voltage distribution is as shown

consisting of the aerial combined with its coupling transformer, so that when a R.F. voltage exists across it nearly all the current will go into the load and very little into the dynamic resistance.

At the receiving end we have the aerial picking up a small signal voltage, and therefore acting as the " generator " ; while the tuning circuit, L_1, C_1 in Fig. 112, together with C_2, R, and the valve itself, is the load. Except at very high frequencies, when certain rather subtle effects cause the grid-to-cathode path of the valve to act as a fairly low resistance, the input resistance of the valve is quite high. R is generally very high, of the order of a megohm. Therefore the dynamic resistance of the tuning circuit may be the lowest of those concerned, thus absorbing the majority of the power ; so it cannot be neglected as a load. As C_2 (in Fig. 112) is generally large enough to have a relatively negligible reactance at radio

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frequency, the dynamic resistance, grid leak resistance, and valve itself are all effectively in parallel and can theoretically be lumped together as a single resistance R in Fig. 113, across a theoretical tuning circuit which is loss-free and therefore of infinite resistance. The problem is to couple the aerial, which may be represented as a generator giving a voltage V_0, in series with the aerial resistance, R_0, to the tuning circuit, in such a way as to produce the greatest R.F. voltage across it.

118. Effect of Primary Turns

The greatest voltage will also result in the greatest amount of power being delivered to R; and we have already seen

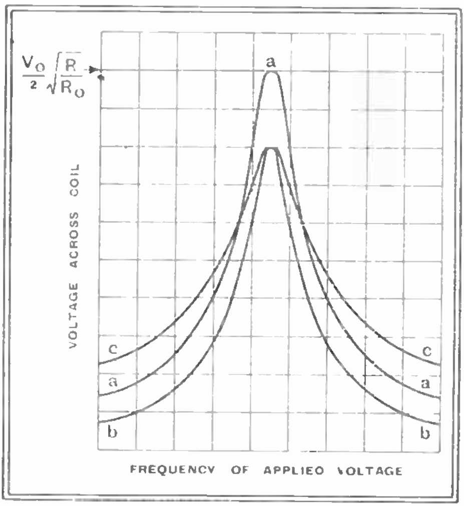

Fig. 114. : A typical resonance curve for optimum primary turns in the aerial coupling transformer is shown as a. A greater number of turns reduces voltage and selectivity (c) ; a smaller number of turns reduces voltage but improves selectivity (*b*)

Secs. 72 and 99) that this condition is obtained when the load esistance is made equal to the generator resistance. Applying ne same transformer rule as before, the turns ratio, *n*, giving ne greatest voltage across the tuning circuit, for applying to ne valve, is $\sqrt{R/R_0}$. For example, if R is 180,000 ohms and $_0$ is 1,800, *n* should be 10, and so L_0 would have one-tenth s many turns as L_1.

The secondary, L_1, being tuned, has to have the right number of turns to give it the inductance necessary to cover the waveband over which it is desired to tune; that leaves us with the primary turns as sole variable.

When this number of turns is adjusted to give the greatest voltage, half the generator voltage is lost in the aerial resistance and half appears across the primary. This half is stepped up to n times the voltage across the secondary, as shown in Fig. 113. At the same time the effect of the aerial resistance on the tuned circuit is equivalent to connecting another resistance equal to R in parallel with it, reducing the " Q " of the coil (see Sec. 52) to half what it would be with no aerial coupled. The selectivity is a half, too The voltage appearing across the tuned circuit as the frequency of the signal is varied would under these conditions be shown by such a curve as Fig. 114*a*.

If the number of primary turns is reduced below the optimum amount just calculated, the voltage across L_0 is stepped up in a greater ratio, across L_1, but this benefit is more than offset by the reduction in the voltage across L_0 because its effective resistance is greatly reduced by the increase in n and so most of the available voltage is lost in R_0. But the selectivity improves. See Fig. 114*b*. On the other hand, if the primary turns are increased in number, the voltage across them increases slightly, but can never quite reach V_0, while this increase is more than offset across L_1 by the reduced step-up. Moreover the effect of R_0 transferred to the secondary causes R to be shunted by a lower resistance, causing the tuned circuit to be more heavily " damped," or less selective, as shown by Fig. 114*c*.

In radio receivers, maximum signal voltage is not everything; it may be, and often is, worth sacrificing some in favour of selectivity. Increasing the number of primary turns above optimum, or *coupling tightly*, loses both signal voltage and selectivity, so has nothing in its favour; reducing it, or *coupling more loosely*, loses signal—not very much for a slight reduction—but gains selectivity, so is often adopted. Another advantage of loose coupling is that less aerial reactance is transferred to the tuning circuit, and so the frequency to which this circuit resonates is less affected by the nature of the aerial used—This is particularly important when several circuits have to be simultaneously tuned by condensers that are ganged (i.e. mounted on the same shaft).

Still another consideration is that when the capacitance is reduced in order to tune to a higher frequency, the dynamic resistance is increased, causing greater signal voltage, and the selectivity is reduced. Simultaneously, assuming the number

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of primary turns to be so small that in combination with the capacitance of the aerial it tunes to a higher frequency than any in the waveband covered by the tuning condenser, the aerial circuit is brought nearer resonance, and so the signal voltage is greater for this reason also. When tuning to the lower frequency end of the band, the reverse takes place, and signal strength is relatively poor. To obtain a more uniform signal strength and selectivity over the waveband, the primary coil is often given a large enough number of turns to make it " peak " (i.e., resonate) beyond the *low*-frequency end of the band, to offset the falling dynamic resistance at that end. To prevent this from causing tremendous damping, the primary and secondary coils are placed so that only a small number of magnetic lines of force link both ; thus the coupling is actually quite loose.

Lastly, an aerial, regarded as an impedance, is not just a simple fixed resistance, as we have conveniently assumed, nor are all aerials alike. The impedance of any one aerial varies greatly according to the frequency.

Taking all these matters into account, then, the practical design of the aerial coupling transformer is quite a complex problem.

119. Tuning Range

Reverting to the circuit of Fig. 112, we see that the total capacitance across L_1 is increased above that of the tuning condenser C_1 by the extra capacitances due to the valve and its holder, the wiring, the terminals or tags to which the ends of L_1 are brought, and by a certain amount of capacitance transferred from the aerial through the primary L_0. If the maximum capacitance of C_1 is the usual 500 *p*F, the total will be about 550 *p*F, from which we find (Sec. 55) that if we are to tune to 550 kc/s, L_1 must have an inductance of about 155 μH.

The highest frequency to which a circuit will tune is almost entirely a function of the extra, or " stray " capacitances, but in any average case a coil of 155 μH will just comfortably tune to 1,500 kc/s.

As we have seen (Sec. 107), usual values for C_2 and R are 200 *p*F and 0·5 M Ω. The effect of the grid circuit of the valve in damping the tuned circuit has also been mentioned ; it is approximately equal to putting across C_1 a resistance $\frac{1}{2}$R, or, in this case, 0·25 M Ω. If the dynamic resistance of the tuned circuit alone is 125,000 ohms it will be reduced by this damping to two-thirds of its original value. Alternatively

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expressed, the valve will increase the equivalent series resistance of the circuit by 50 per cent.

120. The Miller Effect

In addition to this effect, which is solely due to the grid current taken by the valve, there is another, named after its discoverer, Miller, which depends for its existence upon the voltages developed in the anode circuit, and upon the small capacitance between the anode and the grid of the valve. In Fig. 115 *a* there is shown the conventional diagram of the valve used as an amplifier, the impedance in the anode circuit being represented by Z_a. This may be a resistance, a capacitance, or an inductance. C_{ga} represents the total capacitance between grid and anode, which is partly in the valve electrodes themselves and the glass pinch supporting them, and partly in the valve base the valve holder and the wiring.

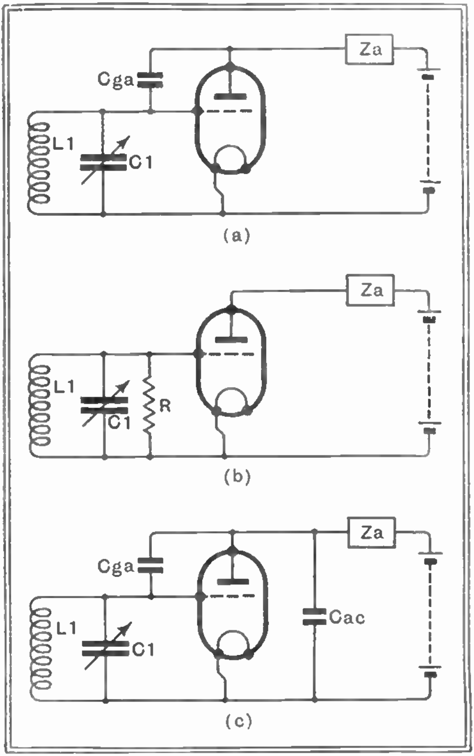

Fig. 115 : Illustrating the Miller Effect. *a* A voltage at the anode of a valve can pass, by way of C_{ga}, back into the grid circuit. *b* If Z_a is a capacitance, the anode-grid feed is equivalent to connecting a damping resistance R across the grid circuit. *c* Owing to the stray capacitances from anode to cathode, any anode-circuit impedance Z_a is necessarily shunted by C_{ac}

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Since the amplifying action of the valve produces a signal voltage at the anode, a small signal current will flow through C_{ga} and the grid circuit to the cathode of the valve. In flowing through the components in the grid circuit, this current will develop across them a voltage, and this voltage might have any phase relationship with the voltage already present due to the signal. If it were in phase with the original signal voltage the two would simply add, and the original voltage would be artificially increased. If it were 180° out of phase, on the other hand, this new voltage would be in opposition to that already there, and the energy fed through C_{ga} would tend to damp out and reduce the signal voltage. In a third case the voltage fed back from the anode might be 90° out of phase with that already present,

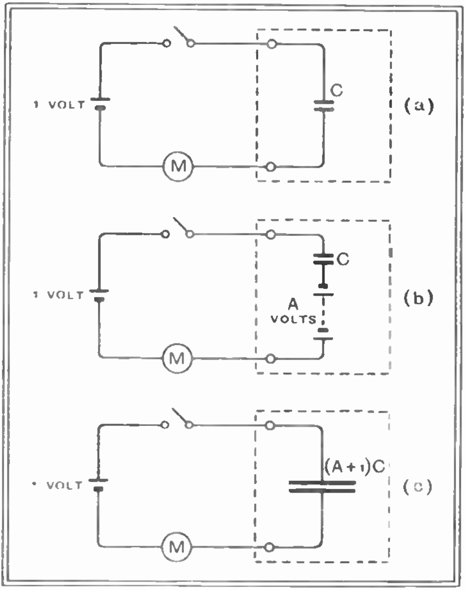

Fig. 116 : Showing how the introduction of an extra voltage, such as that fed back from anode to grid in a valve, is equivalent to a multiplied capacitance

and would therefore neither help nor hinder it, but would produce the same effect as a reactance.

In general there will be a combination of this with one of the other two.

Let us consider the case in which Z_a is a resistance, because we have already seen, in Sec. 71, that the signal voltage at the anode of a resistance-coupled amplifier is opposite in polarity to that applied to the grid, and A times as great (" A " standing for the stage amplification). Therefore every signal volt applied to the grid causes $A + 1$ volts between anode and grid. Consider Fig. 116 *a*, in which C is a condenser hidden in a box, and M is a special sort of meter that measures the charge passing into the condenser when the switch is closed. As the capacitance of a condenser is equal to the charge required to raise the P.D. across it to 1 volt (Sec. 19), we have here a means of measuring the capacitance of C. But suppose, *b*, that a demon in the box connects an A-volt battery in series with C whenever we connect our 1-volt battery. The charging voltage is A + 1 volts, and the meter indicates a charge A + 1 times as large as it would without the demoniacal activity. Knowing nothing of this, we conclude that the box contains a condenser having a capacitance A + 1 times as large as C. So far as external electrical tests can tell, the boxes *b* and *c* are identical.

Thus the effect of the grid-to-anode capacitance C_{ga} is to make our resistance-coupled amplifier behave as if a capacitance A + 1 times as large as C_{ga} were connected across the tuned circuit L_1C_1. As A might be 50 or more, the result is serious.

If Z_a is a capacitance, the energy fed back tends to damp out the voltage already present, while if it is an inductance, the energy fed back reinforces and increases the signal voltage on the grid, perhaps sufficiently to set up continuous oscillation, as in the T.A.T.G. oscillator (Fig. 78), which depends on this principle.

In the case where Z_a is a capacitance, the damping effect on the grid circuit can be exactly reproduced by connecting a resistance R of suitably chosen value across grid and cathode of the valve in the manner shown in Fig. 115 *b*. But a little thought will make it clear that since the whole effect depends on the alternating voltage at the anode, changes in magnitude of this will alter the value of the equivalent damping resistance R. The higher the impedance of Z_a (or since we are considering the case where this is a capacitance, the lower the value of this capacitance) the higher will be the voltage developed, and hence the greater will be the damping effect in the grid circuit. Thus a high value of Z_a corresponds to a low value of R, reducing very markedly the voltage across the tuned circuit of Fig. 112, and flattening its tuning to a considerable extent.

If the capacitive reactance of Z_a is made high (corresponding to a small value for C_3 in Fig. 112) the damping can be very serious indeed ; with C_3 omitted altogether, so that the anode-circuit impedance for radio-frequency currents consists only of the stray capacitances across valve, valve-holder, and telephones (Fig. 115 *c*), the energy fed back from anode to grid may be equivalent, for a signal at 1,000 kc/s, to connecting a resistance of as low a value as 5,000 Ω between grid and cathode. Since

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the dynamic resistance of the tuned circuit L_1C_1 will probably be twenty times as great as this, the effect of the damping in dropping signal strength and flattening tuning is positively catastrophic.

121. The Anode By-pass

This explains the presence of C_3 in Fig. 112 ; it is inserted as a low impedance to the radio-frequency currents so that the voltage developed at the anode may be as low as possible.

Evidently there is a limit to the reduction in damping that can be effected in this way, because although from the radio-frequency point of view the higher the capacitance of C_3 the less the voltage developed across it and the less the damping thrown back into the grid circuit, we have to remember that there are audio-frequency voltages that we want. Fortunately the reactance of C_3 is much greater at audio frequencies, so they are not suppressed so effectively as the radio frequencies. If we make C_3, about 0·001 μF, its reactance at 1,000 kc/s will be little more than 150 ohms, while at the higher audio-frequencies (5,000 cycles) it will rise to 30,000 ohms, which will not be a very serious shunt to the phones, and so will not cause too great a diversion of high notes from their windings. Like almost every other point in a wireless set, the choice of a capacitance for C_3 is a compromise that tries to make the best of both worlds.

122. Reaction

Instead of striving to prevent feed-back from the anode to

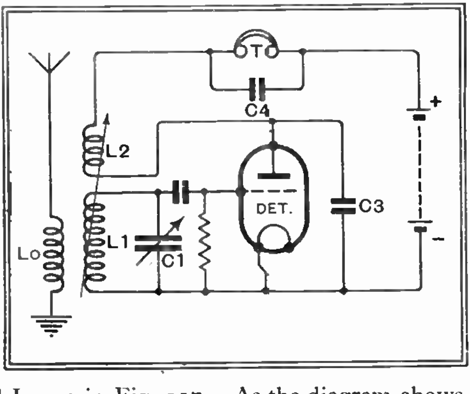

Fig. 117 : Conventional single-valve set with adjustable reaction

the grid circuit, we can deliberately introduce it, so arranging matters that we have it at all times completely under control. This can be done by inserting in the anode circuit a coil L_2, as in Fig. 117. As the diagram shows, this coil is close to L_1, in order to couple with it inductively,

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while the arrow running through them indicates that their relative positions can be adjusted as required.

Part of the radio-frequency current flowing in the anode circuit will pass direct to cathode through C_3, and part will flow through L_2, the capacitance C_4 across the phones, and the anode battery. This latter portion, in passing through the coil, sets up round L_2 a radio-frequency field which, in passing also through L_1, induces a voltage in the latter. By connecting L_2 in the right direction this voltage can be made either to assist or to oppose the voltage there already, the effect in either case becoming more marked as L_2 is brought closer to L_1. We now have our old friend, the reaction-coil oscillator circuit of Fig. 74, described in Sec. 80. The fact that the grid circuit is tuned instead of the anode is a minor variation. The only essential difference is that now we are coupling too loosely to maintain oscillation.

We saw how tuned circuit damping, or resistance, which causes oscillation to die out, can be completely neutralized by a sufficiently closely-coupled reaction coil, so that oscillation continues indefinitely and without any external source. That would not suit our purpose in a receiver, because we want the strength of oscillation in the tuned circuit to depend from moment to moment on what is coming in from the distant transmitter. What happens when the reaction coil is coupled in the direction that tends to cause continuous oscillation, but too loosely to do so, is that oscillations set up by the signals derived from the aerial or elsewhere build up more strongly (because they are reinforced by the induced voltage from the anode) and die away less quickly, just as if the series resistance of the tuned circuit were less, or (what is the same thing) its dynamic resistance greater.

Of course, reaction does not neutralize resistance in any strictly literal physical sense. The sole characteristic of resistance is its absorption of power; if, therefore, we supply power from the anode circuit of a valve, the circuit in which that resistance is located behaves as though it had lost some of its resistance. The valve is used as a source because it is only by making the voltage itself (in the grid circuit) control the power used to enhance it that the two can be locked unalterably together in the required phase.

In discussing tuned circuits (Sec. 54) we saw that reduction of radio-frequency resistance increases both the magnification and the selectivity of a tuned circuit. With the aid of a valve to provide reaction we are now in a position to adjust the resistance of the tuned circuit L_1C_1 to any value that takes ou fancy, simply by approaching L_2 cautiously towards L_1 until the resistance has been reduced to the desired extent. As we do this the voltage developed by the signal across L_1C_1 will steadily rise and the tuning will become steadily sharper.

The effect on the tuned circuit can best be visualized with the aid of a series of resonance curves. In Figs. 118 and 119 the voltage across L_1C_1 is plotted against frequency for a

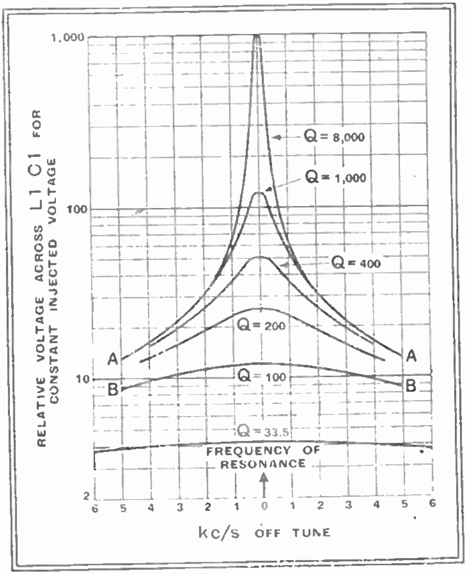

Fig. 118 : Showing relative voltage at resonance (height of peak) developed across L_1C_1 for various coil magnifications Q

number of values of magnification. A glance will show that as the magnification is increased by the application of reaction the signal-voltage rapidly rises* and the sharpness of tuning, as measured by the ratio of the voltage at resonance to that developed a few kilocycles off tune, becomes greater.

So great is the increase that it could not be shown on a curve sheet like those in Chapter 6 : a logarithmic scale of

* The relative heights of the peaks are calculated on the basis of constant injected voltage. This ignores the reaction of L_1C_1 upon L_0, the aerial primary.

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relative voltage is necessary, and compared with Fig. 53 the rise is greater than it looks.

The difference between the two sets of curves in Figs. 118 and 119 is purely one of frequency scale; in the former the frequency scale extends to 6 kc/s on either side of resonance, so that only the peaks of the curves are plotted. In the latter

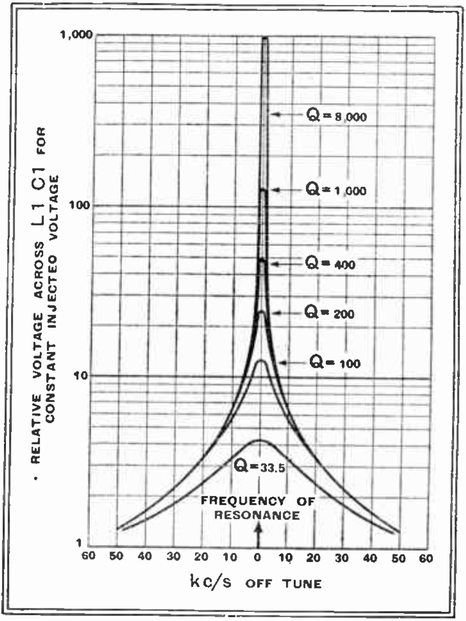

Fig. 119: Extension of Fig 118, showing the voltages across L_1C_1 when tuned exactly (peak) and when detuned to various extents

the behaviour of the circuits is shown over a range of 60 kc/s each way from resonance. In both cases the lowest curve is a fair representation of the behaviour of a tuned circuit of normal radio-frequency resistance connected to a detector. With the reaction coil out of use the circuit assumed has $L = 155\mu H$, $r = 10\ \Omega$, and is supposed to be tuned to 1,000 kc/s. Detector damping across it is taken as 50,000 Ω. For the tuned circuit alone $Q = 98$, $R = 95{,}000\ \Omega$; with detector damping in parallel the total dynamic resistance is reduced to 32,600 Ω, making the equivalent R.F. resistance 29·2 Ω and reducing the effective magnification to 33·5. The curve next in order ($Q = 100$) represents the same tuned circuit with the effects of detector damping almost exactly offset by the judicious application of reaction. Successive curves show the effect of more and more reaction, culminating in the extreme case where the magnification has been increased to 8,000, which is about the highest value known to have been reached, and held, by this means. It corresponds to the neutralization of all natural resistance of the circuit except for a small residue of about one-eighth of an ohm.

123. Over-Sharp Tuning

At first sight it would appear that the reduction of circuit resistance, even to such very low limits as this, was all to the good, since it would increase both the sensitivity and the selectivity of the receiver. If we had to receive only a simple carrier wave this conclusion would be true, but we must remember that the signal from a broadcasting station consists of a *modulated* carrier. As we have seen, the modulation consists of a variation in the amplitude of the carrier at the frequency of the musical note it is desired to transmit. We know also (Sec. 76) that if a tuned circuit had no resistance at all, any oscillation that might be set up in it would persist after the removal of the signal source, unchanged in amplitude, for ever. Such a circuit would evidently be quite incapable of following the rapid variations in amplitude of a modulated carrier.

It follows, therefore, that as we approach towards zero resistance by a greater and greater application of reaction, the voltage across the tuned circuit will tend more and more to "ring", following with greater and greater sluggishness the variations due to the modulation. For the highest audible notes the radio-frequency voltage has to change in amplitude most rapidly; as the resistance of the tuned circuit is decreased these will therefore become weak and vanish at a value of resistance still high enough to enable the low notes, for which the variations in amplitude of the carrier are proportionately slower, to remain substantially unaffected.

The high, sharp peak of a very low-resistance circuit such as that giving the curves "$Q = 8{,}000$", therefore, tells us that high modulation-frequencies cannot be followed. On the other hand, the flatter curves such as that for $Q = 100$, indicate a resistance high enough for any current through the circuit to die away rapidly unless maintained by a driving voltage, thus enabling the voltage-variations across L_1C_1 to be a faithfu copy of the signal as received from the aerial.

124. The Theory of Sidebands

By regarding the modulated wave from a slightly differen point of view, the relationship between sharpness of tuning and the loss of high audible notes can be shown to be very much more intimate than has been suggested. Strictly speaking, it is only an exactly recurrent phenomenon that can be said to possess a definite frequency. The continuous change in amplitude of the carrier wave in response to modulation makes the radio-frequency cycle of the modulated wave non-recurrent, so that in acquiring its amplitude variations it has lost its constancy of frequency.

A mathematical analysis shows that if a carrier of f_1 cycles per second is modulated at a frequency f_2 cycles per second the resulting modulated wave is exactly equivalent to three separate waves of frequencies f_1, $(f_1 - f_2)$, and $(f_1 + f_2)$. It is not easy to perform the analysis of the modulated wave into its three components by a graphical process, but the corresponding synthesis, adding together three separate waves, requires nothing more than rather extensive patience.

Fig. 120 shows at *a*, *b*, and *c* three separate sine-waves, there being 25, 30, and 35 complete cycles, respectively, in the length of the diagram. By adding the heights of these curves point by point, the composite curve at *d* is obtained. There are in its length 30 peaks of varying amplitude, and the amplitude rises and falls five times in the period of time represented on the figure. If this is a thousandth part of a second, curve *d* represents what we have come to know as a 30 kc/s carrier modulated at 5,000 c/s.

Thus a carrier modulated at a single audio-frequency is equivalent to three simultaneous signals, the unmodulated carrier itself and two associated steady frequencies *spaced away from the carrier on either side by the frequency of modulation.* In a musical programme, in which a number of modulation frequencies are simultaneously present, the carrier is surrounded by a whole family of extra frequencies. Those representing the lowest musical notes are close to the carrier on either side, those bringing the middle notes are farther out, and the highest notes are the farthest removed from the carrier frequency. The spectrum of associated frequencies on either side of the carrier is called a *sideband*, and as a result of the presence of these a musical programme, nominally transmitted on a (carrier)

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frequency of 1,000 kc/s, will spread over a band of frequencies extending from about 993 to 1,007 kc/s.

We now have a direct relationship between the selectivity of a tuned circuit and its ability to receive the highest notes likely

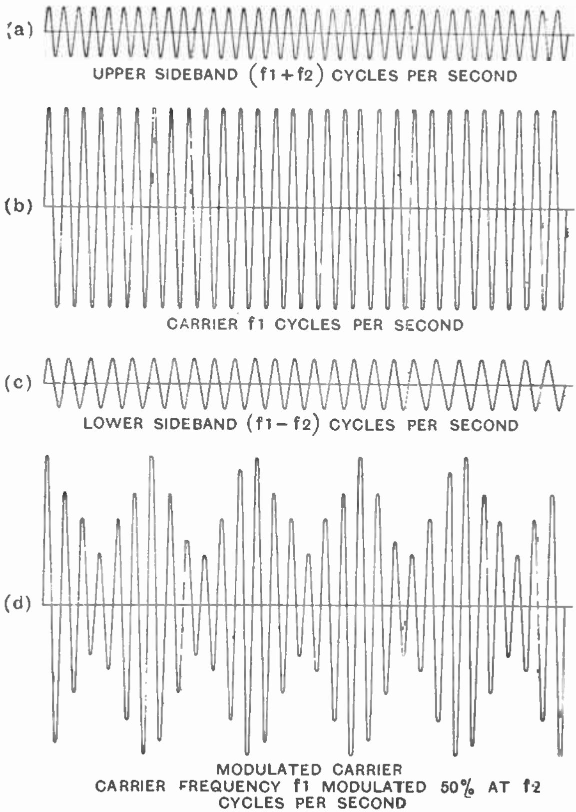

Fig. 120: Showing the relationship of a modulated carrier *d* to its three components

to be present as modulation on the carrier. If the resonance curve of the circuit is not substantially flat over a central portion wide enough to include the whole of the required sidebands, high notes will be attenuated—they will be quite literally tuned out owing to over-selectivity. In the curve for $Q = 8{,}000$,

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in Fig. 118, the sidebands corresponding to a modulation frequency of 5,000 cycles are shown, at points AA, as being transmitted at about 1·3 per cent. of the central carrier frequency. Lower notes are more fully transmitted, higher notes even more greatly attenuated. The result will be "woolly" and more or less unintelligible speech, and "boomy" music. For a tuned circuit in which $Q = 100$, however, 5,000-cycle notes are passed at 70 per cent. of the carrier amplitude (BB in Fig. 118).

It is clear from these considerations that high selectivity is not altogether an unmixed blessing in the reception of telephony, and that too great an application of reaction will sharpen tuning to such a point that the quality of the received programme suffers badly. Nevertheless it remains invaluable for neutralizing the losses due to detector damping, and may, without serious detriment to quality, be pressed far enough to halve or even quarter the natural resistance of a tuned circuit. But much greater amplification than this is needed for the successful reception of distant transmitters.

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CHAPTER 12

RADIO-FREQUENCY AMPLIFICATION: SCREENED VALVES

125. Increasing Range

IF we want to increase the range of our single-valve set sufficiently to enable us to receive transmissions from distant stations, the only alternative to reaction is amplification by a valve. A valve may be used in either of two ways: it may be applied to amplify the modulated radio-frequency signal before detection (radio-frequency amplification) or it may be made to amplify the detected audio-frequency signal (audio-frequency amplification). The choice between these two alternative methods is dictated by the characteristics of the detector.

We know that a large signal can be detected with less distortion than a small one; it is also true that any detector is very insensitive to really weak signals (Sec. 114). Unamplified signals from a distant station (a millivolt or less) would swing the grid of a detector over a portion of its curve so small that it would be virtually a straight line over that tiny range. We are driven, therefore, to amplify weak signals before detection in order to provide sufficient input to operate the detector satisfactorily.

126. Simple Resistance Coupling

At first sight it might seem that, since a resistance behaves alike to currents of all frequencies, one would obtain very satisfactory results by coupling valves together for radio-frequency amplification in the manner suggested in Fig. 121 Here is shown a stage

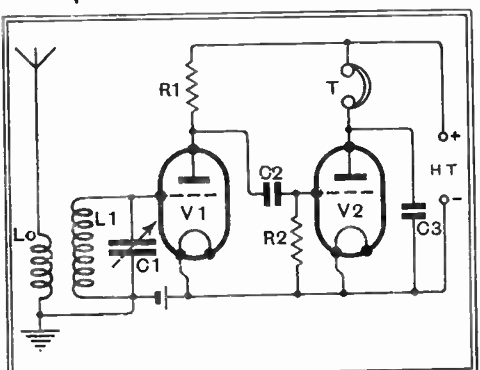

Fig. 121: The resistance-coupled R.F. amplifier is of little more than academic interest

of resistance-coupled amplification preceding the detector valve V_2. The valve V_1, receiving its signal from the secondary of the aerial transformer L_0L_1, produces an amplified voltage across the resistance R_1 in its anode circuit. This voltage is conveyed to the grid of V_2 through the condenser C_2 which, while readily passing R.F. voltages, protects the grid of V_2 from the steady positive voltage at the anode of V_1.

If one could build this receiver without departing from the strict letter of the circuit diagram it would work very well. Unfortunately, there appear in a practical set the stray capacitances from anode to cathode of V_1, and from grid to cathode of V_2, together with wiring, etc. They make, in an average case, a total of 40*p*F or more, which provides at 1,000 kc/s a path of reactance about 4,000 Ω. This sets an upper limit, irrespective of the value adopted for R_1, to the anode-circuit load of V_1. With so low a load V_1 will not provide very high amplification; one may expect a gain of about five times with an average valve.

But even this is not the worst fault of the circuit of Fig. 121. The anode circuit of V_1 being, as we have seen, predominantly capacitive, it damps the tuned circuit L_1C_1 rather heavily (Sec. 120). In the detector we reduced this damping to reasonable limits by inserting a condenser direct from anode to earth in an attempt to reduce the radio-frequency voltage at the anode as nearly as possible to zero; we were then wanting only the rectified audio-frequency signals. In the present case we obviously cannot do this, or we shall short out the signals, and, in consequence of the development of an appreciable radio-frequency voltage at its anode, V_1 is equivalent to a damping resistance of the order of 6,000 Ω across the tuned circuit. If the initial dynamic resistance of this, undamped, were 120,000 Ω, the introduction of this damping would reduce the voltage across it to less than one-twentieth.

With V_1 amplifying this reduced signal five times, the voltage finally delivered to V_2 would be one-quarter of that developed across L_1C_1 unloaded. On the whole, not a very successful amplifier.

The replacement of R_1 by a radio-frequency choke, making a choke-coupled amplifier, leaves the problem untouched; the faults of the circuit lie in the stray capacitances across the anode load and in the anode-grid capacitance of the valve, and not in the type of coupling used.

127. The Tuned Anode Circuit

But if we can find a method of neutralizing the effects of stray capacitances we shall be in a better position. Such a method

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lies ready to hand; we have only to place in parallel with them (i.e., from anode to earth or to the H.T. line) a coil of reactance equal to that of the stray capacitance, thereby forming a tuned rejector circuit. To avoid the awkwardness of having to readjust the value of this inductance every time we want to tune from one wave-length to another, we add a variable condenser for tuning. This gives us the *tuned anode* circuit of Fig. 122.

The diagram shows that LC is connected, as a complete circuit, between the anode of the valve and its battery. The stray capacitance in parallel with this now has no more effect than to make it necessary to reduce C itself a little below the value at which tuning would be attained in the absence of the strays. The whole forms a simple parallel tuned circuit. For the frequency of resonance we have seen that this behaves as a pure resistance R, the dynamic resistance L/Cr of the circuit. We have therefore worked our way back, so far as the electrical behaviour of the system is concerned, to the unrealizable resistance-coupled arrangement of Fig. 121. The amplification given by a tuned anode stage will be that calculated from the

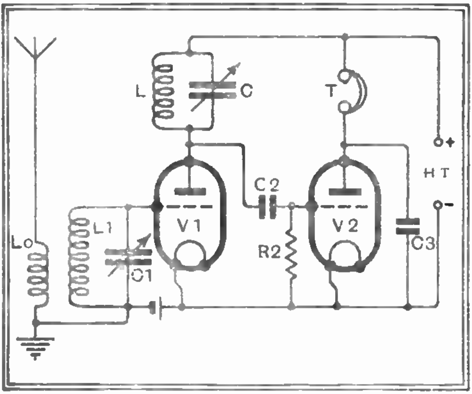

Fig 122: Tuned anode R.F. coupling. Compare with Fig. 121 and note that the various stray capacitances are now in parallel with C and so form part of the tuning capacitance

simple formula $A = \frac{\mu R}{R + r_a}$ given in Sec. 70 for a resistance-coupled stage, but we must now interpret R as the dynamic resistance of the tuned circuit.

We have found a remedy for the effects of stray capacitance in limiting amplification, for the circuit of Fig. 122 will give a gain of some 25 or 60 times with battery or mains valves respectively, even if R is no more than 100,000 ohms. It remains to be seen whether the anode-grid capacitance is equally harmless.

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128. Grid-Anode Capacitance

So long as the anode circuit is exactly tuned to the frequency of the signal being received, the anode circuit of the valve will be purely resistive, and voltage fed back through C_{ga} (Fig. 123) will neither assist nor damp down the voltage on the grid but will make it necessary to use less of C_1 in order to compensate for Miller effect. If the applied frequency (or alternatively the capacitance of C) is now increased, slightly more current will flow through C than through L, so that the anode circuit becomes capacitive. The fed-back voltage will then, as we have seen, tend to damp out the signal.

If, on the other hand, the applied frequency (or alternatively

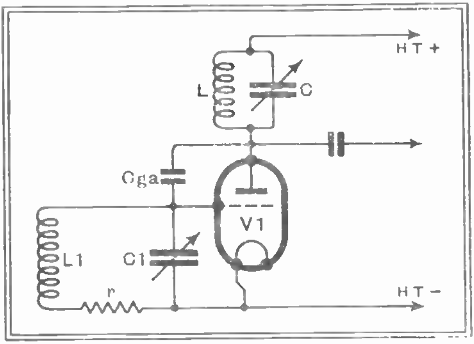

Fig. 123: The grid-anode capacitance of V_1 introduces difficulties into the working of the tuned-anode circuit

the capacitance of C) is reduced, more current will flow through L than through C, giving us an *inductive* anode circuit. Now the coupling provided by C_{ga} between the two tuned circuits will feed back energy that assists and builds up the voltage already present. We have, in fact, the condition described in Sec. 120, where it was pointed out that if the anode circuit is inductive, and the anode-to-grid capacitance feeds back sufficient voltage to the grid, the conditions for a T.A.T.G. oscillator are fulfilled. Comparing Fig. 123 with Fig. 78 (a T.A.T.G. oscillator circuit) it can be seen that essentially they are identical.

129. Instability

Our amplifier circuit therefore turns out to be an oscillator circuit. That does not necessarily mean that it sets up continuous oscillation. It will not do so unless the voltage fed

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back via C_{ga} is at least as great as that causing the anode voltage. In an amplifier of the type considered, the risk is so great as to be a certainty. The better the system is as an amplifier, the more likely it is to oscillate. Thus, if the stage gain is 50 times, it is necessary for only one-fiftieth of the anode voltage to find its way back to the grid in order to cause oscillation.

With coils of fairly good design (low r, or high dynamic resistance) and any ordinary triode, oscillation appears every time an attempt is made to bring L_1C_1 and LC into resonance with the same frequency. Although theoretically there should be no tendency to oscillation when exactly tuned, it is found that the increasing loudness of signals due to the commencement of feedback as C is reduced below the value necessary for resonance completely overwhelms the decrease of loudness that one would expect to find on detuning. In tuning the set there is therefore no aural indication of the true resonance point, so that in trying to tune for loudest signals one is led, every time, straight into the trap of oscillation, which occurs as soon as C is set a fraction low in capacitance.

In a receiver, oscillation results in the production of a rushing noise, and in the development of sundry whistles and squeaks as the set is tuned. For practical purposes, therefore, the circuit of Fig. 123 is unusable.

When the triode was the only valve available, oscillation due to feed-back through C_{ga} was avoided by providing a "faked" circuit by means of which another voltage, equal in magnitude but opposite in phase to that causing oscillation, could be fed back to the grid of the valve. These arrangements were known as *neutralized* circuits. They have now died out entirely, the modern solution to the problem of preventing feed-back through the grid-anode capacitance of the valve lying in the choice of a valve in which, by internal screening, this capacitance has been practically eliminated.

130. The Theory of Screening

The capacitance between any two objects can be reduced to zero by placing between them as a screen an earthed metal sheet of sufficient size. The operation of such a screen can be understood by considering Fig. 124, which shows at *a* two plates A and B separated from one another by an air-space. There will be a capacitance between them, so that the radio-frequency generator V will drive a current round the circuit Earth—V—A—B—Z_1—Earth. Across Z_1, which is an impedance of some kind between B and earth, the current will develop a potential difference, and this P.D. will be the voltage appearing on B as a result of current passing through the capacitance AB.

At *b* a third plate S, larger than either of the two original plates, is inserted between them in such a way that no part of either plate can "see" any part of the other. We now have no direct capacitance between A and B, but we have instead two capacitances, AS and SB, in series. If an impedance Z_2 is connected between S and earth the current round the circuit Earth—V—A—S—Z_2—Earth will develop a P.D. across Z_2. Since Z_2 is also included in the right-hand circuit the P.D. across it will drive a current round the circuit Earth—Z_2—S—

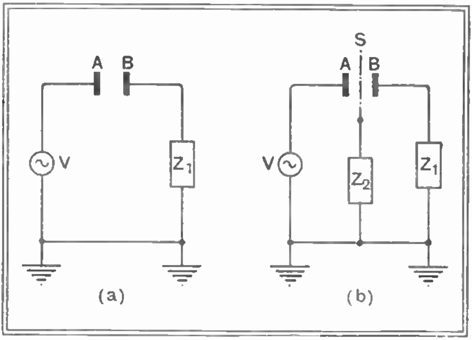

Fig. 124 : Illustrating the theory of screening

B—Z_1—Earth, and this will give rise to a potential on B. So far, S has not screened A from B, there remaining an effective capacitance between them which, if Z_2 is infinitely large, amounts to the capacitance equivalent to that of AS and SB in series. If S is thin this is practically equal to the original direct capacitance between the two plates.

Now imagine Z_2 to be short-circuited. Current will flow round the first circuit, but since there is now no impedance common to both there will be no driving voltage to produce a current in the latter. No matter what alternating voltages are applied to A, none will appear on B, even though large currents may flow via S to earth. The effective capacitance between A and B has therefore been reduced to zero, and B is completely screened from A.

It is very important to note that S is only effective as a screen if it entirely cuts off A from B, thus replacing the direct capacitance AB by AS and SB in series. Even with this proviso, perfect screening is not obtained unless S is definitely connected to earth either by a direct wire or through an impedance Z_2 which is negligibly small.

131. Screening a Valve

This is the principle used in reducing the grid-anode capacitance of a valve. A screen is put between anode and grid within the bulb, while capacitance between the leads running to grid and anode is avoided by taking the lead for one or other of these electrodes out through the top of the bulb. In some recent types of valve the screening has been extended down to and below the base, allowing both anode and grid to be brought out at the same end.

Clearly, a solid metal screen, while providing irreproachable screening, would cut off the electron flow from cathode to anode; it is therefore necessary to use as screen a close-mesh wire gauze through the openings of which electrons can pass. It is found that this necessary compromise with perfection still leaves a completeness of screening that falls short of that obtainable with an unbroken sheet of metal by a surprisingly small amount. In an unscreened valve, C_{ga} is usually of the order of 6 to 8 pF.; with a gauze screen, properly earthed, this is commonly reduced to less than 0·003 pF, and may even be less than 0·001 pF. The structure of a typical screened valve is shown in the sketch of Fig. 125.

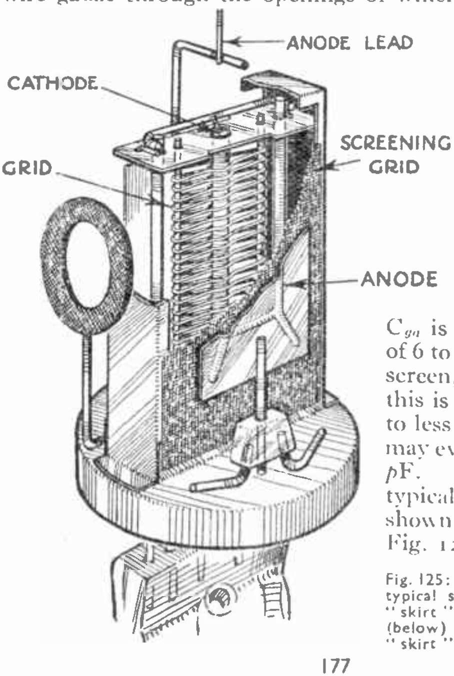

Fig. 125: Showing construction of a typical screened valve. Note the "skirt" screening the grid lead (below) from the anode. This "skirt" is connected to the screen

132. How a Screened Valve Works

If earthed in the strictly literal sense the potential of the screen would be approximately that of the cathode. Since the attraction of the positive anode cannot extend through the screen to any appreciable degree, electrons near the grid of the valve would then not be drawn onwards, and the anode current would fall practically to zero. But since, as Fig. 124 shows, the requirements of screening can be met by making Z_2 negligibly small, we can connect a condenser of large capacitance from the screen of the valve to earth, after which we can supply the screen, from any convenient source, with a positive potential.

The inner portion of the valve, comprising cathode, grid, and screen, is practically unaffected by the voltage at the anode ; so the total current through the valve is almost completely determined by the potentials of grid and screen. But if a electron arriving at the screen should happen to find itself exactly opposite to one of the openings in the latter, the attraction exerted upon it by the screen will come equally from all sides and it will go straight through the opening. With the anode at zero potential it would fall back again to the screen, but if the anode is much more positive than the screen it will be drawn on.

Thus by making the anode more positive than the screen some of the electrons, initially set in motion by the positive potential on the screen, will pass through the latter and travel on to the anode. The more the potential of the anode exceeds that of the screen the more electrons will be drawn on ; with rising anode voltage, therefore, the anode current rises and the screen current falls, the total remaining practically constant.

133. Characteristics of a Screened Valve

Curves of a typical screened tetrode (as this 4-electrode valve is called) are reproduced in Fig. 126, which shows anode current plotted against anode voltage. Each curve refers to the fixed grid-voltage E_g mentioned against it, and all were taken at a fixed screen-voltage of $E_s = 80$ V.. So long as E_a is considerably more than E_s, the anode takes practically all the current ; over the range $E_a = 120$ to $E_a = 200$ V. on the curve for $E_g = -2$, the anode current changes by only 0·08 mA. As E_a falls below 120 V the proportion of electrons pulled through the screen to the anode begins to drop, as the rapid fall in I_a shows. The screen current I_s, if plotted, would show a corresponding rise, keeping the total space-current constant.

The reasons for the peculiar shape of the curves for values

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of E_a lower than E_s will be discussed in connection with pentodes ; for the present it is enough to note that a screen-grid valve of this type is always used with an anode voltage considerably higher than that on the screen.

The extreme flatness of the curves over the working region to the right of the diagram indicates that the anode resistance of the valve is very high (Sec. 65). For the curve $E_g = -2$, the change of I_a by 0·08 mA for a change in E_a of 80 V indicates a resistance of 80/0·00008 = 1 megohm. But this value depends far more than in the case of the triode upon operating

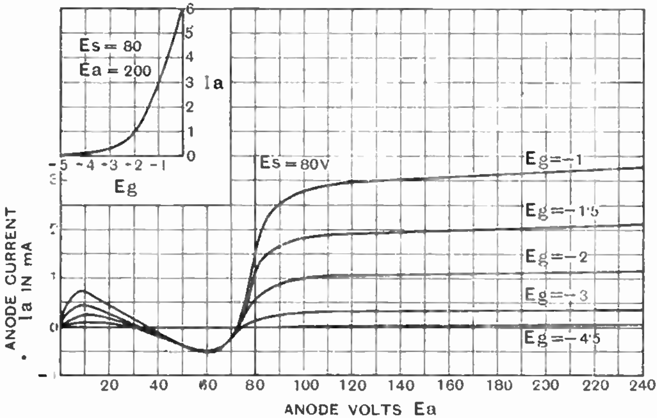

Fig. 126: Characteristic curves of typical screened tetrode. Only the flat parts of the curves to the right of the line E_s are used for amplification. Inset : I_a—E_g curve to show slope

voltages. Reducing the grid bias reduces also the anode resistance ; reading off values from the curve for $E_g = -1$ gives an anode resistance of 350,000 ohms only, which is about one-third of the value found for $E_g = -2$.

The small curve inset on Fig. 126, which shows the variation of anode current with grid voltage at $E_s = 80$ and $E_a = 200$, makes clear that this rapid variation of anode resistance is not accompanied by corresponding changes in mutual conductance or slope. At $E_g = -1$, $g_m = 2·45$, while at $E_g = -2$, $g_m = 1·45$ mA/V. Since the amplification factor of the valve is given by $\mu = g_m r_a$, we can find its value from the figures for g_m and r_a at these two bias points ; at $E_g = -2$, $\mu = (1·45/1,000) \times 1,000,000 = 1,450$, while at $E_g = -1$, $\mu = (2·45/1,000) \times 350,000 = 880$.

In the triode, the amplification factor is determined almost entirely by the geometry of the valve, and therefore does not vary over these considerable ranges; further, it is much lower, seldom exceeding 100. Nevertheless, the screen-grid valve, used as a radio-frequency amplifier, does not give such enormously enhanced gain as these startingly high figures might suggest, for their effect is very largely offset by the valve's very high anode resistance.

134. Finding the Gain

Fig. 127 *a* shows a simple tuned-anode stage of radio

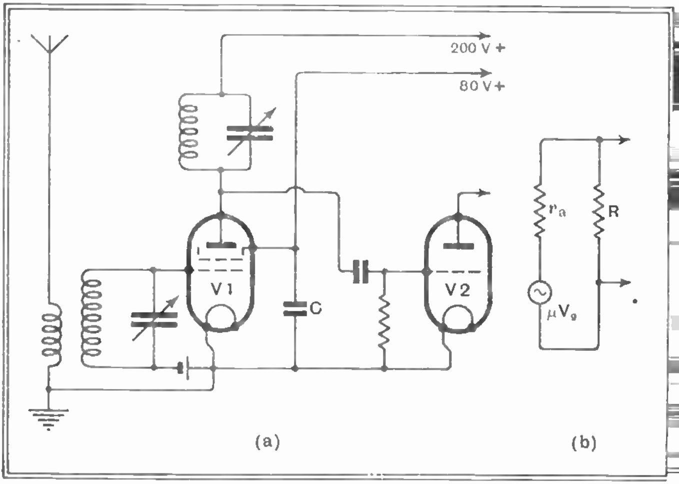

Fig. 127 *a*: A simple R.F. stage employing a screened valve with tuned-anode coupling; *b* equivalent anode circuit of the valve. If R is small compared with r_a, gain of stage is approximately $g_m R$

frequency amplification, preceding a grid-detecting triode V_2 with the exception of the addition of the screen circuit, with its large by-pass condenser to earth, the arrangement is the same as that for a triode. At *b* is shown the equivalent anode circuit of the valve, the signal-voltage V_g at the grid being represented, as before, by μV_g volts in series with the anode resistance of the valve. If R, the dynamic resistance of the tuned circuit, is 100,000 Ω, the amplification given by the

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stage, calculated from the formula $A = \dfrac{\mu R}{R + r_a}$, works out as 196 times for $E_g = -1$ and 141 times for $E_g = -2$. The rising amplification factor has been accompanied by so large a rise in anode resistance that the gain actually *drops* in passing from $E_g = -1$ to $E_g = -2$.

In most practical cases the resistance of the valve is so very much higher than that of the tuned circuit connected to its anode that R is small compared with r_a. A good approximation to the correct value for the stage-gain can then be had by writing $A = \mu R/r_a$, or $A = g_m R$.* The conditions for high gain with a screen grid valve are therefore simply that we choose a valve of high slope and follow it with a tuned circuit of high dynamic resistance.

Apart from these considerations the screen-grid valve behaves exactly like a triode from which the grid-anode capacitance has been removed; all the principles and methods discussed in Chapter 7 can therefore be applied to the tetrode.

135. The Limits of Stable Amplification

The introduction of the screening makes it quite possible to build up and use successfully a circuit such as that of Fig. 127 *a* without running into difficulties due to oscillation. It can be shown that the stage will be stable provided that the numerical value of a quantity H is less than 2. This quantity is given by the relation $H = 2\pi f g_m C_{ag} R_1 R_2$, where f is the frequency of the signal being amplified, and R_1 and R_2 are the effective dynamic resistances of the tuned circuits connected to grid and anode. High values of R_1 and R_2, which imply circuits of low inherent losses, tend, as might be expected, to produce oscillation. So also do high values of valve slope or grid-anode capacitance, while the likelihood of instability is greater, other things being equal, the higher the frequency of the signal it is desired to amplify.

For a valve for which $g_m = 2{\cdot}5$ mA/V, $C_{ag} = 0{\cdot}005\ pF$, used at 1,500 kc/s (200 metres), we can find now the maximum dynamic resistance that the tuned circuits can have without causing oscillation. For critical oscillation $H = 2$, so that we can write $R_1 R_2 = \dfrac{2}{2\pi f g_m C_{ag}} = \dfrac{2}{118} \times 10^{12}$. If the two tuned circuits are alike each may have a maximum dynamic resistance equal to the square root of this; i.e., of 130,000 ohms. Since

* g_m in amps per volt and R in ohms, *or* g_m in milliamps per volt and R in thousands of ohms.

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this represents a tuned circuit only a little better than the average, it is clear that the inter-electrode capacitance assumed for the valve is just on the maximum permissible limit for a single stage of amplification. In such a case the amplifier, though just stable, will be quite near oscillation, and we have a condition in which feed-back through the valve is not far from sufficient to reduce the radio-frequency resistance of the grid circuit to vanishing point.

136. Magnetic Screening

The formula given in the preceding Section assumes that there is no back coupling other than through the anode-to-grid capacitance of the valve, or any external wiring, etc., in parallel with it. In other words, any risk of oscillation results from the amplifier circuit being the same as that of the T.A.T.G. oscillator.

But it must also be considered as a possible reaction-coil oscillator, a simple circuit of which appears as Fig. 74. There is no intentional magnetic (or inductive) coupling, but accidental coupling is difficult to avoid when two tuning coils are mounted in the same unit. We have seen how a very minute capacitive coupling is enough to make a R.F. amplifier unstable, so it is not difficult to imagine that a very small stray induction has the same effect. Most radio apparatus is required to be compact, which means that the coils cannot be spaced far enough apart to avoid risk of instability. The larger the coils, the greater the risk, for two reasons : the magnetic field spreads out more, and the R.F. resistance tends to be lower.

If a coil is enclosed in a metal box or can, constructed so as to give a continuous low-resistance path parallel to the coil-winding, it is equivalent to a transformer secondary winding with a single short-circuited turn. The currents induced in it by the tuning coil set up a magnetic field which opposes that due to the coil ; so the resultant field outside this screen is greatly reduced.

If the coil is surrounded closely by the screen, its own self-inductance is largely neutralised, which means that more turns must be used to tune to a given frequency, and the R.F. resistance is greater. So it is desirable for the diameter of the screening can to be not less than twice that of the coil ; and even then some allowance must be made in the design of the coil for the reduced inductance due to the screen.

Comparing magnetic (or inductive) screening with electric (or capacitive) screening, note that it necessitates low resistance

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paths parallel to the coil winding but it need not be earthed, whereas capacitive screening need not provide closed circuits but must be effectively earthed or kept at some constant potential. Coils need both kinds of screening, which is

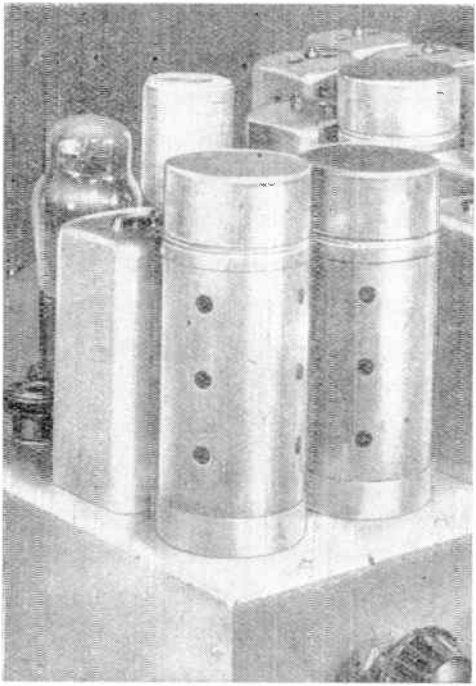

Fig. 128 : Typical coil screening

provided by completely closed and earthed metal covers (Fig. 128), because the capacitance between them would be enough to by-pass the internal valve screening, quite apart from inductive coupling.

137. Two Stages

For reception of the most distant stations, the gain given by a single stage of amplification is hardly adequate and it is

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desirable to add a second. This brings up, in much more acute form, the difficulty of instability, and experience shows that it is very difficult to persuade two tuned-anode stages to refrain from self-oscillation.

Examination of the two-stage tuned-anode amplifier of Fig. 129 shows that the tuned circuit 2, besides being in the anode

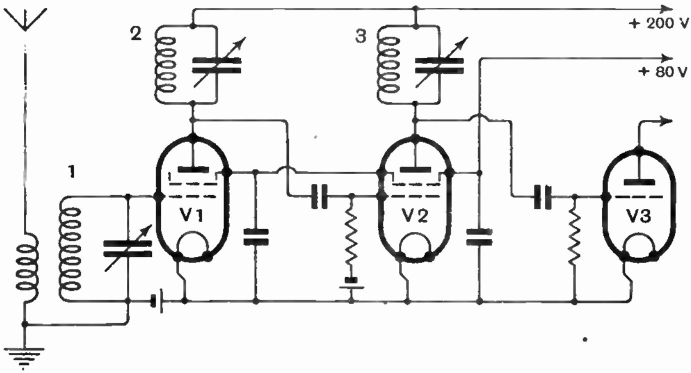

Fig. 129: Two-stage radio-frequency amplifier, using tuned-anode circuits with screen-grid valves

circuit of V_1, serves as grid circuit for V_2, being connected between the grid of that valve and the H.T. line. This, being at zero potential so far as signals are concerned, counts as " earth " from the A.C. point of view. To keep the grid of V_2 at the right potential for amplification, grid bias is connected through a resistance high enough not to damp circuit 2 excessively. In its role of grid-circuit to V_2, the tuned circuit has energy fed into it through the valve, and so has its R.F. resistance reduced well below its normal value. This results in giving it a very high dynamic resistance, and it is this artificially-raised figure that must be taken for R_2 in applying the formula to compute the stability of the first stage. As the formula shows, a rise in R_2 increases the tendency to oscillation, and we conclude that two stages, each individually stable, may oscillate if connected in cascade as in Fig. 129.

In addition, of course, the need for magnetic screening is more acute the greater the amplification.

138. Transformer Coupling

Feedback from the anode of V_1 to its grid can be reduced—by cutting down the signal-voltage at the anode. Naturally

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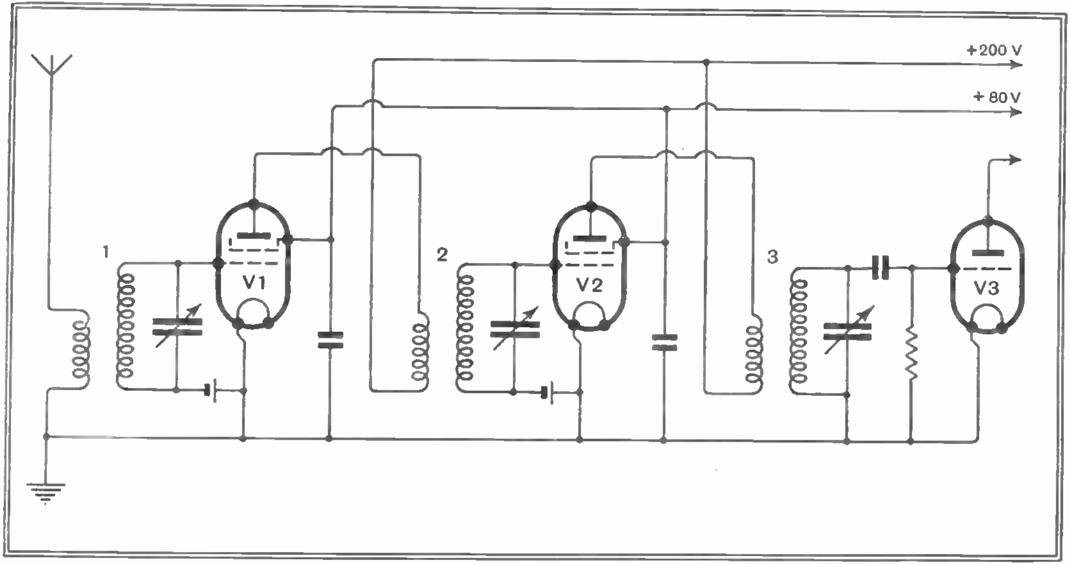

Fig. 130 : Two-stage radio-frequency amplifier, using step-up transformer couplings with screen grid valves. Much more stable than the closely-corresponding circuit of Fig. 129

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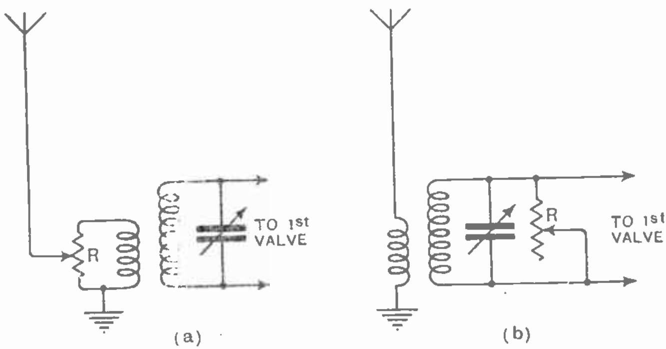
one dislikes sacrificing gain, so that one would like to maintain as nearly as possible the signal-voltage eventually reaching the grid of V_2. This can best be done by replacing each tuned circuit with a radio-frequency transformer of the conventional type, using a tuned secondary and an untuned primary. The conversion of Fig. 129 to this more stable arrangement is shown completed in Fig. 130. The exact turns-ratio that gives best results in such a case is usually best found by experiment, but the gain can readily be computed for any ratio to which a search for stability may lead us.

If the secondary has a dynamic resistance R, and contains n times as many turns as the primary, the effective resistance of the latter will be R/n^2. Following a valve of slope g_m, the signal-voltage at the anode will therefore be g_mR/n^2 times that



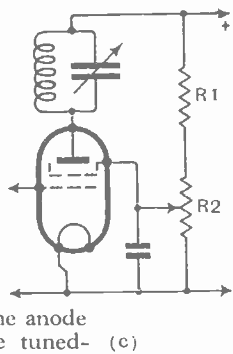


Fig. 131 :- Three general methods of controlling the signal passed to V_3 in Fig. 130

at the grid, while at the grid of the succeeding valve it will be n times this owing to the voltage step-up in the transformer. This makes the gain, reckoned from grid to grid, equal to g_mR/n.

Thus if we replace a tuned-anode coupling, the gain for which is g_mR, by an R.F. transformer of ratio n, we divide the stage-gain by n and the voltage at the anode of the valve by n^2. Thus we can cut down the signal-voltage at the anode to one-ninth of its value in the simple tuned-

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anode circuit at the cost of dividing the gain of the stage by only three.

139. Volume Control

In order to prevent the detector-valve V_3 from being grossly overloaded when receiving a near-by station, it will be necessary to add to the circuit of Fig. 130 some form of *volume control* by manipulation of which the overall gain of the amplifier can be adjusted. By this means it is possible to ensure that the signal-voltage reaching the detector is kept at a constant level irrespective of the voltage produced at the aerial by the particular transmitter tuned in.

Volume control can be obtained in the three general ways illustrated in Fig. 131 : by controlling the input from the aerial,

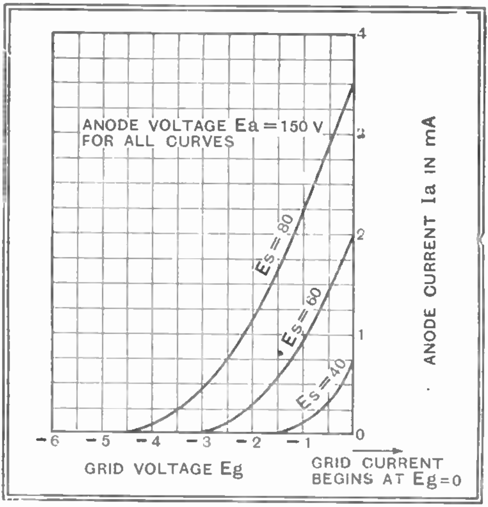

Fig. 132 : Curves of ordinary screen-grid valve. Note that rectification (overload) can occur on quite a small signal, especially when E_s is reduced

as at *a* ; by controlling the magnification of one or more tuned circuits, as at *b* ; or by controlling the gain given by the valve, as at *c*. With method *a* the amplifier works always at full gain, in which condition it is likely to produce a certain amount of background noise (" valve-hiss ") which, while tolerable in listening to a distant station, must be avoided, if possible, while listening to a near one.

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For this reason method *a* is not used save as an auxiliary to some other type of control.

Method *b* suffers from the drawback that in reducing the gain of a tuned circuit its selectivity is reduced also; except for local-station reception, where this is sometimes considered an advantage, this type of control is not used.

Method *c* seems to be ideal, since it supplies a means of controlling gain by reducing the amplification given by the valve, the slope of which drops as E_s is decreased, without affecting any of the other characteristics of the amplifier. In the particular form shown in the diagram, however, it leaves a good deal to be desired, as can be seen by referring to the curves of Fig. 132.

140. Distortion due to the R.F. Amplifier

Here are shown the E_g—I_a curves of a typical screen-grid valve, and it is at once evident that when the voltage on the screen is lowered the available portion of the characteristic, lying between the grid-current region and cut-off, is neither long enough nor straight enough to accommodate a signal of any but very small voltage. As always, a curved characteristic means rectification, with its accompanying distortion, and it is clear that with such a volume control as this, distortion will be greatest where we can least tolerate it—when receiving the local station, because then the input voltage is greatest and the screen voltage lowest.

The effect of reducing the screen voltage is to reduce the *grid base* (the range of voltage between grid current and cut-off); which is just the opposite to what is wanted.

Looking at Fig. 132, suppose that the grid bias is $-1\frac{1}{2}$ volts and the carrier wave has a peak amplitude of 0·5 volt, and is modulated 100 per cent. by an audio-frequency sine wave. The input to the valve will therefore be like Fig. 133 *a*. With the screen of the valve at 80 volts, the anode current variations due to this signal (neglecting the effect of the anode load impedance) are as in Fig. 133 *b*. There is slight rectification—the negative current peaks are not quite as pronounced as the positive—but this in itself may not be serious. The A.F. envelope is still very nearly the same shape as that of the input.

But when the screen voltage is reduced to 40 (Fig. 133 *c*) with the object of reducing the signal output, this reduction is accompanied not only by total rectification, but also by gross distortion of the modulation. The positive peaks of the audio envelope are high and sharp, while the negative are shallow

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and broad. The results would be unpleasantly audible ; still more so if a yet larger input amplitude were employed.

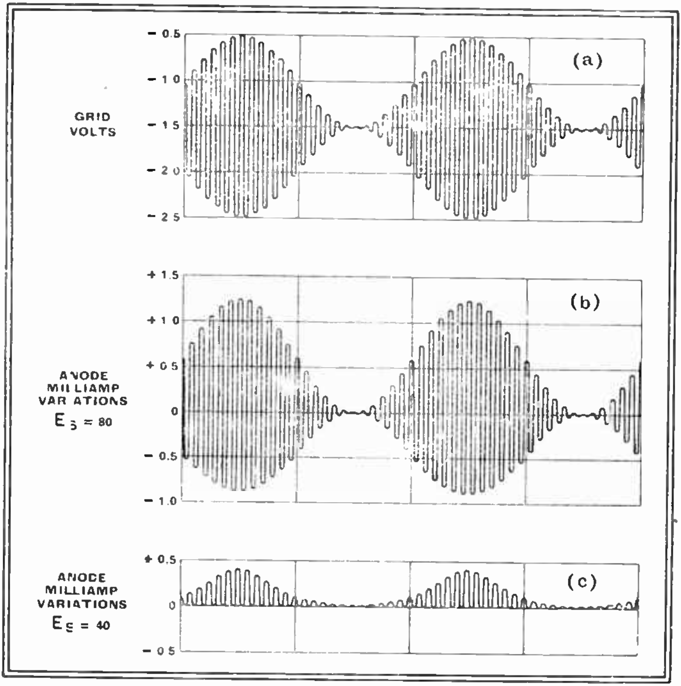

Fig. 133 : If the modulated R.F. voltage represented by *a* were applied to the valve whose characteristics are shown in Fig. 131, the corresponding current output would be *b* or *c* with a screen voltage of 80 and 40 respectively. Note the distorted outline of *c*

141. Cross-Modulation

Besides this distortion of a single modulated carrier, there is a type of distortion, known as *cross-modulation*, which makes its appearance under the misleading guise of lack of selectivity. It arises like this. Suppose that the receiver of Fig. 130 is tuned to a station 45 kc/s away from the local. We may very well assume that the overall selectivity of the three tuned circuits is enough to reduce the local station to inaudibility when they are all tuned 45 kc/s away from it. But the grid of the first valve is protected from the local station by only one tuned

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circuit ; it is not impossible that at this grid this station may produce quite a large voltage. If this voltage is large enough to cause the valve to rectify, its positive peaks will carry the grid voltage to a point at which the valve will amplify more strongly, while the negative peaks will so increase the bias as as to cut off anode current altogether, reducing the amplification to nil. The amplification of the desired programme is therefore varied according to the modulation of the interfering programme, which impresses itself on the carrier wave of the former. Since the set is tuned to this carrier, the remaining two tuned circuits will pass it along, together with its twin programmes, to the detector, which will make both stations audible together. If the station to which the set is tuned switches off its carrier wave the programme of the interfering local station will also disappear, thereby proving beyond all doubt that the interference is due to cross-modulation, and not simply to lack of selectivity in the tuned circuits.

For all practical purposes, the selectivity of a set in which cross-modulation is occurring is no greater than that of the tuned circuit preceding the first grid. It used therefore to be common practice to interpose two tuned circuits between the aerial and the first valve.

For more satisfactory prevention of cross-modulation we have to replace the first valve by one which overloads less readily, so that quite large voltages from the local station can reach it without causing rectification. Further, this new valve must be suited to some means of gain-control other than that obtainable by reduction of screen voltage, which inevitably reduces the signal-acceptance of the valve.

142. The Variable-Mu Tetrode

Controlling gain by increasing negative bias is not going to be satisfactory either, because (referring to the "$E_s = 80$" curve in Fig. 132, for example) with a weak signal, for which the volume control would be set to make the bias small, there is a more than ample range of reasonably straight characteristic, whereas with a progressively stronger input the control has little reducing effect until the bottom bend is reached and anode current is nearly cut off, causing severe distortion.

Simply widening the spacing of the grid wires, to give a longer grid base, is not going to help matters, because the characteristic will still be straightest when the signal voltage is least, and most sharply curved when it is greatest. And the rate of control is still very slight over most of its range, and concentrated at one end.

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What we want is a characteristic extending over a very large grid base, and curving gradually over the whole range, so that over any part of it (and especially the lower reaches) a considerable signal voltage can be handled without serious distortion of the modulation envelope. Over the upper part the curvature can be greater, because the signal voltage is then normally small.

Such a characteristic can be achieved in a screen-grid valve

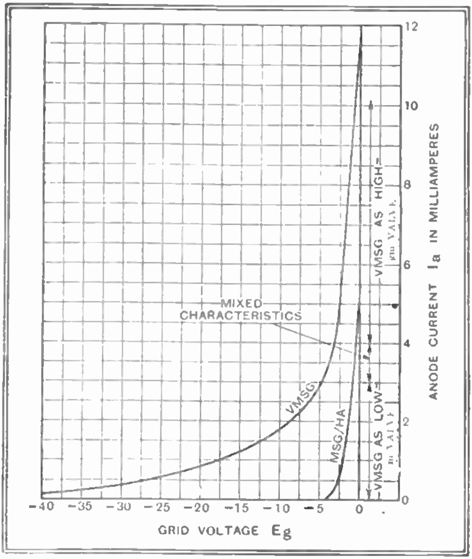

Fig. 134 : Characteristic of variable-mu valve (VMSG) compared with that of standard screen-grid valve. Note the increased signal-handling ability of the VMSG and the slow but steady change of slope with bias

by winding the control grid with a variable pitch. Where the mesh is close the grid base is short, and anode current is cut off by a large negative bias. Where the mesh is widest and the grid base consequently large, the same bias will permit anode current to flow ; but as only a small part of the valve is effective, its mutual conductance is low. As the bias is reduced, more and more of the valve comes into action and the mutual conductance increases smoothly.

We have already seen (Sec. 134) that the amplification given

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by a screen-grid valve is approximately proportional to its mutual conductance, so here is just what we want.

This is known as the variable-mu valve; the name can be regarded as signifying that the mutual conductance is variable over a wide range. The μ is not varied to any great extent

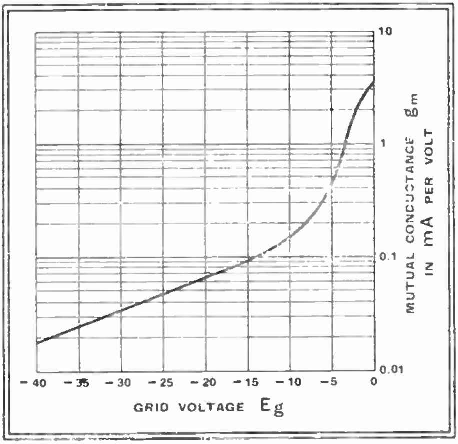

Fig. 135: How the mutual conductance of a variable-mu valve (VMSG) is affected by alterations of bias

by control of grid bias, though it does vary geometrically along the length of the grid itself.

In Fig. 134 is plotted the E_g—I_a curve of a variable-mu valve, the curve of an ordinary screen-grid valve being plotted, for comparison, on the same diagram. For a large input, which might need a grid voltage of about −25 to reduce the output to the required level, the variable-mu type is seen to provide a large range of reasonably straight characteristic. Fig. 135 shows how the slope varies with applied bias, and makes clear how the amplification can be controlled smoothly over a wide range by varying the bias. In fact, it is necessary to use a logarithmic scale (footnote to Sec. 72) for plotting such a large range of g_m. Such a scale is in any case appropriate, because it corresponds more closely than an ordinary scale to the impression of loudness given by the ear.

Because of these advantages of the variable-mu valve, the short-grid-base screen-grid valve is rarely used in broadcast receivers, though it is still useful for applications where variable

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gain is not involved. Other things being equal, its maximum slope is slightly greater than that of the variable-mu type.

In no other respect than those just touched upon is there any difference between the two types of valve ; with the obvious minor modifications, all that has been said about the simpler valve may be applied, without alteration, to its successor.

143. Secondary Emission

While the introduction of variable-mu characteristics overcomes with fair completeness overloading and distortion arising in the grid circuit of the valve, there remain possibilities of trouble in the anode circuit. These arise owing to the peculiar shape of the $E_a - I_a$ curve, which is shown in full line in Fig. 136. If the sole effect of raising the anode voltage were to rob the screen of more and more electrons, the valve curves would take a form such as that shown dotted on the same

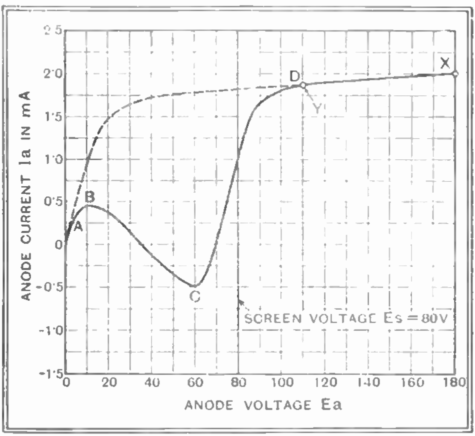

Fig. 136 : "Theoretical" (dotted) and actual (full line) curves of tetrode valve. The extraordinary shape of the latter between A and D is due to secondary emission from the anode. The introduction of a "suppressor" grid, turning the valve into a pentode, enables the dotted curve to be realized in an actual valve

diagram. Why the divergence between theory and observed fact ?

As always when theory and practice do not agree, the theory has overlooked something. In the present case it has omitted to take into account the phenomenon of *secondary emission*, by

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which is meant the ability of a fast-moving electron to knock out another electron when it strikes a metal surface. Once liberated, free electrons so produced will naturally be attracted to the most positively charged object in their neighbourhood.

At low anode voltages the real curve follows the dotted one, but at A the velocity of the electrons has risen enough to enable them to dislodge secondary electrons from the anode on their arrival there. These electrons are attracted to the more positive screen, so reducing the net number of electrons arriving at the anode, and reducing the anode current below the "theoretical" value. Beyond B, the peak of the curve, each extra electron drawn to the anode by rising voltage knocks out more than one when it gets there, and these all reach the screen, which still has the higher potential. The current, therefore, *decreases* with rising anode voltage. It even reverses in direction, this meaning that the total number of electrons arriving at the anode is less than the number they dislodge by secondary emission.

At higher anode voltages than that at C, the secondary electrons begin, in increasing numbers, to return to the anode, allowing the anode current, therefore, to begin to return towards its "theoretical" value. Finally, as soon as E_a exceeds E_s by a small amount (at D) the superior attraction of the anode prevents any from reaching the screen. The observed curve has now joined the dotted curve, showing that secondary emission no longer has any effect on the net anode current.

Secondary emission, although it must occur, does not distort the characteristic curves of a triode valve, for the excellent reason that secondary electrons, when emitted, always return to the anode, since it is the only positively charged object near them. The total anode current is thus not altered by their temporary absence from the anode.

Consideration of the full-line curves in Fig. 136 makes it perfectly clear that if the voltage at the anode is swung by the signal so far that it falls momentarily down to that of the screen, violent distortion is likely to occur. If, for example, $E_a = 180$ V and $E_s = 80$ V, the maximum permissible signal swing at the anode is about 70 volts peak (from X down to Y); after that, rapid curvature begins.

144. The Screened Pentode

Admittedly, signal voltages of this order are generally adequate, in a radio-frequency stage, but are dependent on the screen voltage being kept sufficiently low. It is not always desirable to do this, however. The source of the trouble can

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be removed by inserting between screen and anode an extra grid, connected to cathode, which will serve to protect the electrons dislodged from the anode from the attraction of the screen, so ensuring that, as in the case of the triode, they all return to the anode. This extra grid is called a *suppressor grid* by virtue of the fact that it suppresses secondary emission, and a valve containing it, having five electrodes, is known as a *pentode*. The shape of the $E_a - I_a$ curves of the pentode is, as theory predicts, practically that of the dotted curve of Fig. 136.

Like the screened tetrode, the screened pentode is available in both variable-mu and short grid base types ; the former is intended primarily for R.F. amplification, while the latter has a number of special uses. The addition of the suppressor still further reduces the influence of the anode in determining the total space-current through the valve ; in other words, the pentode has a higher anode resistance (and consequently a higher amplification factor) than a corresponding tetrode of the same slope. Since, in a radio-frequency amplifier, the valve is shunted across the tuned circuit (as in Fig. 129), this high anode resistance results in a slight gain in selectivity as compared with the tetrode.

The pentode thus has several advantages over the corresponding tetrode : a possible slight improvement in selectivity ; the total elimination of the possibility of anode circuit overload through running the anode to a voltage less than that of the screen ; and the simplification of being able to supply the screen voltage from the same source as the anode.

145. The Kinkless Tetrode

An alternative method of preventing secondary electrons from being attracted to the screen-grid is to increase the distance between it and the anode sufficiently for the negative space-charge of the stream of electrons flowing towards the anode to neutralize the positive attraction of the screen. This distance is critical because if too great the electrons will be turned back from the anode. So this valve is sometimes called the critical-distance tetrode. There are other details of design that are employed to the same end, which is to obtain a tetrode without the " kink " in the anode-current characteristic, so evident in Fig. 136. The result is a tetrode so similar in characteristics to a pentode that it is difficult to distinguish between them. Almost the only difference is a tendency in the tetrode for the nearly flat top in the $I_a - E_a$ characteristic to be reached at a lower anode voltage, so increasing still further

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the amplitude of signal that can be handled without distortion. There is no great need for this in a R.F amplifier, but it is one of the prime aims in an audio-frequency output stage, as will be seen in Chapter 14.

It may be useful to give here a summary of the outstanding characteristics of each of the types of valves so far discussed.

Diode : Two electrodes (cathode and anode) only. Rectifies, but will not amplify.

Triode : Cathode, grid, and anode. Amplifies, oscillates, and detects. Is the fundamental type of valve, from which more elaborate structures have developed.

Screened Tetrode : A triode with addition of a screen between anode and grid to prevent instability. Anode resistance and amplification factor very high.

Variable-mu Screened Tetrode : As preceding, but with grid-circuit overload reduced and adequate means of gain-control provided.

Screened Pentode : As screened tetrode, but capable of dealing with large signal at anode.

Variable-mu Screened Pentode : Combines advantages of both the two preceding valves.

Kinkless Tetrode : Handles the largest anode signal voltage of any, and is used mainly for A.F. output stages.

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CHAPTER 13

SELECTIVITY IN THE R.F. AMPLIFIER

146. Resonance Curves

IN trying to raise the sensitivity of a single-valve set we first tried reaction, using it to reduce enormously the radio-frequency resistance of our simple set's one tuned circuit. The terribly over-sharp tuning and consequent loss of sidebands that accompanied this attempt led us to reject it in favour of obtaining amplification by the aid of additional valves as radio-frequency amplifiers. We then found that to make these amplify satisfactorily we had to introduce extra tuned circuits. The question at once arises whether, in adding these extra circuits, we have not committed ourselves to just as great an accentuation of selectivity as we originally got with a single circuit and reaction. To settle this point we shall have to go a little more deeply into the subject of resonance curves, both of single circuits and of several in combination.

From the point of view of the adequate reception of high notes, all we need to know is the amount by which the response of our tuned circuit drops at a frequency removed from resonance by the frequency of the musical note we wish to consider. This depends solely on the ratio of the inductance of the coil to the radio-frequency resistance of the tuned circuit as a whole. For all wireless problems, we are only concerned with the response at frequencies not very far removed from resonance, for finding which the formula that follows, although a little simplified, is amply accurate.

If the voltage across the tuned circuit at resonance is V_0, and that across it for a frequency n cycles from resonance is Vn, then $V_0 = Vn\sqrt{1 + (4\pi n)^2\left(\frac{L}{r}\right)^2}$.

The complete square root, which we will hereafter abbreviate to s (for selectivity), tells us by how much we must multiply the voltage at n cycles from resonance to get the voltage at the resonant point. If $s = 4$ at 10 kc/s off tune, the voltage at this frequency is one-quarter of that at resonance, and we speak of the circuit as being "four times down at 10 kc/s off tune."

The expression for s is rather a troublesome one to evaluate quickly for a rapid comparison of the selectivity of different circuits; actual values of s are therefore shown for L/r ratios up to 500 in the curves of Fig. 137. Separate curves are given for 5, 9, 18 and 27 kc/s off tune.

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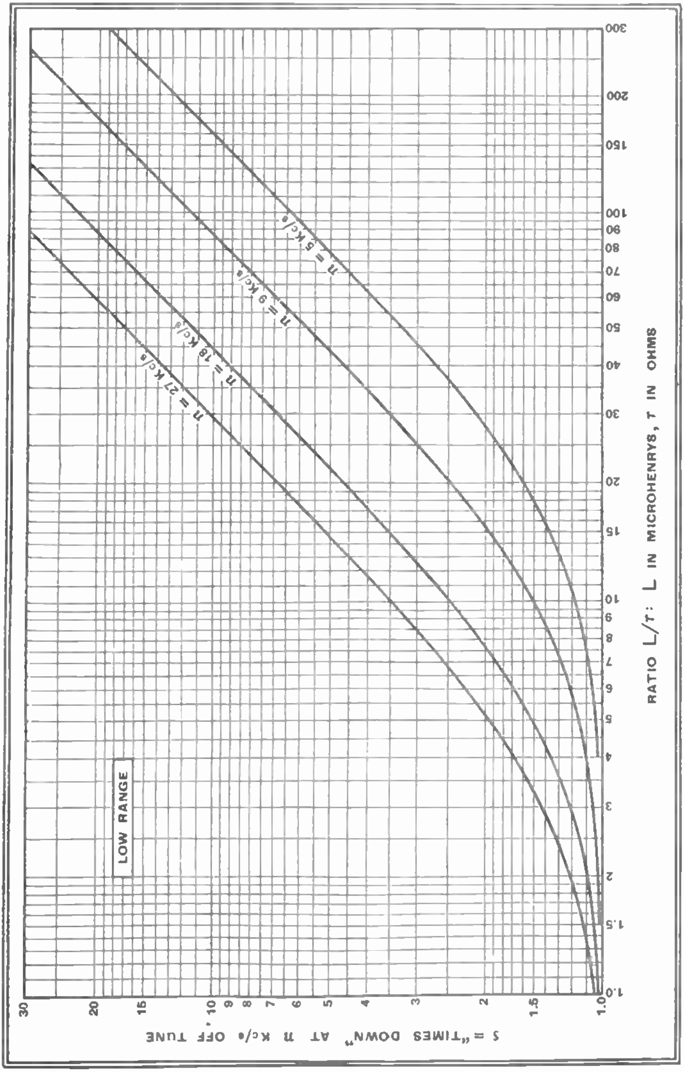

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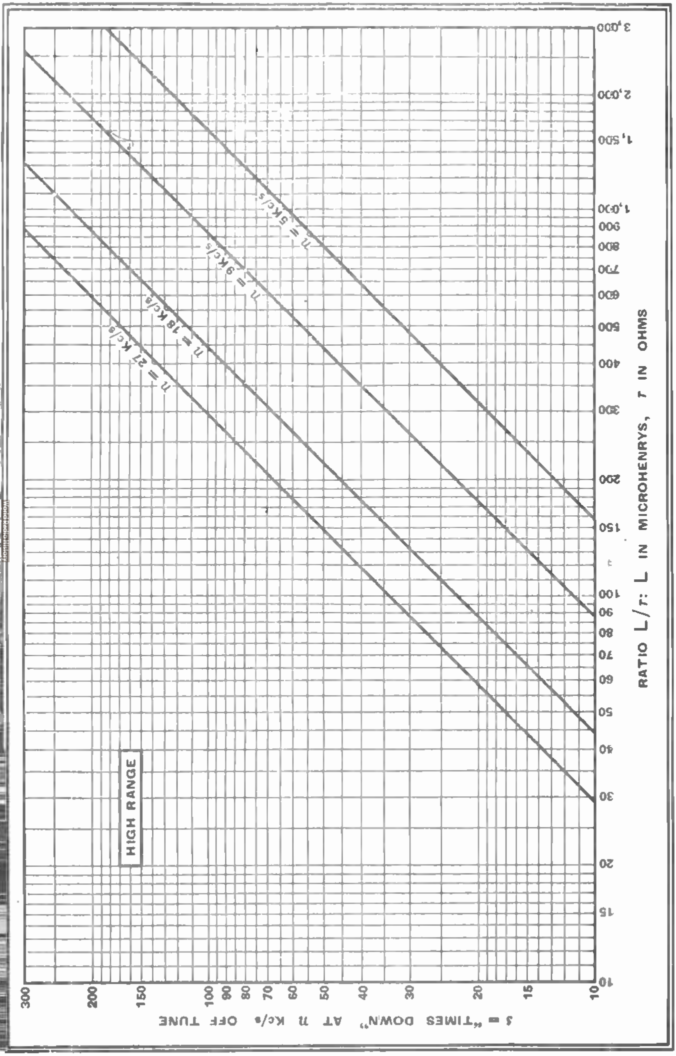

Fig. 137b : Continuation, to higher values, of the curves of Fig. 137a

147. Reaction and Amplification Compared

With the aid of these curves we are in a position to compare at once the selectivity of a reacting detector with that of a set containing a single stage of radio-frequency amplification and therefore employing two tuned circuits. If we assume that at some particular frequency the ratio L/r of the tuned circuits in the amplifier is 10, then we see from Fig. 137 that at 5 kc/s off tune each circuit has its response reduced to $1/1 \cdot 18$ of that at resonance. For two tuned circuits the overall response will be the square of this, or $1/1 \cdot 39$; that is, the amplifier will pass 72 per cent. of the side-bands representing high notes of frequency 5,000 cycles.

If the gain of the stage is assumed to be fifty times, then to get equal amplification by means of reaction we shall have to reduce r to one-fiftieth of its normal value, thereby increasing L/r to 500. Reference to Fig. 137 shows that with L/r raised to this value s becomes 30, making the response at 5 kc/s off tune one-thirtieth that at resonance. In this one tuned circuit side-bands are so cut that only some 3 per cent. of a 5,000-cycle note will reach the speaker. The loss at this frequency is thus some 24 times as great as when using the extra tuned circuit necessitated by the valve, although the amplification afforded is in each case the same.

148. Separating Stations

Selectivity is often regarded as the ability of the set—which means of the tuned circuits in it—to select one station to the exclusion of others. In allotting wavelengths to the various transmitting stations, international agreement has resulted in a uniform spacing, from each station to the next, of 9 kilocycles per second. The 9 kc/s gap between carriers is left to cover the " spread " of frequency taking place as a result of modulation of the carriers by the programme.

If one station transmits at 191 kc/s, its two neighbours will transmit at 200 and 182 kc/s respectively. The wavelengths corresponding to these frequencies are, in order, 1500, 1571 and 1648 metres, making an average spacing between stations of 74 metres. If we consider stations transmitting at much higher frequencies, the same 9-kc. separation holds, because the width of sidebands is determined by the audible *frequencies* in the programme, and has nothing to do with the wavelength of the carrier. Three stations in order from the list transmit on 1474, 1465 and 1456 kc/s: expressed in wavelengths, these frequencies

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are equivalent to $203\frac{1}{2}$, $204\frac{3}{4}$ and 206 metres, a spacing between stations of $1\frac{1}{4}$ metres.

These figures make it abundantly clear that separation between stations cannot intelligibly be expressed in metres ; a proud boast that " My set will separate stations only 20 metres apart " means nothing at all unless there is also specified the wavelength at which this prodigy of selectivity (or woeful lack of it, as the case may be) was observed. We shall therefore have to deal with selectivity exclusively in terms of frequency. The figures further show that we need not specify the actual carrier-frequency in use ; all that concerns us is the amount by which the reponse of our tuned circuit drops at some known number of kc/s from resonance.

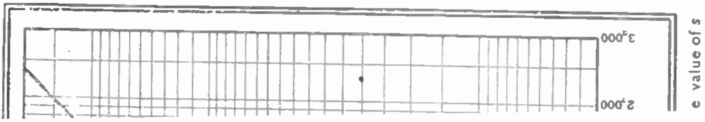
We shall find it convenient to use, therefore, the formula and curves already discussed in considering quality of reproduction.

149. Conflicting Claims

Since we would like to retain at full strength frequencies off tune by at least 5 kc/s for the sake of quality, and yet, for the sake of selectivity, would like to remove as completely as possible frequencies 9 kc/s off tune, we require, if we can get it, a resonance-curve with a flat top and steeply-falling sides.

Some approximation to this can be obtained by using a large number of fairly flatly tuned circuits in cascade. Where several circuits are so used the overall s is found by raising the s-value for one circuit to the appropriate power—squaring for two circuits, cubing for three, and so on. To enable the reader to find for himself the behaviour of any series of tuned circuits in which he may be interested, Fig. 138 gives curves in a rather more general form than Fig. 137. In place of plotting s against L/r, and making a separate curve for each value of n, s is here plotted against the product $n \times L/r$. Curve 1 refers to one tuned circuit, curve 2 to two circuits, and so on up to a total of six circuits, all connected in cascade.

To find, for example, " times down at 9 kc/s " for a series of circuits for each of which $L/r = 10$ we only have to multiply 10 by 9 to find $n \times L/r$, and look up the required figure on the curve corresponding to the number of tuned circuits for which the result is required. For one tuned circuit we find that $s = 1{\cdot}5$, for two $2{\cdot}25$, for three $3{\cdot}38$, and so on. Alternatively, to find the requisite L/r to give 10 times down at 9 kc/s with four circuits, the value of $n \times L/r$ corresponding to $s = 10$ is read off from the curve for four circuits, and is found to be 117. The required L/r is then $117/9 = 13.0$.



FOUNDATIONS OF WIRELESS

150. Equal Selectivity

Suppose, for example, we want to reduce the voltage of an interfering station 18 kc/s off tune to one-hundredth of the voltage it would have if exactly tuned in. As Fig. 137 shows, a single circuit to do this has $L/r = 450$, with which a 5 kc/s side-band will be 28 times down. If we used six tuned circuits the value of $n \times L/r$ required, as Fig. 138 shows, is 152, giving $L/r = 152/18 = 8{\cdot}4$. At 5 kc/s off tune, $n \times L/r = 5 \times 8{\cdot}4 = 42$, corresponding on Fig. 138 to $2{\cdot}1$ times down. Thus, for the same discrimination against an unwanted carrier 18 kc/s removed from that required, six circuits give over 12 times as great a response to a 5-kilocycle side-band.

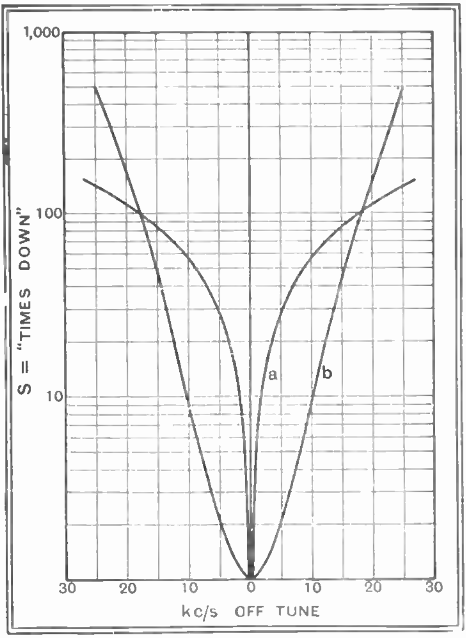


Fig. 139 : Resonance curves of one (*a*) and six (*b*) tuned circuits, chosen so as to give in each case 100 times reduction at 18 kc/s off tune. Note the enormous loss of side-bands in case *a*

To make this point clearer, Fig. 139 shows the complete resonance curve, derived from Fig. 138, for the two cases. The curves show very clearly that, although in both there is the same discrimination against a station 18 kc/s removed in frequency from that required, the single tuned circuit can provide this selectivity only at the cost of lopping off the sidebands of the desired transmission to a very drastic extent. The more rounded curve for six tuned circuits, though by no means perfect, offends very much less in this respect.

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To reach so high a value of L/r as 450 it would be necessary to use a good deal of reaction, so that these two curves may be taken as illustrating, from a different angle, the dangers of trying to make reaction do too much. We have seen already how it destroys quality when used as a substitute for true amplification ; the curves of Fig. 139 emphasize that its use to provide selectivity that should be attained with additional tuned circuits brings just the same dire results in its train. These comments apply, of course, only to the excessive use of reaction ; in off-setting detector damping, and perhaps providing, in addition, a *little* extra selectivity or sensitivity it is invaluable, expecially in the less ambitious receiver.

151. Equal Quality

We have taken, perhaps, an extreme case in comparing the resonance curves of one and six tuned circuits. A more

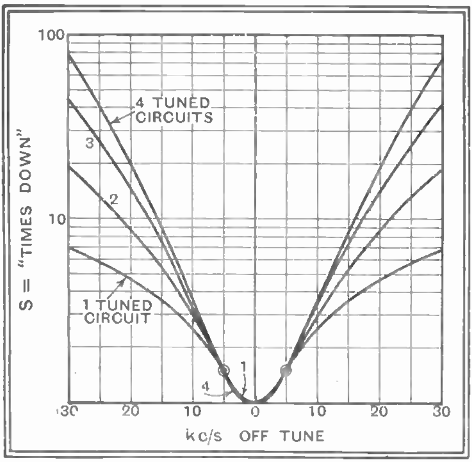

Fig. 140 : Overall resonance curves of one, two, three and four tuned circuits, in each case chosen to give "equal quality", represented by the same response to a 5-kc/s sideband

practical comparison is that shown in the four curves of Fig. 140. Here we can see the differences in selectivity obtained by using one, two, three, or four tuned circuits, the L/r values in each case being chosen to give $1\frac{1}{2}$ times down at 5 kc/s—that is, a reduction of 5,000-cycle notes to two-thirds of their correct

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voltage. This corresponds to a barely noticeable loss at this frequency.

For one tuned circuit we require that $L/r = 18$, which is by no means an outrageous value. The selectivity is poor, a station even three channels (27 kc/s) away being reduced only some six times. With two tuned circuits L/r for each comes out at 11, and a station 3 channels away is now reduced about 16 times. Adding a third circuit and reducing L/r to 8·8 still keeps the quality unchanged, but increases the selectivity to 32 times down at 27 kc/s. A fourth circuit increases this figure to 52.

152. Practical Coil Figures

It is simple enough, on paper, to talk about the choice of correct L/r ratios to provide the response-curves that we desire. In practice it is not always easy, or even possible, to achieve them. Experience shows that a coil designed to tune, with its variable condenser, over a range of frequencies, always has a lower resistance at the longer frequencies. Constant selectivity, of course, requires the R.F. resistance to remain unchanged.

In the ordinary small coils used in the modern receiver, the ratio L/r is found to vary from about 5 or 6 at 1,500 kc/s (200 metres) up to about 20 at 550 kc/s (about 550 metres). If the coil has an iron-dust core and is wound with stranded wire in which the strands are insulated from one another ("Litzendraht") the 550-kc/s figure will probably rise to about 35, that for 1500 kc/s remaining approximately unchanged. The design of a coil for lowest attainable resistance requires the choice of correct wire-thickness, and the thickness required depends on the precise frequency for which the calculation is made. The figure given as representative for L/r can therefore be increased a little at either end of the waveband at the cost of a decrease at the other by designing the coil specifically for the frequency it is desired to favour. But the only really useful method of improving the L/r ratio is by increase in size of coil; this, of course, is effective at all frequencies.

153. Selectivity and Gain

It is an unfortunate fact that the more constant the selectivity is over the frequency-band, the less constant will be the gain. Gain depends, as we have seen (Sec. 134) on the dynamic resistance $R = \dfrac{L}{Cr}$ (Sec. 60); as we reduce frequency by increasing C, r diminishes and tends to hold constant the product Cr, and with it the dynamic resistance, since L does

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not change. It is usually found that R has a maximum at about 1250 kc/s, after which it falls steadily, till at 550 kc/s it is usually about half the maximum value.

If we really succeeded in keeping r, and hence L/r, constant from 1500 to 550 kc/s, we should get constant selectivity accompanied by a steady drop in R which, at 550 kc/s, would have less than one-sixth of its value at 1500. Conversely, constant R would give us marvellously constant gain, but to get it r would have to decrease in the same ratio that C increases, making L/r over six times as great at 550 kc/s as at 1500.

Tuning by varying L, keeping C constant, could theoretically avoid this difficulty, for then constant R would also mean constant L/r. But r shows no particular inclination to be strictly proportional to L in any variable-inductance tuner that has so far appeared.

154. Low Frequencies

On the low-frequency band, from 150 kc/s up to about 350, the coils generally used have an inductance of round about 2 millihenries in conjunction with an L/r ratio varying from 30 to 50 over the band. At these frequencies higher figures can quite easily be attained, but they are hardly desirable on account of the severe loss of sidebands to which they give rise.

Consideration of the various figures that have been mentioned will make it clear that the ordinary set reduces the side-bands to a considerable extent, and yet suffers to some degree at least from insufficient selectivity. In spite of many attempts, the problem of making a satisfactory compromise between the conflicting claims of selectivity and quality is really not soluble in the case of the radio-frequency amplifier. A nearer approach to the desired results can be attained in a superheterodyne receiver, in connection with which we shall return to the question in Chapter 17.

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CHAPTER 14

AUDIO-FREQUENCY AND OUTPUT STAGES

155. Resistance-Coupled Amplification

AUDIO-FREQUENCY amplification, by which is meant amplification of the signals after detection, is generally carried out in modern sets by some form of resistance amplifier. In Chapter 7 this method of amplification was discussed fairly fully, being taken as the type of amplification in general.

In Chapter 12 we found the method unsuitable for radio frequencies owing to the inevitable stray capacitances. In dealing with audio-frequencies these strays are naturally less harmful, but they may lead to a certain reduction in gain at the highest notes, for which their reactance is of course least, if care is not taken in the choice of component values.

156. High Note Loss

It can be shown that high notes of frequency f are reduced to 70·7 per cent. of their correct voltage when $1/2\pi fC = Rr_a/(R + r_a)$ (Fig. 141); that is to say, when the reactance of the stray capacitance is equal to the anode resistance of the valve in parallel with the load resistance. Any reduction in reactance (increase in capacitance) or increase in R or r_a leads to greater proportionate loss of high notes. It will be clear that where a high capacitance is inevitable (as in long screened leads, for example, or feeder lines to a distant amplifier), or the frequency is very high, the choice of a valve of low anode resistance, with a coupling resistance of low value, will help to keep the loss of higher frequencies within reasonable bounds. A better method for special purposes is described in Sec. 168.

157. Low Note Loss

In Fig. 141 the grid condenser and leak, C_1 and R_1 form a potential divider across the source of amplified voltage (anode of V_1 to earth). Only the voltage appearing on R_1 reaches the grid of V_2, any dropped on C_1 being lost. For the lowest frequencies, at which its reactance is highest, there may be an appreciable wastage of signal on C_1; correct relative values must be chosen if this is to be avoided.

Low notes of frequency f are reduced to 70·7 per cent. of their full voltage when $1/2\pi fC_1 = R_1$; another way of stating the same thing is to say that the time constant of C_1R_1 is $1/2\pi$—

roughly one-sixth—of the time period of one cycle of the voltage. Of course a reduction in R_1 or in C_1 further increases the proportionate loss of low notes. A usual combination is $C_1 = 0{\cdot}01\ \mu F$, $R_1 = 0{\cdot}5\ M\Omega$, with which a note of frequency

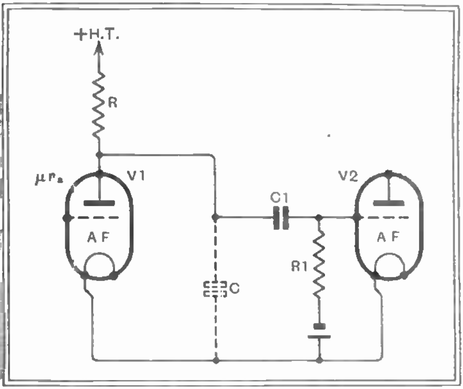

Fig. 141 : Showing stray capacitances C in an A.F. stage. High notes of frequency f receive 70·7% of the amplification which they would have were it not for the presence of C, when $1/2\pi fC = Rr_a/(R + r_a)$

about 32 cycles is reduced to 70·7 per cent. Doubling either C_1 or R_1 will reduce this frequency to 16 cycles, but it is doubtful whether the resulting improvement in bass reproduction would be noticeable with the average loudspeaker.

158. Transformer Coupling

A transformer is sometimes substituted for the resistance with the double aim of allowing a greater D.C. voltage to reach the anode of the A.F. amplifying valve (or detector) and of obtaining extra gain by virtue of the step-up ratio of the transformer.

In Fig. 142 there are shown skeleton diagrams of two varieties of transformer-coupled stage. Since we desire to amplify signals of all frequencies to the same extent, the voltage developed across the primary in circuit *a* must be independent of frequency. The primary constitutes an inductive load, the reactance of which rises with frequency ; to attain uniform amplification it follows, therefore, that the voltage across it must be substantially equal to μV_g at even the lowest frequency in which we are interested, since it will certainly rise to within a fraction of this figure at the highest. For this, the inductance of the primary must provide a reactance which, even at a low frequency, is high compared with the anode resistance of the

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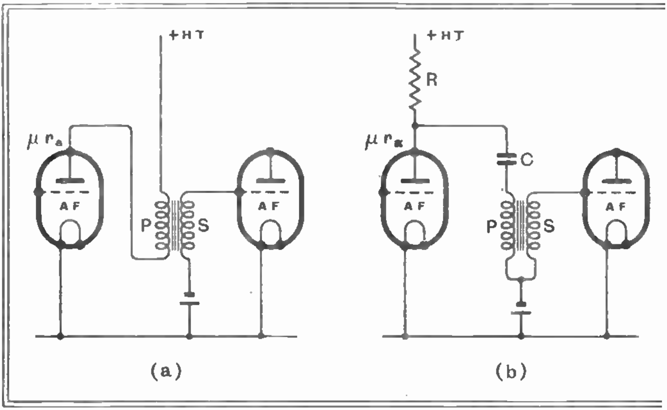

Fig. 142 : Transformer-coupled A.F. stages. In *a* the steady current of the first valv passes through the transformer primary P ; in *b* it is carried by R

valve. In the "equivalent anode circuit" of Fig. 143, th primary inductance L_p is in series with the anode resistance r of the valve, and receives $2\pi fL_p/\sqrt{r_a^2 + (2\pi fL_p)^2}$ of th generated voltage μV_g. If the primary reactance $2\pi fL_p$ i equal to r_a, the voltage across L_p will be 0·707 of what it is a a high frequency at which $2\pi fL_p$ greatly exceeds r_a. If w accept this condition as representing a tolerable drop in gai at the low frequencies we have at once a convenient desig formula : $r_a = 2\pi fL_p$.

For a given valve and transformer, this tells us the lowes frequency that is satisfactorily amplified ; for a 10,000-Ω valve and a transformer for which $L_p = 100$ henries, the dro to 70 per cent. of maximum amplification will occur at $f = r_a/2\pi L_p = 15{\cdot}9$ cycles. Evidently, with so good a transformer as this a valve o higher r_a, and hence higher μ, might b chosen. If we are content to set ou limit at 50 cycles, then, with the sam transformer : $r_a = 2\pi \times 50 \times 100 =$ 31,400 ohms. This, therefore, is th highest permissible value of valve resist ance. Or if r_a stays at 10,000 Ω, we ca use a less bulky transformer, for whicl L_p is given by $L_p = r_a/2\pi f = 10{,}000/2\pi \times 50 = 31{\cdot}8$ henries.

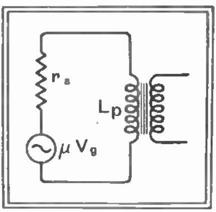

Fig. 143 : Equivalent anode circuit of a transformer coupled stage

It is important to note that the necessary value for the primary inductance is that which holds in actual use, with the steady anode current of the valve passing through the winding. The permeability of the iron core, on which the inductance depends, falls off very severely when the magnetising force due to the current in the coil exceeds a certain amount. This is described as magnetic saturation. So a large initial anode current tends to prevent the core from responding to the signal current, which has to superpose on it the varying magnetization from which the secondary derives its energizing voltage. In other words, the inductance is decreased below its "open-circuit" value by the steady current.

159. The Resistance-Fed Transformer

This effect can be allowed for by making sure that the minimum value of L_p prescribed by the formula is reached even with the steady current passing through the winding, or alternatively by diverting the steady current through another path, as in Fig. 142 *b*. Most modern transformers have cores of high-permeability material (Mu-metal, etc.) which attain magnetic saturation with quite a small primary current. For these the "parallel" circuit shown at *b* is essential. The feed-condenser, if large enough, has negligible effect on the voltage across L_p at any frequency, but by cunning choice of a suitable value for C it may be made to maintain the bass response of a transformer at frequencies lower than that to which it would respond satisfactorily with a condenser of infinitely large capacitance. In effect C and P form a tuned circuit, tuning flatly on account of R and r_a which are virtually in parallel across it, by which the extreme bass can be maintained. Instructions for the choice of C, R, and r_a are generally given in the instruction-slip accompanying a transformer.

The matter of high-note response from a transformer is a complex one, depending partly on the stray capacitance across the transformer—which should evidently be kept at a minimum—and on a transformer characteristic (leakage inductance) not usually known to the ordinary user. Owing to the lack of readily available data on this point, no discussion of high-note response will be embarked on here.

160. The Output Valve

When amplified sufficiently, the signal is passed from the last valve in the set to the loud speaker, there to move a diaphragm which recreates, with more or less fidelity, the sound-waves from which the original modulation was derived. To

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drive the diaphragm of a loud speaker *power* is required ; th output valve has therefore to be so chosen, and so worked, tha the greatest possible amount of power, consistent with th desired standard of fidelity of reproduction, is delivered to th loud speaker. To provide large power, high anode curren and high anode voltage are required ; an output triode i

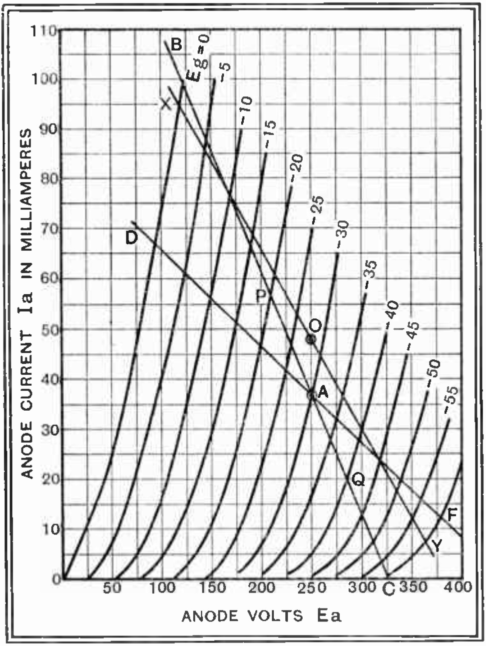

Fig. 144 : Curves of an output triode rated for 250 v. max. E_a. The load-lines shown are discussed in the text

therefore a valve of low anode resistance and may be rated t operate at voltages up to 400 or more.

The properties of an output valve are deduced, in much th manner already discussed in Chapter 7, from load-lines draw across the $E_a - I_a$ curves. A set of such curves for an outpu triode are reproduced in Fig. 144. In discussing a resistance coupled stage we saw that the load-line (Fig. 67) cuts the lin $I_a = 0$ at the voltage of the anode battery, thereby indicatin that the voltage at the anode of the valve could only rise t

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his value at zero anode current. In the case of an output alve the load consists of the windings of the speaker itself or f an output transformer, either of which has a comparatively ow resistance. If, for the sake of simplicity, we regard this esistance as negligibly low, the voltage at the anode of the alve will be that of the anode battery itself; and the curves of he valve-plus-loudspeaker combination, if measured with lirect current, will be those of Fig. 144. Let us suppose, then, hat we decided to work the valve at the rated $E_a = 250$ V, nd that we set the bias at -30 V. This gives the working point A, for which $I_a = 37$ mA.

Even though the speaker offers no resistance to D.C., it will have quite a large impedance to signal currents; if we consider his impedance as purely resistive and as having the same value or all the frequencies in which we are interested, we can epresent it by a load-line passing through A. Let us try, for start, a load of 2,000 ohms, represented by the load-line BAC.

161. Harmonic Distortion

By means of this diagram we can study the result of applying

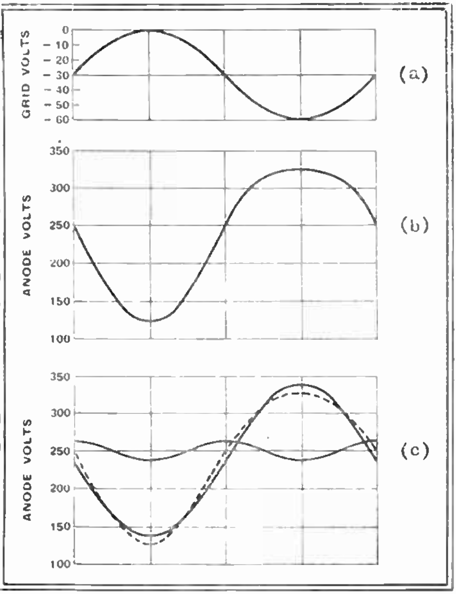

Fig. 145: Showing that the distorted output (*b*) resulting from applying the input *a* to the valve whose characteristics are given in Fig. 144 is equivalent to *c* (dotted line) the combination of an undistorted wave with its second harmonic

a signal to the grid ; say a sinusoidal waveform of 30 volts peak represented by Fig. 145*a*. As the grid bias is -30 volts, the signal swings the grid between 0 and -60 volts. Following downwards from the points where the load-line intersects these grid voltage lines, we can read off the anode voltage swing, 123 to 325 volts. By taking intermediate points from Fig. 145 *a* we can find on Fig. 144 the corresponding anode voltages, and so construct the output voltage waveform, Fig. 145 *b*.

This is obviously distorted ; the blunting of the positive half cycle is due to the crowding together of the characteristic curves near the bottom bend. In other words, the greater r_a in this region reduces the amplification, as one would expect from the formula $A = \mu R/(R + r_a)$.

Fig. 145*c* shows, in full line, an undistorted sine curve of the same frequency and mean amplitude as *b*, together with a smaller sine wave of twice the frequency. Adding these together we get the dotted curve, which is practically identical with *b*. *The distorted wave produced by this output valve is just the same as an undistorted wave plus a wave of twice the frequency.* This double-frequency wave is called the *second harmonic* of the original wave, which is the first harmonic or more commonly the *fundamental* ; and distortion of this kind, that introduces a second harmonic, is described as second-harmonic distortion. Other types of distortion bring in the third harmonic, and higher harmonics. The amplitude of the second harmonic in our example is about 13 volts, compared with about 100 volts fundamental. The amplifier is therefore giving about 13 per cent. second-harmonic distortion.

The ear tolerates a fairly large percentage of second-harmonic distortion, less of third-harmonic, and so on. A fraction of one per cent. of the higher harmonics such as the eleventh is unpleasant. Actually it is not so much the harmonics themselves that are disagreeable as the extra frequencies (called intermodulation products) that are generated when several frequencies are being amplified at once, as when an orchestra is playing. The condition—*non-linearity*—that causes harmonics also causes intermodulation (compare Secs. 140, 141).

Harmonic distortion is not confined to output stages ; but it is generally more pronounced there because the signal amplitude is greatest. If the amplitude is much reduced, so that the signal involves only nearly straight (or *linear*) parts of the valve characteristic curves, the distortion is negligible ; but the output is uneconomically small in relation to the power being put into the valve. It is therefore not a question of adjusting matters until distortion is altogether banished,

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because that can be done only by reducing the output to nil. It is necessary to decide how much distortion is tolerable, and then adjust for maximum output without exceeding that limit.

Fixing such a limit is no simple matter, because it should (in the case of reproduction of music, etc.) be based on the resulting unpleasantness judged by the listener. Different listeners will have different ideas about this, and their ideas will vary according to the nature of the programme. And, as has just been pointed out, the amount of distortion that can be heard depends on the order of the harmonic (2nd, 3rd, etc.) as well as on the percentage present.

Various schemes for specifying the degree of distortion in terms that can be measured have been proposed, but the nearer they approach to a fair judgment the more complicated they are; so the admittedly unsatisfactory basis of percentage harmonic content is still generally used. Usually the strongest harmonic only is included, generally second or third; and the most commonly accepted limit is 5 per cent. As we have seen, the distortion with a typical triode, in which the positive and negative half-cycles are unequally amplified, is almost entirely second harmonic. It can easily be worked out, on the lines of Fig. 145*c*, that a third harmonic sharpens or flattens both positive and negative halves equally.

162. Finding the Best Load

In Sec. 72 we saw that to get the greatest output from a valve (or any other generator) the load resistance had to be equal to the internal resistance (r_a). In Sec. 99 we found that for a valve transmitter it might be wise to modify this rule, which gives a theoretical efficiency of 50 per cent., in order to get a higher efficiency at some sacrifice of output power. Now, with an amplifier, we are limited to a certain percentage distortion, and the load resistance that gives the greatest output within that restriction may be far from equal to r_a.

It can be shown that when the distortion is all second-harmonic, the ratio of the lengths AB, AC (Fig. 144) representing the signal swing each side of the working point A, can be used to find the percentage, which equals $50 \times \frac{AB - AC}{AB - AC}$, which in the example shown amounts to about 13; and the output is (Sec. 87) 202 × 98/8 = 2,480 milliwatts. For 5 per cent. harmonic, the two lengths must be in the ratio 11 : 9. The greatest grid amplitude along the 2,000-ohm load-line within this restriction is about 15 volts, giving a total output

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voltage swing of 110 volts and current swing 53 mA, the output power therefore being $110 \times 53/8 = 730$ milliwatts only.

The restriction of the grid-swing made necessary by the early attainment of the 5 per cent. distortion limit indicates that the load has been wrongly chosen. Going through the same process of drawing load-line, investigating permissible grid-swing before the distortion-limit is reached, and calculating from the current and voltage swings the power delivered to the speaker, enables us to find the power that can be delivered into each of a series of loads of different resistance. The results are given as a curve in Fig. 146.

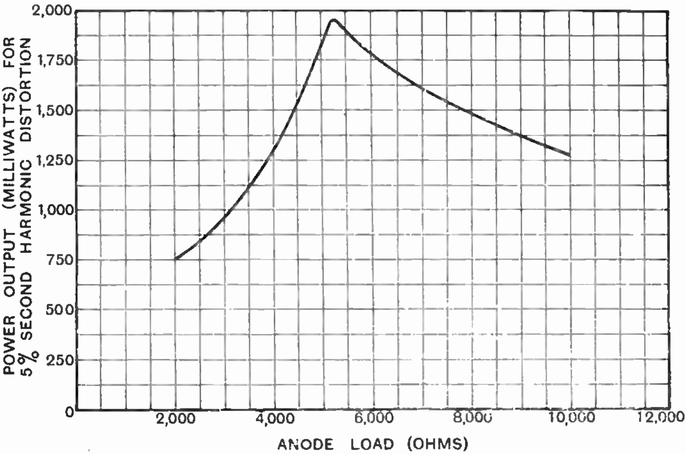

Fig. 146 : Relation between anode load and available power, allowing 5 % second-harmonic distortion, for the valve of Fig. 144 worked at point A

The *optimum load*, being that into which the greatest power can be delivered, is evidently about 5,200 Ω—the corresponding load-line is drawn at DAF on Fig. 144. To achieve this power the grid requires a signal that swings it from 0 to —60 V, giving a swing in anode current from $12\frac{1}{2}$ to 67 mA. The two excursions from A are exactly in the ratio 9 to 11, showing that distortion equivalent to the introduction of 5 per cent. second harmonic has just been reached. The power available for the loud speaker is now

$$\frac{(67 - 12\frac{1}{2}) \times (378 - 94)}{8} = \frac{54\frac{1}{2} \times 284}{8} = 1935 \text{ mW}$$

It will be remembered that the choice of A as the working-

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point was purely arbitrary—it is quite possible that some other point would give greater power. Still keeping to $E_a = 250$ V, which, being the highest voltage for which the valve is rated, will quite certainly give the greatest output within that rating,* other points can be investigated in the same manner as A, and then, by comparing the outputs given by the best load for each point, we can finally pick the best possible working-point and load. For the valve of Fig. 144 this is given by $E_a = 250$, $I_a = 48$, $R = 2{,}930\ \Omega$. For this, the available power is 2,670 mW, as can be deduced from the final load line XOY. It is necessary to make sure that the standing power put into the valve, $250 \times 48/1{,}000 = 12$ watts, does not exceed its *rated dissipation*, fixed by the amount of heat the valve can safely stand.

In general, the user of a valve is not compelled to go through this elaborate examination of valve-curves, for the makers' recommendations as to anode voltage and current, grid bias and optimum load are set forth in the instruction-slip accompanying each valve. The user has only to do as he is told.

This is just as well, because actually, although there is a middle range of frequencies over which a loudspeaker presents an approximately resistive load, at the extreme frequencies the reactance predominates, and this makes the matter too complex to deal with here. The resistive load line does, all the same, give a useful guide.

In the matter of providing the optimum load the user is rather at sea; he can do no more than ask the maker of his chosen loudspeaker to supply it with a transformer suited to the valve he proposes to use. The ratio of the transformer, as Section 48 shows, should be $\sqrt{\frac{R_p}{R_s}}$ where R_p and R_s are respectively the required load and the mean impedance of the speech-coil.

163. The Output Pentode

The ordinary tetrode is not suitable as an output valve owing to the distortion that would occur when the signal swung the voltage at the anode close to or below that of the screen (Sec. 143). But a pentode, or a kinkless tetrode (Sec. 145), which in what follows will be regarded as the same thing, can be used as an output valve.

Compared with the triode, the pentode offers the two advantages of being more *efficient*, in the sense that a greater pro-

* The power output given by a valve is related to the anode voltage applied thus: Power is proportional to $E_a^{3/2}$.

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portion of the power drawn by its anode circuit from the H.T. supply is converted into A.C. power for operating the speaker; and of being more *sensitive*, in that a volt of signal applied to its grid produces a larger output. For these two reasons the pentode has largely supplanted the triode as output valve for sets where cost is a prime consideration.

Screened and output pentodes differ in minor points, but not in principle. In the latter, since screening is no longer vital, grid and anode are both taken to pins in the base. High

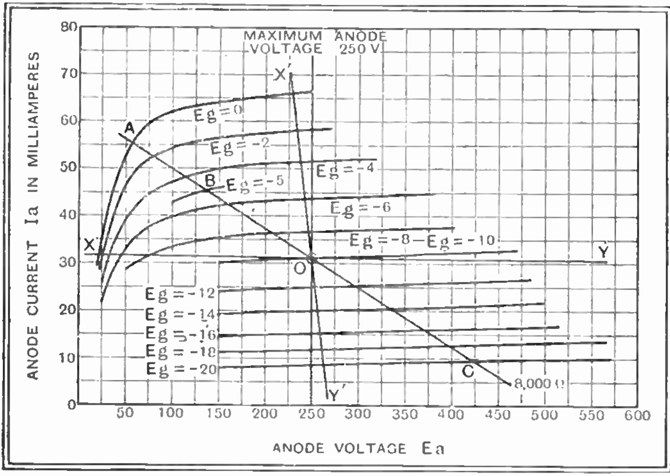

Fig. 147 : Curves of a typical indirectly heated output pentode. The load-line ABOC represents a usual load. Curves taken at E_s = 200

output is obtained by designing the valve to operate with a screen voltage little, if at all, below that at the anode.

In Fig. 147 are reproduced the curves of a typical indirectly heated output pentode; their similarity to the usable portion of the curves of a tetrode will at once be evident. We see again the high anode resistance (curves nearly horizontal) typical of valves using a screening-grid between control-grid and plate.

164. Loading the Pentode

The optimum load resistance for a triode is nearly always greater than its anode resistance; but for the tetrode or pentode is generally much less than the anode resistance. At the working point O (E_a = 250 V, E_g = − 10 V, I_a = 31 mA)

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the resistance of the valve is some 125,000 Ω (change in I_a of 2 mA brought about by change in E_a of some 250 V); XOY is a load-line representing 250,000 Ω drawn through O. Towards X it cuts the curves for $E_g = -8$ to $E_g = 0$ in very rapid succession, while towards Y it looks as though it will never reach the curves for $E_g = -12$ to $E_g = -20$. With a load such as this, the application of a signal swinging the grid from 0 to −20 would very evidently result in the most appalling distortion, together with the development of amazingly high audio-frequency voltages at the anode. (At what value of E_a does the line XOY cut the curve $E_g = -20$?)

If we were to fly to the other extreme and draw a load line (X′OY′) representing a very low load, distortion would again result, owing to the line now cutting the curves for high bias in very rapid succession, while the intercepts with the low-bias curves are widely spaced. Since these two types of distortion, for high and low loads respectively, occur at opposite ends of the total grid-swing, it is fairly evident that some intermediate load is going to be found best.

We are led to the same conclusion if we consider the power developed (still for the grid-swing 0 to −20 V) in the two loads. XOY offers high voltages and negligible current, while X′OY′ provides high current but negligible voltage. To get both voltage and current reasonably large an intermediate value of load is clearly required.

Let us investigate an 8,000 Ω load, which experience suggests as a possible load for a pentode. This is indicated by the line ABOC. The power delivered to this load when a signal swings the grid from $E_g = 0$ to $E_g = -20$ can be obtained, as with a triode, from the voltages and currents at the points A and C; it is

$$(424 - 56) \times (56 - 9)/8 = 368 \times 47/8 = 2{,}160 \text{ mW}.$$

165. Harmonic Distortion and the Pentode

How about distortion? With the triode, as we have seen, second-harmonic distortion predominates, and we accepted the convention that the permissible limit of this is 5 per cent. With the pentode we have to take into account distortion due to the introduction of both second and third harmonics of the original signal.

In Fig. 148 is plotted the dynamic characteristic (that is to say, a characteristic in which account is taken of the effect of the load impedance as well as r_a) of a triode working under conditions of 5 per cent. second harmonic; the data for this are taken from the load-line XOY of Fig. 144. To show up

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the non-linearity of the curve, a straight line joins its extremities; the divergence between the current at the actual working point O and that shown, for the same bias, on the straight line, is a measure of the second-harmonic distortion. Calling the currents at X and Y respectively I*max* and I*min*, that at P is midway between the two, or $\frac{1}{2}$ (I*max* + I*min*). The difference between this and I*o*, the actual current at O,

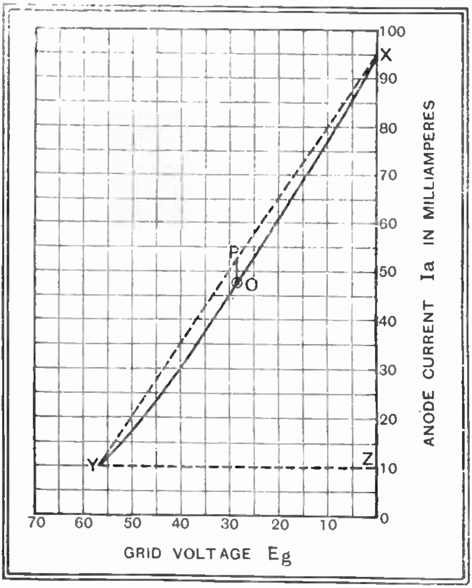

Fig. 148 : Dynamic curve of output triode giving 5% second-harmonic distortion, and, in dotted line, ideal characteristic for no distortion. Percentage second harmonic $= \frac{PO}{XZ} \times 100$

divided by the total current swing (I*max* — I*min*), gives the proportion of second harmonic, requiring only to be multiplied by 100 to give the percentage. The formula for calculation is thus :

$$\text{Percentage second harmonic} = \frac{\frac{1}{2}(I_{max} + I_{min}) - I_o}{I_{max} - I_{min}} \times 100.$$

This method is basically the same as that given in Sec. 162. To introduce third harmonic, as with the pentode, the curve must bend *both ways*, as in Fig. 149, which shows the dynamic

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curve of a valve introducing about 12 per cent. third harmonic, but zero second. Freedom from second harmonic is shown by the fact that O now lies on the straight line joining A and C, but it will be seen that the curve lies below the line between C and O, and above it between O and A. This particular type of divergence from linearity usually implies third harmonic. It can be numerically estimated in a similar way to second-harmonic distortion, using now only *half* the curve. It is

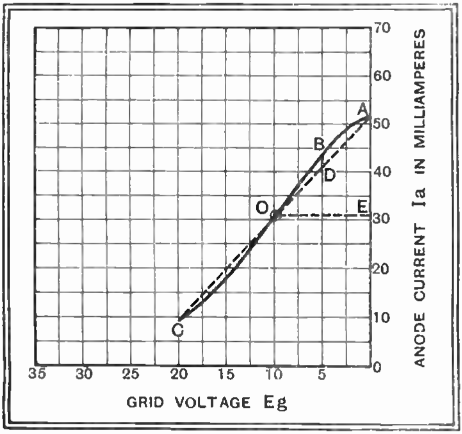

Fig. 149: Dynamic curve of pentode with 10,000-ohm load. Ideal characteristic giving zero second and third harmonics is shown dotted. Percentage third harmonic = $\frac{BD}{AE} \times 67$ approximately

approximately found from the difference between the actual current at B and the current at D (which, being on the straight line, is the mean between the currents at O and A), this difference being divided by the total change in current in passing from O to A, and multiplied by 67 to give percentage.

166. Relation Between Load and Distortion

By drawing a number of load-lines across the curves of Fig. 147 and calculating second- and third-harmonic distortion for each, the results summarized in the curves of Figs. 150 and 151 have been obtained. The difference between the two sets of data is that in making the calculations for Fig. 150 it was assumed that the signal had a peak voltage of 10 V, thus swinging the grid between zero and — 20 V, whereas in Fig. 151 the calculations have been made for an 8-volt signal, swinging the grid from — 2 to — 18 V only. As might be expected, the distortion is less for the restricted input.

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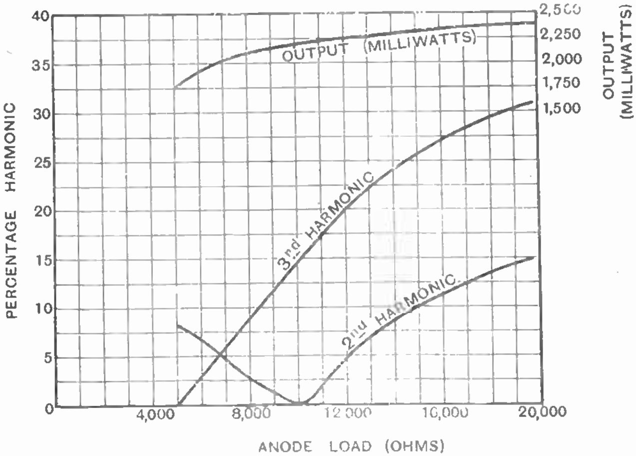

Fig. 150: Output and second and third harmonic distortion for pentode of Fig. 147
Working-point O ; input signal 10 V. peak

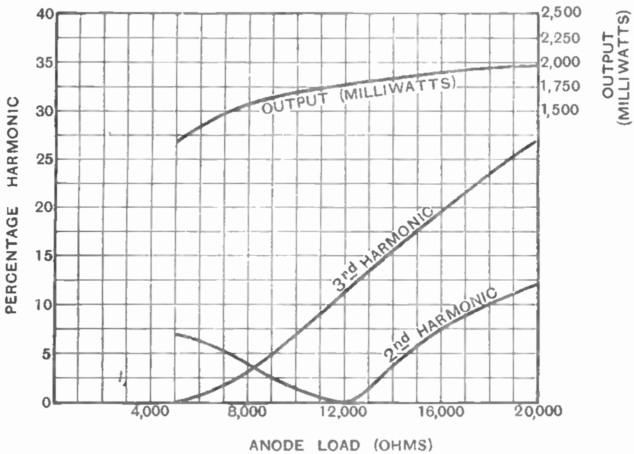

Fig. 151 : Output and second and third harmonics for pentode of Fig. 147.
Working-point O ; input signal 8 V. peak

In both cases the second-harmonic distortion is high for a low load, but drops away to zero as the load is increased. This is the load for which the dynamic characteristic has the form shown in Fig. 149. Still higher loads reintroduce second-harmonic distortion, which then rises rapidly with increasing load. Third-harmonic distortion, as the curves show, increases steadily with increasing load, as does the power delivered to the speaker. It is from a number of curves such as these, calculated not for one but for several alternative working points, that the final operating data for a pentode are determined by its designer.

167. Negative Feedback

The advantages of a pentode, which are high gain and high output power on moderate anode voltages, are to some extent

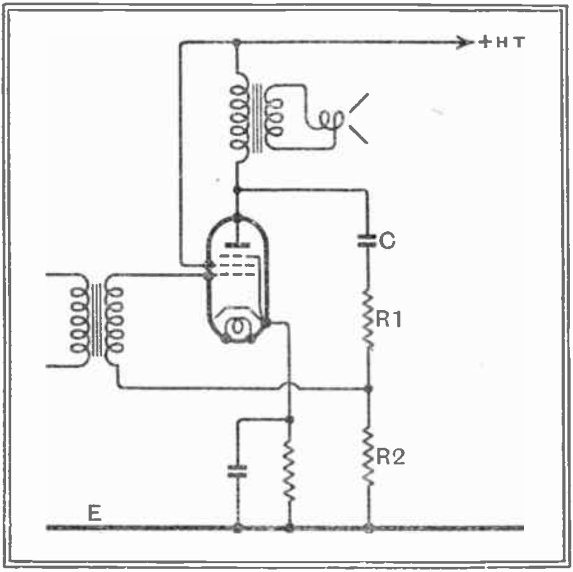

Fig. 152: Circuit for negative feedback. R_1 and R_2 form a potential divider across the output, the voltage developed across R_2 being fed back into the grid-circuit in series with the transformer secondary.

offset by the too-ready development of third-harmonic distortion, which (Sec. 161) is more objectionable than that associated with second harmonic.

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If a reduction in the gain of the valve can be tolerated, it is possible to decrease very considerably the proportion of harmonics in the output without decreasing the power available. This is done by feeding back into the grid-circuit a small proportion of the amplified voltage present at the anode, in the opposite polarity to that required to maintain oscillation (Sec. 80). It is therefore called *negative* feedback.

This can be done in any one of several ways, but it is necessary, in order to maintain the high input impedance of the valve, that the voltage fed back into the grid-circuit should be inserted in series, and not in parallel, with the original signal voltage. The circuit of Fig. 152 is a very suitable one for the purpose, the voltage fed back being that developed across R_2, the lower member of the potential divider across the output. C, of capacitance about 1 μF., serves simply to isolate R_1 and R_2 from the D.C. voltage at the anode of the valve. To avoid appreciable loss of output power, R_1 and R_2 together should have not less than about ten times the load resistance, and it is usually desirable to make R_2 about one-fifth to one-eighth of R_1, thus feeding back from one-sixth to one-ninth of the output voltage.

If A is, as before, the gain of the stage (i.e., ratio of anode voltage swing to grid swing) without negative feedback, and B is the fraction fed back in series with the input, then for every signal volt applied between cathode and grid, $-A$ volts appear at the anode, and $-AB$ volts are fed back. The input source is now required to supply, not 1 volt, but $1 + AB$ volts to produce the same output. The gain of the stage with negative feedback (which we might call A') is thus $A/(1 + AB)$.

Taking as an example the load-line AOC in Fig. 147, $A = \text{anode swing/grid swing} = 368/20 = 18{\cdot}4$. If one-fifth of this is fed back, $B = 1/5$, $AB = 3{\cdot}68$. $AB + 1$ is $4{\cdot}68$ (so the preceding stage must deliver $4{\cdot}68$ times as much signal voltage as if there were no negative feedback) and $A' = 18{\cdot}4/4{\cdot}68$ = very nearly 4. As the signal applied between grid and cathode is as before (the increase being solely on account of having to balance out the fed-back voltage), the valve does not actually handle any increased signal and so requires only its normal bias.

The harmonic-content of the output voltage can be shown to be reduced in about the same proportion as the gain of the stage, so that under conditions where a pentode would normally give a gain of 30 times, accept a signal of 5 volts, and deliver 2 watts of power with 10 per cent. harmonic distortion, the addition of negative feed-back might reduce the gain to six

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times, and make it necessary to supply a signal of 25 volts. This would result in 2 watts of power with only about 2 per cent. harmonic distortion.

Although the load required by the valve is unchanged by the introduction of negative feedback, its apparent resistance is enormously reduced. The new value of this is

$$\frac{r_a}{1 + \frac{r_a + \mu R_2}{R_1 + R_2}}$$

and with the normal values used for the circuit this amounts approximately to dividing the anode resistance of the valve by $\frac{\mu R_2}{R_1 + R_2} = \mu B$. As μ for an output pentode may be of the order of 600, and B may be about one-seventh, the anode resistance of the valve when feed-back is used is not far from one-hundredth of its normal value.

The practical value of this is in dealing with the fact that loudspeakers inevitably resonate at certain frequencies, giving excessive prominence to reproduction of them, as well as causing " ringing " or prolongation of the sound beyond that present in the original performance. By shunting the loud-speaker with a low resistance, such resonances can be flattened out or " damped " just like any tuned circuit. A triode output valve is such a low resistance, but a pentode is not, and an appreciable part of the poor quality of reproduction associated with pentodes can be traced to speaker resonances. By means of negative feedback applied to a pentode, its resistance effective for damping out loudspeaker resonances can be made even lower than in a triode without feedback.

With the addition of negative feedback, and at the cost of no more than a reduction in gain, a pentode gives as great freedom from speaker resonance as a triode and gives less than the triode's harmonic distortion, while retaining the high power-output and moderate bias of the pentode.

Care must be taken that the increased " drive " demanded by the grid of the output valve when negative feedback is used does not overload any preceding stage. To avoid this risk, and at the same time extend the benefits of feedback, the output may be fed back over more than one stage, perhaps from the secondary of the output transformer. A wide variety of circuits have been worked out to suit circumstances.

Pentodes are available giving the output of that illustrated by Fig. 147 with only about one-third of the drive, and so are particularly suitable for negative feedback.

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168. The Cathode Follower

In all valve circuits we have considered up to this point, the load has been between anode and H.T. supply. If, instead, it is connected between cathode and H.T. we get an extreme case of negative feedback, in which the *whole* of the output voltage is fed back. In its simplest form the circuit is as Fig. 153. The valve is not necessarily a pentode, and in fact we shall consider it to be a triode.

Suppose a signal of 1 volt to be applied between grid and cathode. This causes A signal volts to appear across R (as explained in Sec. 70). When the grid swings positive, it causes more anode current to flow, increasing the voltage drop across R, and therefore making the cathode more positive too. The A volts across R are therefore in series with the 1 volt between grid and cathode, making a total of 1 + A volts to be supplied by the driving source; 1 volt actually to drive the valve and A volts to oppose the A volts fed back across R.

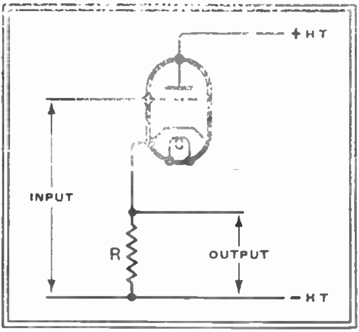

Fig. 153 : Simplest form of cathode follower circuit

This brief consideration reveals several features of the arrangement. The first is that since the input is 1 + A volts and the output is A volts, the voltage amplification of the stage as a whole must be less than 1, no matter what the characteristics of the valve and the value of R. The value of B in the formula given in the previous Section is 1, so the "gain" is A/(1 + A) as just found.

If A is made large by using a high-μ valve and high R, the output is very nearly as much as the input, but never quite as much. A stage that gives out a less voltage than is put in may not appear worthy of further attention; but it is too soon to jump to such a conclusion.

The next point is that as the "live" side of the output—the cathode—goes more positive when the grid is made more positive (and *vice versa*), the output is not an inversion of the input as in the anode-coupled amplifier, but is in the same phase.

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A third result follows from this; namely, that as the cathode signal voltage is normally very nearly the same as that on the grid, the signal voltage between them is small in relation to the input. It is this that earns for the device the name *cathode follower*; the cathode follows the changes of grid voltage. This is just the opposite of the demoniacal intervention explained with the aid of Fig. 116, in which the input capacitance of the valve is inconveniently multiplied. The demon is a friendly one this time, connecting the A-volt battery the opposite way round so as to reduce the charging current and *divide* the effective grid-to-cathode capacitance (C_{gc}) by $A+1$. On the anode side, the Miller effect is eliminated because the anode is at a fixed potential, so the grid-to-anode capacitance (C_{ga}) is just normal. The contrast can be illustrated by an example. Suppose the real C_{ga} and C_{gc} are each 5 pF and the amplification A is 40. Then with the load resistance in the anode side C_{ga} becomes multiplied by 41, making 205 pF, while C_{gc} is still 5pF (because the cathode potential is fixed). The total input capacitance is thus 210 pF. Transferring the resistance to the cathode side, C_{ga} remains 5pF, and C_{gc} is divided by 41, giving 0·12 pF. Total, 5·12 pF. This feature is of great value when amplifying a wide range of frequencies (such as in television signals) from a high impedance source, as any substantial input capacitance in the next stage would provide a low-impedance shunt at the high frequencies, cutting them down in relation to the low.

A fourth feature is the extreme reduction in apparent internal resistance of the valve. The formula given in the last Section stated that approximately this resistance is equal to $r_a/\mu B$ in a valve with negative feedback. In the cathode follower B is 1, so we are left with r_a/μ, which is equal to $1/g_m$. For instance, if g_m is 5 mA/V, which is 0·005 amps. per volt, the valve resistance is $1/0{\cdot}005 = 200$ ohms. This is notwithstanding that the valve may be a pentode with a r_a several hundred times greater.

Another way of looking at this is to consider what happens when the load impedance is much reduced, say by feeding into some low impedance. This causes the signal current to rise, which would not affect the voltage if the valve had no internal resistance. If r_a is very large, as in a simple pentode amplifier, the output voltage falls almost in proportion with the fall in load impedance. But in the cathode follower such fall in voltage reduces the voltage fed back, so that more of the input voltage is available for driving the valve, thus increasing the output. There is thus a strong compensating action tending

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to keep the output steady regardless of changes in load impedance ; which is just what is implied by the very low effective valve resistance mentioned above.

This, too, is a valuable feature when amplifying a wide range of frequencies and feeding them to some distant point via a line, which inevitably has a considerable amount of shunt capacitance. If the source had much internal resistance, this would cause the higher frequencies to be discriminated against ; but the resistance of a cathode follower is so low that the circuit it feeds into has little effect unless its impedance is lower still.

We have, then, a very useful device for feeding a low-

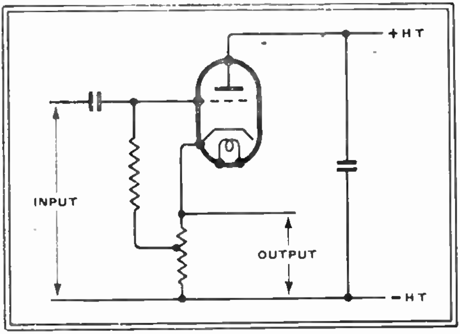

Fig. 154 : A more practical form of cathode follower, in which provision is made for reduced grid bias, and keeping the anode at a steady voltage

impedance load from a high-impedance source. The fact that there is a slight loss in voltage is outweighed by the great step-up in current. Compared with the cathode follower, a transformer is an impedance-changing device that sacrifices far more voltage in yielding a current step-up, is difficult to design for handling a very wide range of frequencies, and is much more liable to distort and to introduce stray coupling.

It is obvious that R in Fig. 153 incidentally provides grid bias (Sec. 209). If less bias is needed, it can be tapped off through a resistance too high to divert an appreciable proportion of the signal, as shown in Fig. 154. The grid condenser is to prevent a conductive input circuit from shorting out the effect of this bias connection. A very large capacitance is commonly connected across the anode supply as shown, to help keep the anode potential constant.

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169. Valves in Parallel and in Push-Pull

If more power is wanted than can be provided by a single output valve, two (or more) may be used. By simply adding a second valve in parallel with the first, connecting grid to grid and anode to anode, the swings of voltage at the anode are left unchanged, but the current swings are doubled. So, therefore, is the power, while the load resistance needed for two valves is half that needed for one. The performance of the whole output stage can be deduced from the $E_a - I_a$ curves of one of the valves merely by multiplying the figures

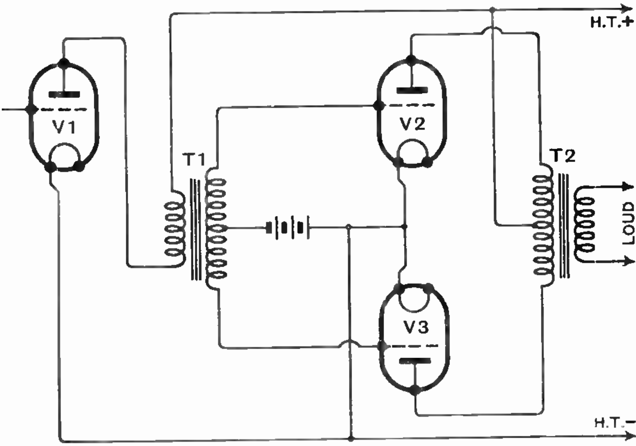

Fig. 155 : Two output valves, V_2 and V_3, in push-pull. The same circuit also applies to Q.P.P. and "Class B", the differences being only in the operating voltages and choice of valves

on the anode-current scale by the number of valves it is proposed to use.

Alternatively, the valves may be connected in *push-pull*, as shown in Fig. 155. Here the output valves are driven from a transformer T_1, in which the mid-point, instead of one end, of the secondary is earthed. At an instant when, with the normal connection, the " live " end of the secondary would be at $+20$ V, the other (earthed) end being zero potential, the centre-point of the winding would be at $+10$ V. With the push-pull arrangement this centre-point is brought to earth potential, the two ends, therefore, being respectively $+10$ and -10 V. Thus each valve receives half the available voltage, the two halves always being in opposite phase.

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The resulting out-of-phase anode currents, which would cancel one another if passed in the same direction through a transformer, are made to add by causing them to flow through separate halves of a centre-tapped primary, as shown at T_2 in Fig. 155. The voltage induced into the secondary, and hence the current flowing in the loud speaker, is due to the combined currents of the two valves.

This mode of connection has several advantages over the more obvious parallel arrangements. These are :—

(1) The steady anode currents, since they pass in opposite directions through their respective half-primaries, cancel one another so far as saturation of the core of the transformer is concerned (compare Sec. 158). A smaller transformer can, therefore, be used for two valves in push-pull than for the same two valves in parallel.

(2) Signals fed through the common H.T. connection cancel ; valves in push-pull are, therefore, unable to feed magnified signals into the H.T. line of a set, and so cannot give rise to undesired feed-back. Conversely, disturbances on the H.T. line (hum, etc.) cancel in the two valves.

(3) *Second*-harmonic distortion produced by either valve is cancelled by equal and opposite distortion from the other. Two *triodes* in push-pull will, therefore, give a greater undistorted output than they would if connected in parallel.

Third-harmonic distortion does not cancel in this way. Pentodes, whose output is limited by third harmonics (see Fig. 150), consequently give no greater output in push-pull than in parallel. Advantages (1) and (2), however, apply to pentodes as much as to triodes.

170. Phase Splitters

Owing to the difficulty of designing transformers to handle the very low and very high audio frequencies without distortion, and their high cost, and liability (as will be seen later) to pick up " hum ", the idea of using resistance coupling to drive a push-pull stage is very attractive.

Several circuits have been designed to carry this idea into effect, of which two examples will now be given.

In the ordinary resistance-coupled amplifier the resistor is on the anode side and the inverted output is taken from the anode. In the cathode-follower the resistor is on the cathode side, and the output taken from the cathode is " right way up ". If we had both of these outputs simultaneously they would be in opposite phase and therefore suitable for driving a push-pull pair. There is no reason why we should not have them,

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simply by splitting the coupling resistance into two equal parts and putting one on the anode side and the other on the cathode, giving the circuit shown as Fig. 156.

Because the same signal current passes through R_1 and R_2, the making of these resistances equal ensures that the two outputs are equal, or balanced—a matter of great importance in a push-pull system. To facilitate matching these resistances, the grid biasing arrangement (which will be dealt with in Chapter 19) is independent, and made so by shorting out the

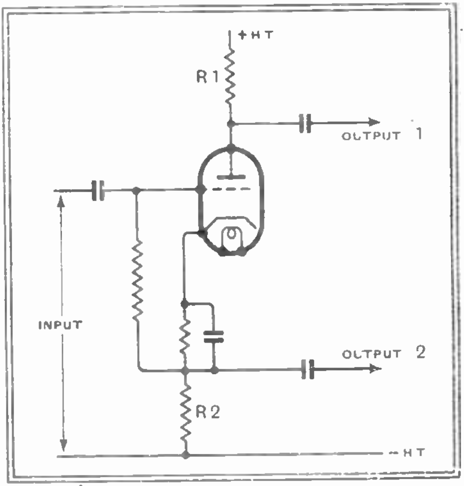

Fig. 156: The "concertina" circuit for providing two equal outputs in opposite phase for driving a push-pull stage

bias resistor to signal currents by means of a high-capacitance shunt condenser.

It will be seen that half the total output is fed back to the grid circuit, and therefore each of the two outputs is slightly less in voltage than the input (Sec. 168)

Looking at the circuit diagram, and imagining the two output points moving in and out in potential, the aptness of the name "concertina" circuit is clear.

The other system, called the "paraphase" circuit, is shown as Fig. 157. It requires two valves each of which is an ordinary resistance-coupled stage, capable of giving a considerable voltage gain. For the outputs to be in phase opposition it is necessary to drive V_2 in opposite phase to V_1, which can be

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done by driving it from the output of V_1. It is also necessary for the outputs to be equal, which means that the input to V_2 must receive only $1/A$ of the voltage output of V_1 (assuming the gain, A, of both valves is the same). This is done by tapping off the necessary portion of the anode resistor of V_1; this tapping is generally variable for adjusting the balance of the outputs.

Each of the two foregoing circuits has its advantages and disadvantages, which must be considered in relation to the design of the whole set in which it might be proposed to use them.

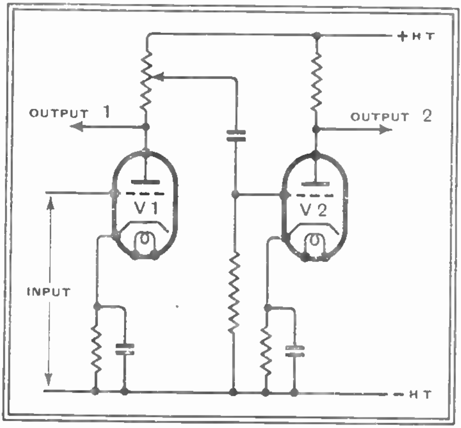

Fig. 157 : Another resistance-coupled push-pull driver — the "para-phase" circuit

171. Q.P.P. and Class B

In all the output stages considered so far in this chapter, the efficiency is at best only moderate, even when giving the maximum possible output within the limits of tolerable distortion. But in a broadcast programme that condition is—or should be—reached only rarely, during the extreme peaks of loudness. Most of the time the volume is far less, and the valves are called upon to handle no more than a few milliwatts; while during intervals and between words and sentences the output drops to zero. Nevertheless, the output valve continues to draw a steady anode current from the supply, representing in many cases a considerable number of watts. The average efficiency is therefore extremely small, possibly a fraction of one per cent.

In mains-driven receivers the cost of supplying this anode current is almost negligible; but in battery sets, where the power costs perhaps twenty to one hundred times as much, this waste is serious.

We have already seen (Sec. 88) how efficiency is improved by biasing the transmitter valve near or beyond cut-off, in Class "B" and "C" systems respectively. Receiver output stages stand to gain even more in proportion by such methods, because it is so seldom, if ever, during a normal programme that a large current drain is fully utilised. But adoption of Class B in a receiver, in the form illustrated by Fig. 80, would obviously lead to extreme distortion, with second and many other even harmonics strongly represented. In a transmitter this does not matter much, because the distorted anode current is used merely to jerk a resonant system into oscillation. In a receiver output there should be no resonant system, but one that follows faithfully the waveform applied. Class C would be much worse still, because all weak signals would be silenced entirely, and only the tips of the strong ones would get through.

In a push-pull system, however, the distortion—mainly second-harmonic—caused by over-biasing is balanced out. If the valves are biased to cut-off, one half of each signal cycle is amplified by one valve and the other half by the other valve. The centre-tapped output transformer puts these separately-constructed halves together again to form an undistorted whole.

If the valves would normally be biased to -10 volts, each would then require a 20-volt total grid swing, making the total swing on the transformer secondary 40 volts. Both valves would then amplify at every instant, and the standing anode current might perhaps be 20 mA per valve, remaining almost unchanged on applying the signal.

Now, suppose each valve biased to -20 V, and the signal doubled. The no-signal anode current might now be only 3 mA per valve, the two valves giving alternate kicks up to 40 mA when the full signal is applied. The *average* anode current, even when delivering full output, is less than in a normally biased push-pull stage; while if the applied signal is well below the maximum that the valves can handle, the average current, made up now of alternate kicks up to perhaps 6 mA, is quite small.

Such an arrangement, although fitting in with the general definition of Class B, in a receiver is generally termed *quiescent push-pull*, commonly abbreviated to "Q.P.P." Special valves

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are produced for the purpose, consisting of two pentodes in one " bottle ". Owing to the need for doubling the input signal, the less sensitive triode is not favoured for Q.P.P.

What is described as a Class B receiver output valve is actually slightly different, although it still uses the basic push-pull circuit of Fig. 155. No bias at all—or very little bias—is used, the anode current cut-off being obtained by special design of the two triodes comprising the Class B valve. As a result of the lack of bias, *the grids are swung heavily positive* by the signal. Grid current inevitably flows, thereby consuming audio-frequency power ; the preceding valve must therefore be so chosen that it can deliver this power without overloading, while the transformer feeding the Class B valve must be a properly designed " *driver* " transformer of the correct ratio and of low D.C. resistance. By removal of the no-grid-current limitation large powers can be obtained from a Class B output stage at the cost of a remarkably low average anode current.

As in the case of output stages of other types, many details of the performance of push-pull, Q.P.P., or Class B output stages can be obtained by careful study of the appropriate $E_a - I_a$ curves ; but the design of Class B, especially, is very complex, and likely to give poor quality and disappointing results generally unless many factors are taken into proper consideration.

172. The Loudspeaker

Whatever output stage is used, the amplified currents in the anode circuit of the last valve eventually reach the loudspeaker, the duty of which, as we have already seen, is to convert the audio-frequency currents into corresponding air-waves. More strictly expressed, it has to convert the audio-frequency electrical power

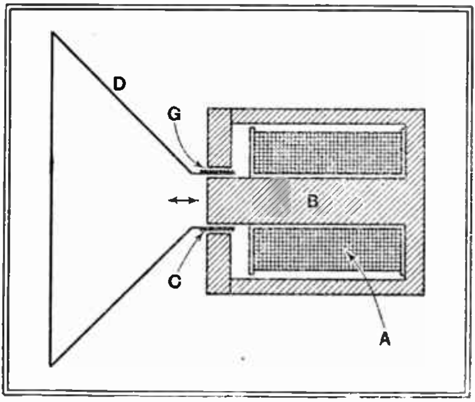

Fig. 158 : Cross-section of a moving-coil loudspeaker

supplied to it into acoustic power at the same frequency. As in every case where electrical energy is converted directly into mechanical energy, this is done by taking advantage of the magnetic field set up by the current.

Fig. 158 shows the cross-section of an energized speaker, in which the magnet is provided by passing a current through the winding A. The iron core, B, is extended by means of an outer iron shell, so that except for the small circular gap at G there is a complete iron circuit. The high permeability of the iron (Sec. 24) results in a very intense magnetic field. Alternatively a powerful permanent magnet, usually made of an alloy of nickel and aluminium, may be used instead of an electro-magnet.

In the gap is suspended the coil of wire C, wound on a former firmly attached to the diaphragm D. If we lead a current through C the coil will tend to move along the gap, driving D towards or away from the face of the magnet according to the direction of the current.

In the anode circuit of the output valve of a set receiving a tuning-note there is flowing an alternating current of frequency equal to that of the note. If C is connected in that anode circuit, it is driven in and out, as suggested by the arrow in Fig. 158, at the frequency of the current, and so the diaphragm, moving with it against the resistance of the air, converts into acoustic energy the power supplied by the valve. It thus sets up an air-wave conveying to the ear, at a loudness depending on the power in C, a note at the frequency of the current.

If the signal has the enormously more complex waveform of a piece of orchestral music, the movements of the coil, and hence of the diaphragm, still follow it—or would, in a perfect speaker—so faithfully reproducing that music.

It will be evident that at an instant when the diaphragm in Fig. 158 is moving to the left, there will be compressed air in front of it and rarefied air behind it. If the period of one cycle of movement of the diaphragm is long compared with the time in which the resulting air wave can travel round its edge from front to back, these pressures will equalize and no sound will be sent out. To prevent this loss, evidently worst at the lowest notes, the loudspeaker is always mounted so that it "speaks" through a hole in a *baffle*. This consists of a piece of wood, flat or in the form of a cabinet, designed to lengthen the air-path from front to back of the diaphragm and so to ensure that the bass is adequately radiated.

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CHAPTER 15

DESIGNING A RECEIVER

173. The Specification

BEFORE going on to consider the peculiar properties of the superheterodyne, it is proposed to devote a short chapter to the practical discussion of the design of a typical simple receiver, with the idea of making a kind of summary of the ground already covered. In order to help the reader to look up any points about which he may be doubtful, numbers in brackets refer him to the section in which fuller elucidation may be found.

We will suppose that we have been asked to design a set which will have an average sensitivity of about one millivolt. By this is meant that if a carrier-voltage of this magnitude, modulated to a depth of 30 per cent. (95) is applied to the aerial terminal, the overall magnification of the set will be such that the "standard output" of 50 milliwatts of modulation-frequency power will be delivered to the loudspeaker. The selectivity of the set is to be that associated with three tuned circuits—since their L/r ratio is bound to vary widely over the waverange covered (152) no numerical specification of selectivity is practicable. The whole is to be driven by batteries, and, for the sake of economy in upkeep, is to consume a maximum of 10 milliamps. in the anode circuits.

174. The Outlines of the Circuit

The first points to be settled are the type of output stage to be used, the kind of detector we shall choose, and whether the three tuned circuits shall be associated with one or with two radio-frequency amplifying valves. These points are inter-related and involve also the limitation in total anode current already imposed.

This latter limitation immediately suggests the choice of a quiescent output stage (Q.P.P. or Class B) (171), but also implies that a small-size H.T. battery is likely to be used. Now small batteries generally fail, except when new, to hand out the large instantaneous currents (171) demanded by quiescent output stages, and by so failing introduce very evident distortion. We will therefore play for safety and choose as output valve a pentode, on the grounds that it gives more output per milliamp. than does a triode (163).

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A battery pentode, if of the high-resistance type, takes about 5 mA at 120 V, in return for which it will deliver some 250 to 300 mW before overloading. This, though small, is an acceptable output for a set of the type contemplated. Allowing another milliamp. for the screen of the pentode, 6 of our available 10 mA are already accounted for.

With only three tuned circuits in the set it is quite certain that occasions will arise when the selectivity will not be adequate for separating the station required from others on neighbouring frequencies (148). In order that selectivity can be enhanced when desired, reaction will have to be available to the user (122). The use of fairly flatly-tuned circuits with adjustable reaction as an auxiliary will enable the inevitable selectivity-quality compromise (123 ; 149) to be readjusted by the user as he tunes from station to station.

For providing reaction the diode detector (105) is obviously useless. The anode-bend detector is not good from this point of view either, because it depends for its results on being operated on a part of its characteristic curve where the slope is very slight (113), and hence delivers comparatively little power to the anode circuit for reaction. We shall therefore choose a grid detector (109).

Either a screened pentode or a triode may be successfully used for this purpose, the former giving much the higher amplification. To set against this advantage it has so high an anode resistance (about 0·5 M Ω) that the use of a transformer to couple it to the output pentode is out of the question if we have any respect at all for our low notes (158). Shunting the transformer by a resistance (159) would limit the high-note gain to that available for low notes, but in so doing the gain would be reduced to about that of a simple triode. If we try to use resistance coupling, the voltage at the anode will be found to be seriously restricted by the voltage-drop in the resistance, and detector overload (109) will set an uncomfortably low limit to the available output, especially with deep modulation. To provide our output pentode with the signal (approximately 3 V peak) that it needs to develop full output, and at the same time to make reaction behave satisfactorily, it will be safest to choose a triode detector followed by a transformer of step-up ratio not less than one to three.

True, we shall now have serious input damping (120), which we could have avoided by choosing a screened valve, but reaction will take care of this (122). Unless a little reaction is used this input damping will make tuning rather flat, and sensitivity perhaps a shade disappointing. But by attention to

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tuned-circuit design this effect can be considerably reduced, as we shall shortly see.

To avoid all risk of overloading, even on low modulation, we shall hardly be safe if we allow the detector less than about 1 to 1½ mA of anode current—which, with the 6 mA of the output valves, leaves us 2½ to 3 mA for the R.F. side of the set. This is about the current of a single screened valve, but by biasing back we could keep the total current of *two* valves within this limit, and still have more gain than one valve could

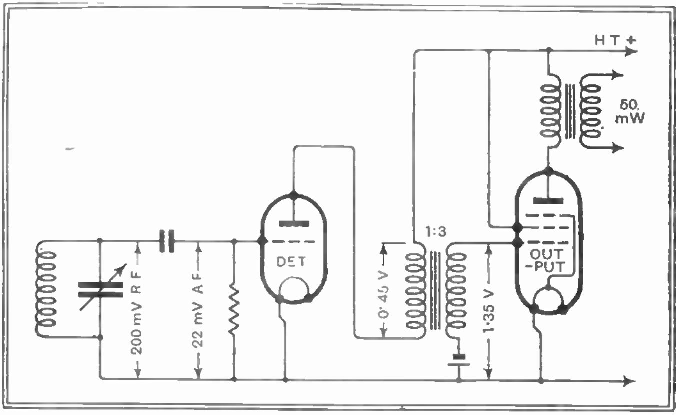

Fig. 159: Approximate evaluation of voltages on pentode and detector for 50 mW output

yield. What gain do we need? To find this we must work back from the output valve, as in Fig. 159.

175. The Amplification Required

We know from the makers' data that the pentode requires 3 V peak signal at the grid to give 250 mW output; to get our reference output of 50 mW we need, not one-fifth of 3 V (for that would give one-fifth current and one-fifth voltage, so one-twenty-fifth power), but $3/\sqrt{5} = 1 \cdot 35$ V peak across the secondary of T. Across the primary, assuming a 1 : 3 ratio because experience shows that to be a good compromise, we shall require 0·45 V. If the detector valve has $r_a = 20{,}000\ \Omega$, $\mu = 24$, under operating conditions, we can reckon on an audio-frequency gain of getting on for 20 times from grid to anode (158), so that we shall require a rectified signal, inside the grid condenser, of about 0·022 V or 22 mV.

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For so low an input as this implies, detector efficiency will be very low (113; 125), and, over-emphasizing this inefficiency so as to be on the safe side, we might reckon that 200 mV of carrier-voltage, modulated at 30 per cent., will be needed to produce a rectified signal of this magnitude.

This tells us that for a sensitivity of one millivolt we must have a radio-frequency gain of about 200 times between aerial terminal and detector grid. The gain given by one mains-driven valve, ignoring detector-damping, will be about 180 times (134) from grid of R.F. valve to grid of detector; a battery valve has a lower g_m and we shall do well to assume a gain of not more than about 60 times, so that we shall need some 3 to 4 mV at the first valve's grid. Across the second of two coupled circuits, the voltage is usually about four to eight times that actually applied to the aerial terminal, owing to the step-up effect of the tuned circuits (51; 118); we see, therefore, that 1 mV on the aerial terminal will comfortably give us the required 50 mW output with only a single R.F. valve, provided that, as assumed, reaction is used to an extent just sufficient to offset detector damping (122). We shall certainly not need a second R.F. valve; in fact, if we were to use one, the sensitivity of the set would be too high for its selectivity. By this is meant that the additional stations brought in by the extra sensitivity, being necessarily those which give only weak signals at the aerial, would all be liable to serious interference from stronger ones. Unless it were added simply with a view of making up for the deficiencies of a tiny aerial, the extra sensitivity would therefore be of no value in practice.

176. The Circuit Completed

Our set, then, will be arranged thus: two tuned circuits, R.F. valve, tuned circuit with reaction, grid detector, transformer, output pentode. Such a bald skeleton description as this does not prescribe an exact circuit; a dozen designers would produce a dozen circuits all differing from one another in minor ways. One of the many possible variations on the theme is shown in Fig. 160, where the complete receiver, including wave-band switching, is shown.

Careful inspection of this rather elaborate diagram will show that it really consists of a combination of separate circuits, each of which, regarded individually, is by now perfectly familiar. With but one or two unimportant exceptions, every separate circuit has been discussed somewhere or other in past pages. Dissection of the diagram is best performed by tracing grid, anode and screen circuits right through, starting at the electrode

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in question and continuing, through H.T. or bias battery, until the cathode of the valve is reached. Observe that sometimes the same components can be common to two circuits—for example, the tuned circuit $C_5L_7L_8C_6$ is included both in the anode circuit of V_1 and in the grid circuit of V_2.

Some small points in the circuit may be puzzling at first sight, even though their meaning could be seen by arguing from basic principles. The coupling of aerial to first tuned circuit is done by the combination of the primary winding L_0 and the condenser C_0, of capacitance about 20 pF. The two together, if suitably dimensioned, can be made to give more or

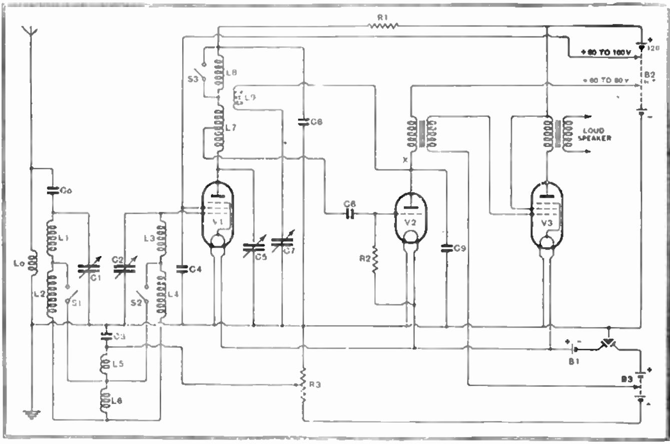

Fig. 160 : Complete circuit of three-valve set to conform with specification laid down in Section 173

less constant step-up at all frequencies on the lower (medium-wave) band. On long waves, S_1, S_2 and S_3 are open so that the tuning inductances in use are $L_1 + L_2$, $L_3 + L_4$, and $L_7 + L_8$. One section of each composite coil is shorted out for medium-wave reception.

Energy is transferred from the first tuned circuit to the second by making the coil $L_5 + L_6$ (on medium waves, L_5 only) common to both circuits (compare 130), so that the voltage developed across it by the current in the first circuit acts as driving voltage for the second. L_5 will need to be about 3 μH, while L_5 and L_6 together will be about 30 μH. The condenser C_3 is inserted to close the circuit for R.F. currents while allowing a variable bias, taken from the potentio-

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meter R_3 connected across the bias battery B_3, to be applied to the grid of the variable-mu screened pentode V_1 to control its amplification (18; 142).

The tuning condenser C_5 goes from anode to earth instead of directly across its coil L_7L_8 in view of the fact that C_1, C_2, and C_5 will normally be in the form of a three-gang condenser, with rotors on a common spindle. The tuned circuit is completed through the non-inductive condenser C_6, which, in order to maintain the ganging of the set, should have the same capacitance as C_3. Each may be 0·25 μF or over; much less would begin to reduce the tuning-range appreciably (38).

Since R.F. currents flow in the anode circuit of the detector, which is completed through the H.T. battery B_2, any R.F. voltage developed across this will be conveyed to the anode of V_1, and so to the grid of V_2. The resistance R_1, of some 5,000 Ω, serves as protection against instability (129) from this cause.

Damping imposed by the detector (107) on the tuned circuit is decreased, if only for medium waves, by connecting the detector grid to a tap on L_7. If the tap is at the centre of the coil, damping will be reduced to one-quarter (118). The reaction-coil L_9 is coupled to both L_7 and L_8, and the current through it is controlled by the variable condenser C_7. The inductance of the reaction coil must be such that C_7 does not tune it to any wavelength within the tuning range of the receiver, or reaction control will be difficult. The increase in sensitivity and selectivity (122) produced by applying reaction will also be felt in the circuit $L_3L_4C_2$, owing to a certain amount of energy feeding back through the screened valve and by way of stray couplings (128; 135).

As shown, the circuit does not include a radio-frequency choke in the anode circuit of the detector, the primary of the A.F. transformer serving as substitute. This attempted economy may lead to difficulty in obtaining proper reaction effects. Alternatively, by allowing R.F. currents to stray into the output valve, and then back, via loudspeaker leads, to the aerial side of the set, it may lead to hooting and grunting noises when receiving a signal, especially when much reaction is being used. In such cases an R.F. choke must be inserted at X making sure that the anode by-pass condenser of the detector (121) is still directly connected to the anode.

As shown, the set requires three positive connections to the H.T. battery. This enables the technically-minded user to adjust the voltages at detector anode and pentode screen either for maximum sensitivity or for economy of current. In a

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commercially-built set, to be handled by non-technical users, it would be better to provide a resistor of fixed value in each of the movable leads and to take them all to maximum H.T. voltage.

It is hoped that this chapter has given the reader a glimpse of the way in which all the various matters discussed in earlier parts have to be brought together when considering the design of a set, and of the process by which a concrete design emerges from a brief specification of intended performance. Any reader who may be taking this book really seriously may like to complete the design here only begun ; by a sufficiently close study of earlier chapters he could find a suitable value for every component in the set, after which, adding some data from a valve catalogue, he could work out, at least approximately, the overall sensitivity, selectivity, and fidelity of the receiver at a number of different wavelengths. It will have been gathered, however, that at many points the designer falls back on experience in selecting a particular value or component. A design arrived at purely theoretically may not be realizable in practice ; on the other hand, guesswork and rule of thumb are not enough. The good designer is the one who knows how and when to combine both.

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CHAPTER 16

THE SUPERHETERODYNE AND ITS FREQUENCY-CHANGER

177. The Need for Selectivity

SOME idea of the problem of separating one station from another when they are spaced only 9 kc/s apart, without losing the sidebands of the wanted station, will have been gained from Chapter 13. The problem is intensified by the demand of the listener to be able to hear a station undisturbed even when its next-door neighbour (in frequency) is many times more powerful than itself. It is complicated still more by the fact that selectivity varies as the tuning condenser is adjusted.

In the usual " straight " set, as discussed in the last chapter, pre-detector amplification is carried out at the frequency of the signal. So long as we have only two or three circuits to be retuned every time we pass from one station to another, this system is convenient enough, but if we were to demand selectivity of so high an order than ten tuned circuits were needed to provide it, the set would become impossibly cumbersome.

178. The Principle of the Superhet

When high selectivity in conjunction with simplicity of control is required, the supersonic heterodyne receiver (conveniently known as the " superhet.") is the only possible type of set. In Fig. 161 is given a block diagram of a superhet., in which the various sections of the set are shown as labelled boxes. Of their contents we shall speak later.

The signal received from the aerial is first put through a stage of *pre-selection*, containing tuned circuits enough to ensure that signals of wavelengths far removed from that of the station required shall not pass farther into the set. This box may or may not contain a stage of ordinary radio-frequency amplification of the type with which we are now familiar.

The next stage, the *frequency-changer*, operates upon the signal in such a way as to produce a carrier wave of a new frequency, this new carrier still carrying the modulation of the original carrier. In most cases the new carrier has a frequency lower than that of the original signal, though it is always *supersonic*, or higher than any frequency within the audible range. It is, in consequence, usually referred to as the *intermediate frequency*, commonly abbreviated to I.F. At

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this new frequency it undergoes further, and often considerable, amplification in the third box of Fig. 161, after which it is passed to detector and A.F. amplifier in the ordinary way. Pre-selection, frequency-changing, and I.F. amplification thus take the place of the R.F. amplifier and associated tuned circuits of an ordinary set.

Whatever may have been the frequency of the received signal, when it has been tuned in it always emerges from the frequency-changer at the one fixed intermediate frequency, this being determined by the designer of the set. To perform this conversion, the frequency-changer subtracts from (or adds to) the signal frequency whatever frequency is necessary in order to bring it to the I.F. As the signal frequency varies according to the station it comes from, whereas the I.F. is fixed, it follows that the frequency to be subtracted must be varied to suit the

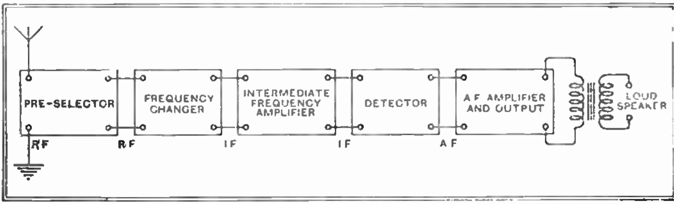

Fig. 161 : Block diagram of supersonic heterodyne receiver (" superhet.") showing the functions of the various parts and the changes in the signal in passing through i

station it is desired to receive, and therefore the frequency-changer must be tuned to subtract the right frequency. It is so tuned that a 1,000 kc/s signal has its frequency converted to 465 kc/s (by subtracting 535 kc/s); and signals at 991 and 1,009 kc/s, due to stations working on *adjacent channels* in the frequency band, are also present; the frequency-changer converts them to 456 and 474 kc/s respectively. It follows that if the I.F. amplifier is accurately and selectively tuned to 465 kc/s, these two signals will not pass through it, and hence will not reach detector, A.F. amplifier, or loud-speaker.

Adjacent-channel selectivity, or selectivity aimed at removing stations on frequencies closely bordering on that of the desired station, can therefore be provided entirely by design of the I.F. amplifier without reference to any other part of the set.

Since the I.F. amplifier is tuned to the one fixed frequency, it becomes practicable to include in it just as many tuned circuits as are needed to provide the selectivity we require; they have only to be tuned once, when the set is first made.

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Nor is this the only advantage of operating on a fixed frequency ; by careful and finicky adjustment we can shape the overall resonance-curve to give us any desired compromise between selectivity and sideband response with the comforting knowledge that this compromise will hold unchanged for every station received. Further, its constancy permits of judicious faking of the A.F. amplifier to strengthen high notes if we find that we cannot get the selectivity we desire without undue cutting of side-bands in the I.F. tuned circuits.

Although the highest usable adjacent-channel selectivity can be provided in the I.F. amplifier, tuning is still required in the pre-selector stage. This is so because the characteristics of the frequency-changer are such that stations *on certain frequencies widely removed* from that of the station required can set up in it a carrier of the intermediate frequency, so causing interference with the station to which the set is intended to be tuned. It is the duty of the pre-selector to eliminate these outlying frequencies before they can reach the frequency-changer, leaving the task of providing adjacent-channel selectivity to the I.F. amplifier.

179. The Frequency-Changer

The function to be performed by the frequency-changer, as we have just seen, is to convert all the signal frequencies into new frequencies, of which only those coming from the desired station must be allowed to get through the I.F. amplifier. Suppose a sample of the carrier wave coming from the desired station in one hundred-thousandth of a second to be represented by Fig. 162 *a*. As it contains 10 cycles, the frequency is 1,000 kc/s. If another signal generated in and by the frequency-changer itself is as shown at *b*, its frequency is 1,450 kc/s. Now add the two together. The result is shown at *c*, and can be seen to be a 1,450 kc/s wave varying in amplitude at a rate of 450 kc/s. But it is only a variation (or modulation), at 450 kc/s, of a signal of higher frequency. *No signal of* 450 *kc/s is present*, for each rise at that frequency above the centre line is neutralized by an equal and opposite fall below the line. The average result, when applied to a circuit tuned to 450 kc/s, is nil.

We have been up against this difficulty already—see Fig. 94—and the solution is the same now as then ; rectify it. By eliminating all the negative half-cycles (*d*) the smoothed-out or averaged result of all the positive half-cycles is a 450 kc/s signal—the difference between the frequencies illustrated at *a* and *b*—shown dotted at *d*. This mixture of frequencies is passed to the I.F. tuned circuits, which reject all except the 450 kc/s, shown now alone at *e*. The unwanted frequencies

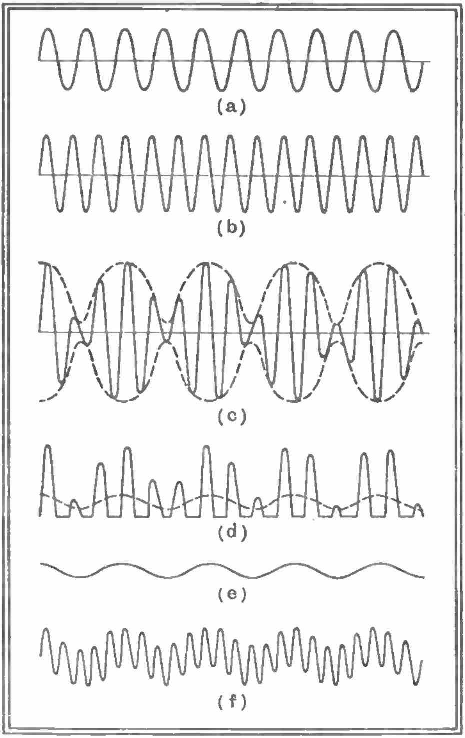

Fig. 162 : A sample of incoming carrier wave, of ten millionths of a second duration, is represented at *a*. To the same scale *b* is the local oscillator signal. When the two are added (in a "first detector") the result is c, which must be rectified (*d*) to yield a difference frequency (e). But if *a* and *b* are *multiplied* together (*f*), the difference frequency comes directly

are so far away from that to which the I.F. circuits are tuned that they are soon eliminated.

So far, the result is only a 450 kc/s carrier wave of constant amplitude, corresponding to the constant amplitude of the 1,000 kc/s carrier wave (*a*) received from the aerial. It is not

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difficult to see, however, that if *a* is modulated the modulations are repeated in *e*; so long, at least, as *a* does not exceed *b* in amplitude. Looking at the modulation of *a* as the addition of sideband frequencies (Sec. 124), these sidebands when added to *b* and rectified give rise to frequencies that are sidebands of *e*. Suppose that our 1 kc/s tuning note is broadcast: in addition to the 1,000 kc/s at *a* there are 999 kc/s and 1,001 kc/s. When added to 1,450 kc/s the 450 kc/s beat itself beats at 1 kc/s, and when rectified there are present 449 kc/s and 451 kc/s. As the I.F. amplifier is designed to cover a band of several kc/s each side of 450, these sidebands are amplified along with the carrier wave; and so far as the following detector is concerned the combination might equally well have been due to a station transmitting on a carrier frequency of 450 kc/s.

The early types of frequency-changer, not commonly used now except for extremely high frequencies, worked on the principle just described. In some, the functions of mixer and rectifier (or, as it was generally called, *first detector*) were allocated to one valve and that of oscillator to another; in others, a single valve did everything.

180. A Two-Valve Frequency-Changer

An example of the former type is shown in Fig. 163. V_2 is the oscillator which generates the added frequency. The

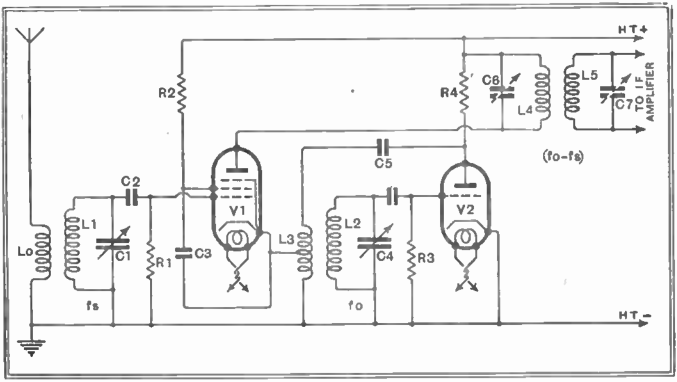

Fig. 163: Circuit of simple two-valve frequency-changer. The signal-frequency *fs* is applied to the grid of V_1, and the oscillator V_2 is tuned to *fo*. Currents at the intermediate frequency (*fo* − *fs*) appear in the anode circuit of V_1

resistor R_4 serves in lieu of an R.F. choke to divert the radio-frequency anode current through the reaction-coil L_3 besides

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being useful in limiting the average anode current of the valve. Further help in this direction is supplied by grid rectification of the oscillation, which biases V_2 negatively (Sec. 105). The frequency of the oscillation is that to which the tunable circuit L_2C_4 is adjusted ; for convenience of reference we will call this, the oscillator frequency, *fo*.

The signal, of frequency *fs*, is collected from the aerial or other source by the tuned circuit L_1C_1, and applied to the grid of V_1, the screened pentode used as first detector.

The cathode of V_1 is taken to a tapping on the reaction coil L_3, thereby including that part of L_3 that lies between tap and earth in the grid circuit of the valve (remember that the grid circuit includes everything between grid and cathode). The amplitude of the oscillation thus applied to the grid of V_1 will require to be about 10 to 15 V peak in a circuit of this kind ; suitable choice of tapping point on L_3 ensures a correct voltage. Like the oscillator, V_1 will bias itself back until the applied oscillation just, and only just, runs the grid into grid current. Assuming a 10-volt peak oscillation at this point (represented by *b* in Fig. 162), the bias of V_1 is being swung, at the frequency of the oscillation, from zero to -20 and back again.

The negative peaks therefore occur in a region of the valve's characteristics at which anode current is almost if not entirely cut off. The incoming signal, which relatively is far smaller than suggested by Fig. 162 *a*, is meanwhile causing both positive and negative peaks to fluctuate at the intermediate frequency ; though far less violently than shown at *c*. As the negative peaks are cut off, only the positive ones are effective in varying the anode current ; and so we get *d*, from which the I.F. amplifier selects *e*.

181. Modern Frequency-Changers

The early types of frequency-changer, especially those in which one valve was made to do everything, fell short of the ideal requirements :

1. Maximum I.F. output for a given R.F. input.
2. Absence of radiation from oscillator.
3. Frequency stability of oscillator.
4. Ability to oscillate freely at very high frequencies.
5. Absence of " pulling " between oscillator and pre-selector circuits.
6. Application of volume control, and especially A.V.C. (Chapter 18).

All of these requirements became much more difficult to fulfil

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as the demand arose for reception of very high frequencies. One of the most tricky to deal with is 5. The I.F. in vogue during this early period in the development of the superhet was 110 kc/s. This represents a difference between received signal and local oscillation frequencies of 55 per cent. at 200 kc/s (long wave) or 11 per cent. at 1,000 kc/s (medium wave), which causes no great difficulty; but at 20 Mc/s (15 metres, short wave) the difference is only about 0·5 per cent. Any slight coupling due, say, to stray capacitance, between the oscillator and signal preselector tuning circuits tends to cause the latter to be pulled into step with the former, and to provoke other undesirable effects. As the products of both tuning circuits must be applied to one valve in order to combine them, the interelectrode capacitance must be reduced to a very small figure indeed if an undesirable amount of coupling is not to result; and even if it is eliminated by suitable screening there are other more subtle forms of coupling that give trouble. The situation was relieved by a general adoption of an I.F. in the region of 450-470 kc/s, increasing the disparity between the two frequencies.

As this alone did not solve all the problems, intensive work by valve designers has resulted in an almost bewildering variety of frequency-changers, some with separate oscillators but most combining all functions in one bulb. It is not possible in a limited space to deal with all of these, but one outstanding feature is a departure from the principle outlined in Sec. 179, known as the *additive* principle because the incoming signal and the local oscillation are added together and then rectified in order to yield the desired difference frequency. Nearly all modern frequency-changers are of the *multiplicative* type, in which no rectification is needed, the difference frequency being produced directly. Look again at Fig. 162: curve *c* was formed by adding the heights above or below the centre lines of *a* and *b*. If these heights are multiplied instead of added, remembering that two negative heights when multiplied give a positive height, the result is curve *f*. The difference frequency, 450 kc/s, is here already without rectifying; and so is the sum frequency, 2,450 kc/s, which could quite possibly be used as the I.F. instead of 450 (there are as many crests in *f* as in *a* and *b* together).

But how does a valve multiply? It is easy enough to add two signals in a valve by feeding them in series between grid and cathode so that the net grid-cathode voltage is the sum of the two.

Sec. 134 showed that the amplified output of a valve of the

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high anode resistance type is practically equal to the input voltage multiplied by $g_m R$. R is the dynamic resistance of the output tuned circuit, which in a superhet is the first I.F. transformer, so is fixed. If g_m, the mutual conductance, can be made proportional to the local oscillation voltage, then the output is proportional to the input signal voltage multiplied by the oscillation voltage at every instant.

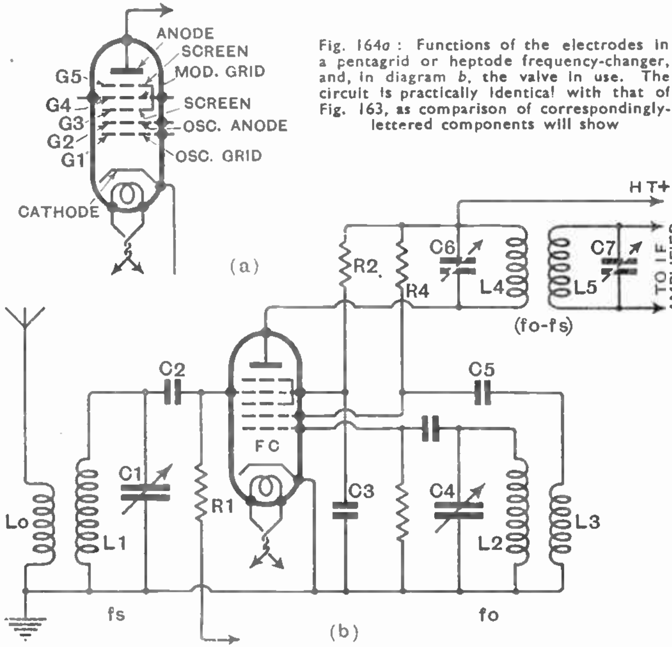

Fig. 164*a*: Functions of the electrodes in a pentagrid or heptode frequency-changer, and, in diagram *b*, the valve in use. The circuit is practically identical with that of Fig. 163, as comparison of correspondingly-lettered components will show

The first successful valve of this type was the heptode or pentagrid, shown in Fig. 164.

As its name implies, the valve has five grids, the uses of which are shown on the diagram. G_1 and G_2 form the grid and anode of a triode oscillator, the circuit of which, as Fig. 164 *b* shows, in no way differs from that of V_2 in Fig. 163. G_4 and G_5 serve as control grid and screen of a screened tetrode performing the functions of V_1 in Fig. 163. The additional grid G_3, connected within the valve to G_5, serves to screen the modulator grid from the oscillator, and so prevents

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G_4 from biasing itself back, as does V_1 in the two-valve circuit. This valve, therefore, remains responsive to control of amplification by variation of bias ; G_4 is consequently given variable-mu characteristics, and the controlling bias is fed to it through the resistance R_1.

The almost exact identity of the two frequency-changing circuits is emphasized by the fact that exactly the same components are used in both ; for convenience, they have been identically lettered in the two diagrams. The parallel can be made even closer by replacing the pentagrid with an *octode*, for in this valve there is yet another grid between G_5 and the anode, thus converting the tetrode outer portion of the pentagrid into a screened pentode.

The sole real difference between the two lies in the method of arranging that the oscillation shall vary the mutual conductance of the screened valve that deals with the signal.

Every electron that reaches the modulator (made up of G_4, G_5, and the anode) has to pass *through the oscillator* (G_1 and G_2) on its way.

The slope of the modulator is low when the oscillator grid is strongly negative, and high when its potential is zero or slightly negative. When oscillations are present on G_1, this grid will bias itself back until only the extreme positive peaks cause grid current to flow ; the total excursion of the grid will therefore be from approximately zero to double the peak voltage of the oscillation. It is therefore evident that we have drastic variations of modulator slope at the frequency *fo* of the local oscillations generated by the triode portion of the valve. Since the incoming signal, at frequency *fs*, is applied to the grid of the modulator, we have a system in which the amplitude of the original signal is in effect multiplied by that of the local oscillator, leading to the production of sum and difference frequencies in the anode circuit. Of these, that desired is picked out by the tuned circuit L_4C_6, and passed, through a second tuned circuit L_5C_7, to the I.F. amplifier valve.

Actually, as the slope of a valve is not a thing that can have negative values, the waveform shown in Fig. 162 *f* is somewhat different in practice, the difference consisting in the presence of the original signal frequency along with the others, but that is also rejected by the I.F. amplifier.

Although there is screening within the heptode between the oscillator and preselector tuning circuits, ordinary capacitance is not the only form of coupling possible. The large fluctuating mob of electrons controlled by the oscillator section constitutes a *space charge* (Sec. 63), which induces a current at oscillator

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frequency in the preselector circuits. This effect becomes serious at the higher frequencies and, together with other still more obscure phenomena, causes the heptode to be largely ineffective at over 20 Mc/s. Some of the later types of octode have, by skilled design, been rendered useful at these very high frequencies; but the most popular frequency-changer at the present time is the *triode-hexode*. Here we have virtually two valves, only the cathode being common to both—in point of

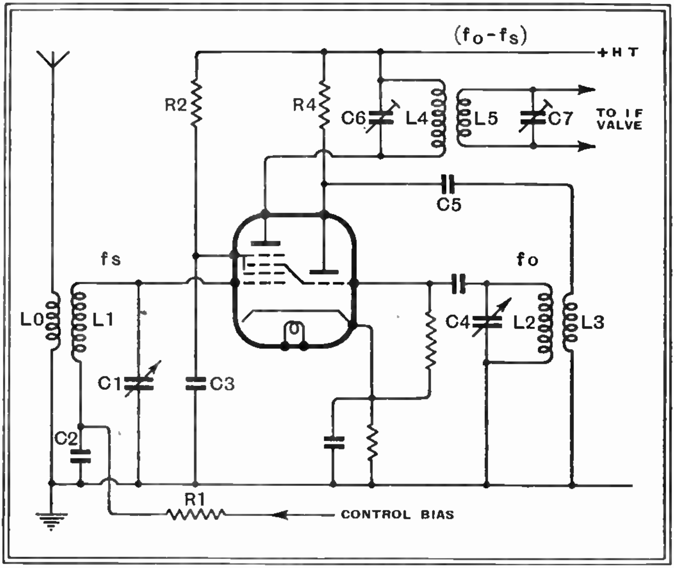

Fig. 165 : Typical triode-hexode frequency-changer circuit, lettered to correspond with Figs. 163 and 164*b*

fact, a separate triode and hexode are used in some sets—and the noteworthy feature is that the signal control grid comes first, and the really large movements of electrons, produced by the local oscillation, occur beyond, where they can exert very little undesirable influence on the preselector circuits.

Fig. 165 shows the connections, and again the lettering of the components is the same. A cathode resistor provides a small amount of initial bias for the signal, or modulator, grid. The grid of the oscillator section is internally connected to the oscillator injection grid (No. 3 in the hexode), which is screened from both signal grid and anode by screen grids. With some valves it is preferable for the oscillator anode to be tuned, rather than the grid. It is possible to give the triode section a higher mutual conductance than in the heptode, enabling oscillation to be readily obtained at higher frequencies in spite of the various influences that tend to increase losses at those frequencies ; and the other requirements specified at the beginning of this Section are also generally better met by the triode-hexode frequency-changer.

182. Conversion Conductance

In the ordinary amplifying valve the mutual conductance is expressed in terms of milliamps of signal current in the anode circuit per volt of signal applied to the grid. The same rating can be applied to a frequency-changing valve, but it is not of much help in receiver design. In this particular case we are interested in milliamps of current *at intermediate frequency* per volt of signal (radio-frequency) on the grid. This is known as the *conversion conductance* of the valve, and quite evidently depends on the efficiency of conversion as well as on the amplifying abilities of the modulator.

Looking at Fig. 162, it can be seen that the peak fluctuations in the amplitude of *b* due to *a*, indicated in *c* by the dotted lines, are equal in amplitude to *a*, on each side of *c*. The fluctuation in *mean* amplitude, shown by the dotted line in *d*, which represents the I.F. output, is less. With full-wave rectification, it would be $\frac{2}{\pi}$ or 63 per cent. of the peak value. But half-wave rectification leaves gaps between each half-cycle, and this figure must be halved, giving $\frac{1}{\pi}$ or 37 per cent. Assuming perfect rectification, then, the conversion conductance with additive frequency-changing is a little over one-third of the mutual conductance. If the oscillation amplitude is insufficient, rectification is imperfect, and conversion conductance falls off, reaching zero, of course, when the oscillation amplitude is zero. On the other hand, if the oscillation amplitude is excessive, only the extreme peaks will pass anode current, there will be bigger gaps between them, and the mean value and consequently the conversion conductance will fall, though not so rapidly. There is therefore an optimum oscillation amplitude for every valve.

In Fig. 162*f* the amplitude of the I.F. component is half that of *a*, so the maximum conversion conductance is half the

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mutual conductance in the multiplicative type of frequency-changer, and therefore rather better than in an additive type having the same mutual conductance. The condition for optimum oscillation amplitude is that it shall be just sufficient to vary the mutual conductance from maximum to zero.

Fig. 166 illustrates this for a heptode, and it is evident that

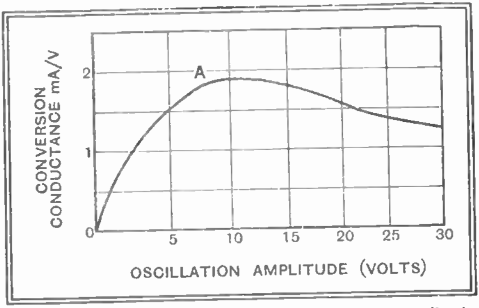

Fig. 166 : Type of relationship between oscillation amplitude and conversion conductance of frequency-changer. The oscillation amplitude is not critical so long as it exceeds a certain value in the neighbourhood of A

there is less danger of losing gain by too powerful than by too weak an oscillation. It is usual, therefore, to arrange that at no part of the wave-band to be covered by the set shall the oscillation amplitude fall below a value corresponding to a point at or near A on the curve. This is done by adjustment of turns on the reaction coil (L3, Figs. 163, 164 and 165), after which the oscillator can be left to look after itself.

183. Ganging the Oscillator

We have seen that the intermediate frequency is in all usua cases equal to the difference between the signal frequency anc the oscillator frequency. With an I.F. of 450 kc/s the oscillatoı must therefore be tuned to a frequency either 450 kc/s greate or 450 kc/s less than the signal. If the oscillator frequency *f*(is higher than the signal frequency *fs* the intermediate frequenc: is (*fo*—*fs*). If it is lower, the I.F. is (*fs*—*fo*). At first sight i would seem a matter of indifference which of these alternative was chosen. There are, however, marked practical advantage in making *fo* higher than *fs*.

Suppose the set is to tune from 1,500 to 550 kc/s (200 t 545 metres). Then, if of higher frequency, the oscillator mus

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run from (1,500 + 450) to (550 + 450), i.e., from 1,950 to 1,000 kc/s. If, on the other hand, the oscillator is of lower frequency than the signal, it must run from (1,500 — 450) to (550 — 450), or 1,050 to 100 kc/s. The former range gives 1·95, the latter 10·5 as the ratio between highest and lowest frequency. Since even the signal-circuit range of 2·72 is often quite difficult to achieve, owing to the high minimum capacitances likely to be present in a finished set, the oscillator range from 1,950 to 1,000 kc/s would always be chosen in practice.

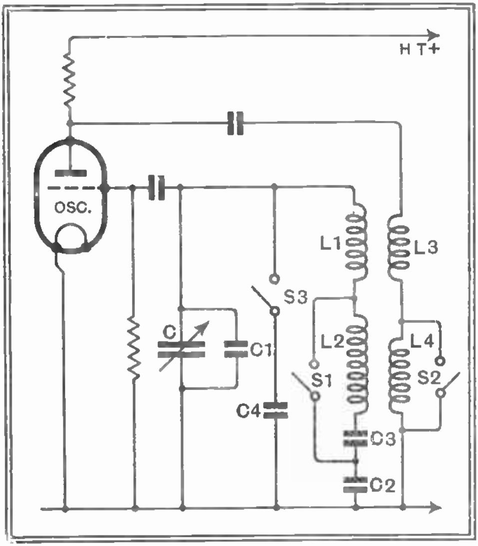

Fig. 167 : Complete dual wave-range oscillator circuit, showing arrangement of series (" padding ") condensers C2 and C3, and of parallel (" trimming ") condensers C1 and C4. By correct choice of values for these, ganging may be made practically perfect

It is evident, since the frequency difference between signal and oscillator must be kept constant, that the oscillator must be tuned in a manner that is in some way different from the tuning of the signal-frequency circuits.

At first sight it might appear impossible to tune the oscillator with a condenser section identical with those tuning the signal-frequency circuits, because the required ratio of maximum to minimum capacitance is different. This difference, however, can readily be adjusted by putting a fixed capacitance either in

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parallel with the oscillator condenser to increase the minimum capacitance, or in series to reduce the maximum. Having got the ratio of maximum to minimum correct in either of these ways, correct choice of inductance for the oscillator coil will ensure that it tunes to the correct frequency at the two ends of the tuning-scale.

In the middle, however, it will be widely out, but in opposite directions in the two cases. It is found that a judicious combination of the two methods, using a small parallel condenser to increase the minimum a little, and a large series condenser to decrease the maximum a little, will produce almost perfect "tracking" over the whole wave-band.

The resulting circuit is that of Fig. 167. Here C is a section of an ordinary gang condenser, and has at every dial reading the same capacitance as its companion sections tuning the signal-frequency circuits. With S_1 and S_2 closed, we have C_1 to increase the minimum capacitance and C_2 to decrease the maximum, their relative values being critical for accurate ganging. Opening S_1 increases the inductance of the tuned circuit to enable the long-wave band (150 to 300 kc/s) to be covered by the set, at the same time decreasing the series condenser to the resultant of C_2 and C_3. At the same time S_2 is opened to throw in the extra reaction winding L_4, and S_3 is closed to add C_4 to the minimum capacitance in the circuit. The arrangement as a whole is shown, for simplicity, with a triode as oscillator, but it is equally suitable for use with a pentagrid or other specialized frequency-changer.

184. Whistles

Owing to the characteristics of the frequency-changer, a superheterodyne is susceptible to certain types of interference from which an ordinary set is free.

The most noticeable effect of these is a whistle, which changes in pitch as the tuning control is rotated, rather like the results of using a "straight" receiver in an oscillating condition but generally less severe. There are many possible causes some of which are quite hard to trace.

The best-known is *second-channel* or *image* interference. In the preceding Section we have seen that it is customary for the oscillator to be adjusted to a frequency higher than that of the incoming signal. But if, while reception is being obtained in this way, another signal comes in on a frequency higher than that of the oscillator by an equal amount, an intermediate frequency is produced by it as well. If the frequency difference is not *exactly* the same, but differs by perhaps 1 kc/s, then two

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I.F. signals are produced, differing by 1 kc/s, and the second detector combines them to give a continuous note of 1 kc/s.

An example will make this clear; and for simplicity the I.F. will be made a round number, 100 kc/s. Suppose the station desired works on a frequency of 950 kc/s. When tuned to it, the oscillator is 100 kc/s higher, 1,050 kc/s, and yields a difference signal of 100 kc/s, to which the I.F. amplifier responds. So far all is well. But if a signal of 1,149 kc/s is also able to reach the frequency-changer it will combine with the oscillation to produce a difference signal of 1,149 — 1,050 = 99 kc/s. The I.F. amplifier is unable to reject this (Sec. 186), so both are amplified together by it and are presented to the detector, which produces a difference frequency, 1 kc/s, heard as a high-pitched whistle. Slightly altering the tuning control alters the pitch of the note, as shown in the table:

TABLE

Set Tuned to: kc/s	Oscillator at: kc/s	IF Carrier due to 950 kc/s signal (wanted)	IF Carrier due to 1,149 kc/s signal (interfering)	Difference (Pitch of Whistle)
		kc/s	kc/s	kc/s
945	1,045	95	104	9
946	1,046	96	103	7
947	1,047	97	102	5
948	1,048	98	101	3
949	1,049	99	100	1
949½	1,049½	99½	99½	0
950	1,050	100	99	1
951	1,051	101	98	3
952	1,052	102	97	5
953	1,053	103	96	7
954	1,054	104	95	9

Since both stations give rise to carriers falling within the band to which the I.F. amplifier must be tuned to receive one of them, this part of the set can give no protection against interference of this sort. That is why it is necessary to have preselector tuning circuits. At first sight it might appear that their task is easy, for the interfering station is twice the intermediate frequency away from the wanted station, and even with an I.F. as low as 100 kc/s that is 200 kc/s. But it must be remembered that (1) the interfering carrier may be thousands of times stronger than the one desired; (2) the product of two carriers is always considerably stronger than one carrier and its

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sidebands ; and (3) 200 kc/s is only 1 per cent. off tune at 20 Mc/s. The demand for short waves and cheap receivers (and therefore a minimum of variable tuned circuits) led to the I.F. in common use being raised from about 110 to about 460 kc/s, giving a separation of nearly 1 Mc/s from the image frequency. The problem has thus becomes less serious on medium waves, but is still very present on short waves, resulting in most stations being received by the cheaper sets at two settings of the tuning knob. Whistles, however, are not as bad as one might imagine, because the number of powerful stations within a given frequency band is less than on medium waves.

Another form of interference, much more serious if present, but fortunately easy to guard against, is that due to a station operating within the I.F. band itself. Clearly, if it is able to penetrate as far as the I.F. tuning circuits it is amplified by them and causes a whistle on *every* station received. Again, a good preselector looks after this ; but the 550 kc/s end of the medium waveband may be dangerously close to a 465 kc/s I.F. If so, a simple rejector circuit tuned to the I.F. and placed in series with the aerial does the trick.

The foregoing interferences are due to unwanted stations. But it is possible for the wanted station to interfere with itself ! When its carrier arrives at the detector it is, of course, always at intermediate frequency. The detector, being a distorter, inevitably gives rise to harmonics of this frequency ; that is to say, currents of twice, thrice, etc., times the frequency. If, therefore, these harmonics are picked up at the aerial end of the receiver, at the same time as it is tuned to a station working on nearly the same frequency, the two combine to produce a whistle. It is easy to locate such a defect by tuning the set to two or three times the I.F. The cure is to by-pass all supersonic frequencies that appear at the detector output, preventing them from straying into other parts of the wiring and thence to the preselector circuits (Sec. 121).

Oscillator harmonics are bound to be present, too ; and are a possible cause of whistles. Suppose the receiver is tuned to 220 kc/s and the I.F. is 465. Then the oscillator frequency is 665, and that of its second harmonic is 1,330 kc/s. If now a powerful local station were to be operating within a few kc/s of 865, its carrier would combine with the harmonic to give a product beating at an audio frequency with the 465 kc/s signal derived from the 200 kc/s carrier. Such interference is likely to be perceptible only when the receiver combines poor preselector selectivity and excessive oscillator harmonics.

The exclusion of most of the varieties of interference peculiar

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to the superhet is fairly easy so long as there are no overwhelmingly strong signals. But if one lives under the shadow of a transmitter it is liable to cause whistles in a number of ways. Besides those already mentioned, an unwanted carrier strong enough to force its way as far as the frequency-changer may usurp the function of the local oscillator and introduce signals of intermediate frequency by combining with other unwanted carriers. Any two stations transmitting on frequencies whose sum or difference is nearly equal to the I.F. may cause interference of this kind ; and with the additive types of frequency-changer are actually known to do so. But a perfect multiplicative frequency-changer does not yield a combination frequency merely by mixing the two, and so is immune. The designer also guards against the danger by a suitable choice of I.F., provided that the frequencies of all pairs of stations that might be "local" are known. And, of course, preselector selectivity comes to the rescue once more.

The foregoing list of possible causes of interference is by no means exhaustive ; but it is only in exceptional situations that whistles are conspicuous in a receiver of good modern design. Some of the very cheap models, however, tend to revive a fault that used to be characteristic of the early superhets—the radiation of energy from the oscillator. One listener may interfere with another tuned to a station separated in frequency by the I.F. If he is on the long wave band, say 200 kc/s, his oscillator on 665 kc/s will, if radiating strongly enough, interfere with a neighbour tuned in to, say, 668 kc/s on the medium band.

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CHAPTER 17

TUNING CIRCUITS IN THE I.F. AMPLIFIER

185. The Task of the I.F. Amplifier

THE I.F. amplifier of a superhet. has to perform exactly the same duties as the R.F. amplifier of a "straight" set. It is really a fixed-tune R.F. amplifier which derives its signal not from the aerial direct, but from the frequency-changer, since this is the point at which the I.F. currents first appear. Just as in the case of the R.F. amplifier, the problems concerned consist mostly of the design of the tuned circuits involved.

The double advantages of fixed tuning and of having to deal with signals of comparatively low frequency completely transform the problem. The fact that our tuning is to be fixed allows us to use more tuned circuits without extra complication, and also to make careful adjustments that could never possibly hold constant over a waveband. The lower frequency, as we saw in Secs. 152 and 154, means that we shall have at our disposal coils of much higher L/r ratio—and hence of much higher selectivity—than we could possibly hope for when dealing with our signals at frequencies round about the 1,000 kc/s mark. We therefore set out, from the beginning, to attain a much higher standard of selectivity than we should dream of attempting in the design of a signal-frequency amplifier.

186. Characteristics of I.F. Coils

The most-used intermediate frequency is 450 kc/s, or values not far removed from this. Previously, 110 kc/s was usual. Experience shows that the values of L/r set out in the table below can be achieved, even with comparatively small coils, with the various types of winding indicated. The figures make rough allowance for the damping effects of valves and other components connected, in the finished set, across the tuned circuits.

Frequency (kc/s)	Type of coil	L/r Microhenries and Ohms
450	Solid wire, air core	25 to 30
450	Litz., air core	40 to 50
450	Litz., iron core	70 to 80
110	Solid wire, air core	50 to 60
110	Litz., air core	Up to 140

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As mentioned in Sec. 152, these figures will rise or fall with the dimensions of the coil, so that they are necessarily only approximate. In addition, for a given frequency they depend on the value of L, growing less as this is increased owing to the fact that the series resistance r equivalent to dielectric loss or other forms of parallel damping is proportional to the *square* of the inductance.

It is not easy, unless one is very familiar indeed with the implications of these figures, to draw any immediate conclusions from them. We will therefore assume that we are called upon to design the I.F. coils for a superheterodyne that

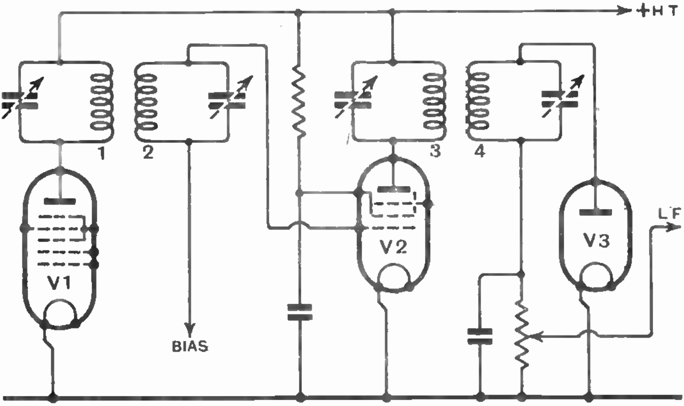

Fig. 168 : Skeleton diagram of single-stage I.F. amplifier. V_1 is the frequency-changer, V_2 the I.F. amplifying valve proper, and V_3 the detector

includes one stage of I.F. amplification. A typical circuit for the relevant part of the receiver is given in Fig. 168, where it will be seen that each of the two I.F. couplings includes two tuned circuits, making four in all. One at each point would suffice to provide the necessary coupling between valves, but as we know (Sec. 150) that the larger the number of tuned circuits the better the compromise between selectivity and high-note reproduction, this minimum is doubled.

To get an idea of the meaning of the L/r values just given, we will draw two overall resonance curves for four *cascaded* tuned circuits, one curve corresponding to circuits of L/r = 140, and one to circuits of L/r = 25, these being the highest and lowest figures in the Table. The curves, drawn from the data curves of Fig. 138, are reproduced in Fig. 169.

The inner one, corresponding to $L/r = 140$, shows the most impressive selectivity—but also shows the most appalling loss of high notes. At 3 kc/s off tune (3,000 cycles audio) the response is little more than one-thousandth of that corresponding to the carrier (and the lowest notes)

The outer curve, corresponding to $L/r = 25$, is more reasonable, being nearly 100 times down at 9 kc/s ; selectivity will be good, while at 5 kc/s (5,000 cycles audio) the response is still one-tenth of that for the bass. Even this curve, if realized in a receiver, would give very " boomy " and deep-toned reproduction of music, badly lacking in the life-giving high notes.

187. The Tuned Filter

The curves of Fig. 169 have been worked out on the assump-

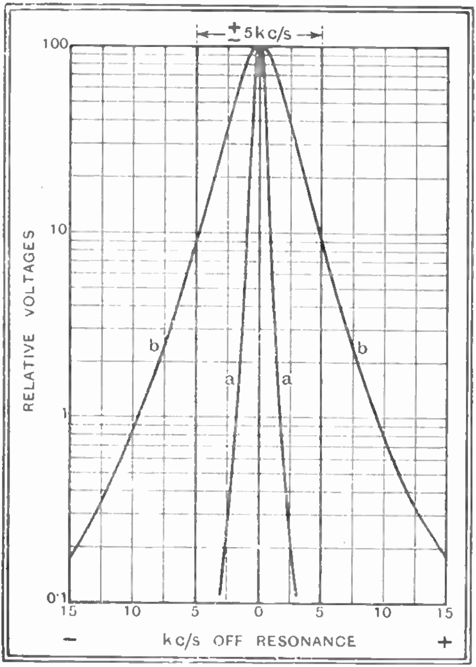

Fig. 169 : Overall resonance curves of four tuned circuits in cascade. *a* $L/r = 140$. *b* $L/r = 25$

tion that the tuned circuits are in cascade, by which is meant that each retains its own individual resonance curve, unmodified

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by the presence of the others. But to pass energy through the intervalve couplings of Fig. 168, some coupling has to be provided by which this energy can pass from circuit 1 to circuit 2, and from 3 to 4. This is done by mutual inductance, the second coil lying in the magnetic field of the first.

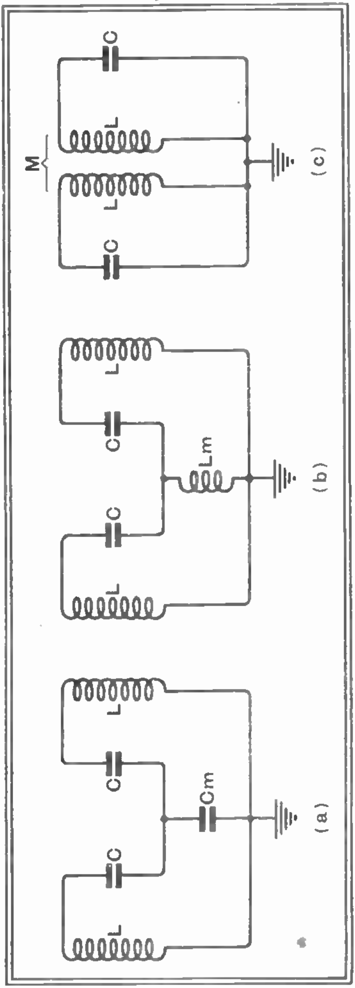

Fig. 170: Three common types of tuned filter. Coupled in *a* by *Cm*, common to both tuned circuits; *b* by *Lm* replacing *Cm*; *c* by mutual inductance M between the coils themselves

In such a case each circuit reacts upon the other, and each modifies the other's resonance curve. There emerges a new joint resonance curve, with characteristics that we have not yet discussed. This effect can equally be had by providing coupling of any other sort between the two tuned circuits. Fig. 170 shows three methods of coupling that are frequently used; in any one of these the complete two-circuit system is known as a *filter*, or *band - pass filter*. More elaborate structures, containing more than two tuned circuits, can be built up, but in ordinary wireless practice the use of tuned filters is generally restricted to a simple two-member combination.

We have seen that the resonance curve of a single tuned

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circuit is determined entirely by the ratio L/r. In a filter we have a second variable in the coupling between the coils, which determines the degree of "spread" round the peak.

If we denote by X the reactance of the coupling element (Cm, Lm, or the mutual inductance M in Fig. 170), then the effect of the coupling in modifying the resonance curve from that of the same two circuits in cascade depends upon the ratio X/r. If, therefore, we know the sharpness of tuning of the individual circuits, as given by L/r, and also the effect of coupling, as given by X/r, we can plot the complete resonance curve of a filter. The formula necessary for this is given at the end of this Chapter.

To investigate the nature of the curve, we will take the very practical case of two tuned circuits (one intervalve coupling in Fig. 168), each of which has $L/r = 40$. If the coupling between them is very weak, so that the reaction of one circuit upon the other is negligible, we get, for the two circuits in cascade, the innermost resonance curve 1 of Fig. 171. This shows a reduction of voltage to 19 per cent. at $\pm$ 4 kc/s from resonance, and to 4·5 per cent. at $\pm$ 9 kc/s. The weak coupling further ensures that even if a large voltage appears across the first coil that across the second will be extremely small.

As the coupling between the two coils is increased by bringing them closer together, the voltage across the secondary increases and the peak of the resonance curve broadens, until at *critical coupling* the curve takes the shape shown at 2 in Fig. 171. The response at $\pm$ 4 kc/s has now risen to 44 per cent., thereby improving the transmission of high notes, but at the cost of a reduction in selectivity, the response at $\pm$ 9 kc/s now being 9·4 per cent. At this coupling the voltage across the secondary is half that which would appear across the primary used as simple tuned-anode coil.

With still closer coupling the voltage, at exact resonance, across the secondary begins to fall a little, while the joint resonance curve takes on the shape shown at 3. The rounded peak of curve 2 has now split up into two separate peaks, with a trough at the actual resonant frequency itself. The response at $\pm$ 4 kc/s is now 98 per cent. of the maximum, while at $\pm$ 5 kc/s it is equal to that at exact resonance. Selectivity has necessarily dropped further, the response at $\pm$ 9 kc/s having risen to 23 per cent.

It would appear that curve 3 offers a suggestion for a very satisfactory design. It provides a rising response up to 5 kc/s from resonance, thereby compensating for probable losses in other portions of the receiver, while at the same time giving

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selectivity which, by using a large enough number of pairs of circuits, might be made sufficiently high. In practice it is found that resonance curves of this type are very hard to achieve, for differences in the L/r values of the two circuits generally lead

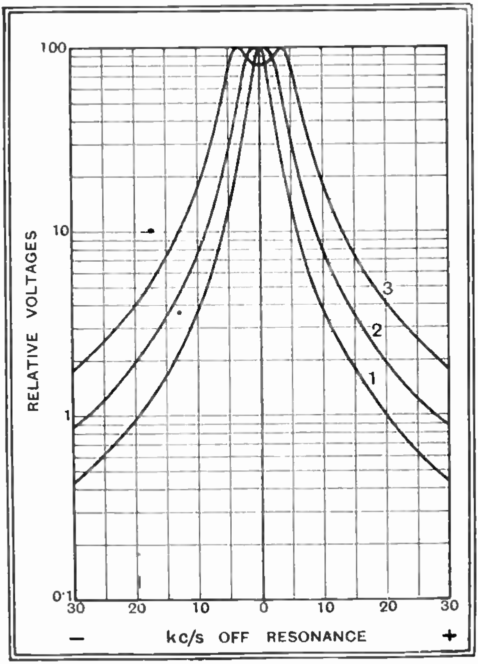

Fig. 171 : Resonance curves of two tuned circuits, each $L/r = 40$. (1) Cascaded : $X/r = 0$. (2) Critically coupled $X/r = 1$. (3) Coupled to give overall band-width $\pm$ 5 kc/s. $X/r = 2{\cdot}04$. (See formula 6)

to a curve in which one peak, being predominant, is brought exactly to resonance, while the other is represented by no more than a slight irregularity on one side or the other of a steeply falling curve. On the whole, it is safest for a designer to content himself with trying to get a peak only a little wider than that of curve 2, which represents the case of critical coupling and maximum gain.

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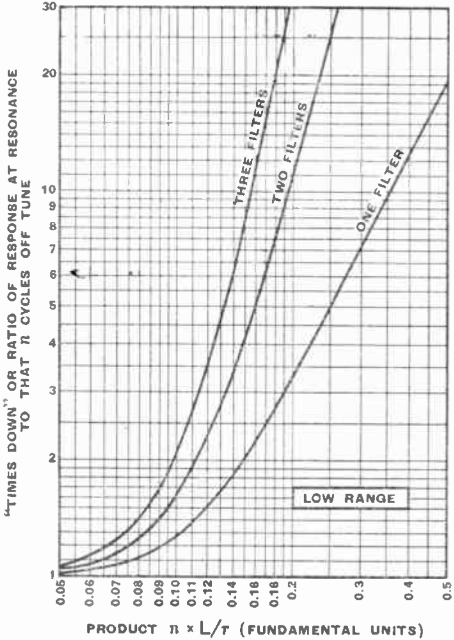

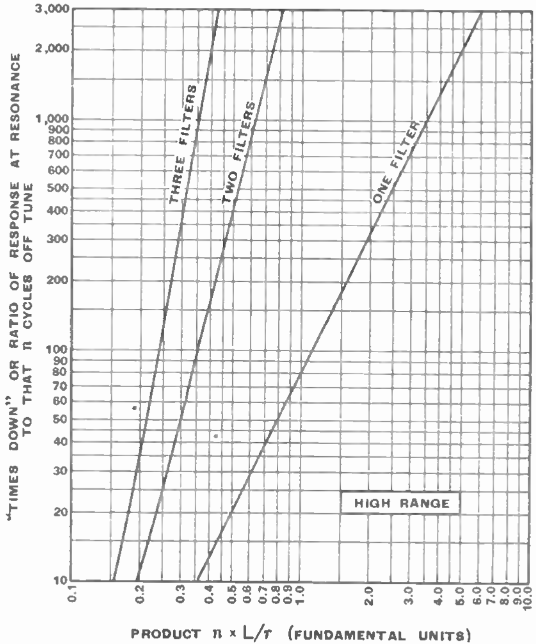

Fig. 172a: Design curves from which overall resonance curves of one, two, or three *critically coupled* filters may be found if

Fig. 172b: Continuing Fig. 172a to higher values

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188. Critical Coupling

Two circuits are critically coupled when the coupling is so close that the peak of the curve is just on the verge of breaking up into two separate peaks. This occurs when the coupling reactance X is made equal to the high-frequency resistance r of either of the circuits (assumed identical), or when the *relative coupling* X/r is made equal to 1. Naturally, the higher r is made the broader will be the peak, since raising r flattens the tuning of each individual circuit and at the same time involves an increase in X to maintain coupling at the critical point. A rapid estimate of the width of the peak can be made by dividing L/r for the circuits concerned into 150,000, which gives the number of cycles off tune at which the response has fallen to half that at resonance.

Thus for two circuits of $L/r = 50$, critically coupled, the curve would fall to half-height at 150,000/50 cycles = 3·0 kc/s off tune. Data for plotting rapidly a complete resonance curve for the particular case of critical coupling are given in Fig. 172. Here "times down" at n cycles off tune is plotted against the product $n \times L/r$, the latter being in fundamental units (cycles, *henries* and ohms). The curve applies to the simple case where the two tuned circuits are identical; in the case of any difference between them an approximation at least could be made by taking a mean value for L/r. This figure fulfils for a filter what the design curves in Fig. 138 do for circuits in cascade.

189. Coupling Closer than Critical

In the third (peaked) curve of Fig. 171 there are two peaks at about $3\frac{1}{2}$ kc/s either side of resonance. A curve of this type is just as easy to plot from the full formula as one for critical coupling, but short cuts are less simple. Owing to the difficulty of obtaining such characteristics in practice, we will do no more than refer the reader to formulæ at the end of this Chapter, which give the number of cycles off tune at which the peaks occur, their height, and the number of cycles off tune at which the final fall of the curve outside the peak brings the response down again to equal that at resonance.

Attention is particularly drawn to the impossibility of combining a flat-topped curve with high selectivity by closely coupling a pair of very low-resistance circuits. Fig. 173 shows the curve of a filter in which each circuit has $L/r = 140$, coupled to give peaks at 5 kc/s off tune. Apart from the fact that the tuning of each circuit reacts upon that of the other to

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such an extent as to make the realization of the curve a matter of extreme difficulty, the great height of the peaks will lead the user of the finished superheterodyne so to tune his oscillator

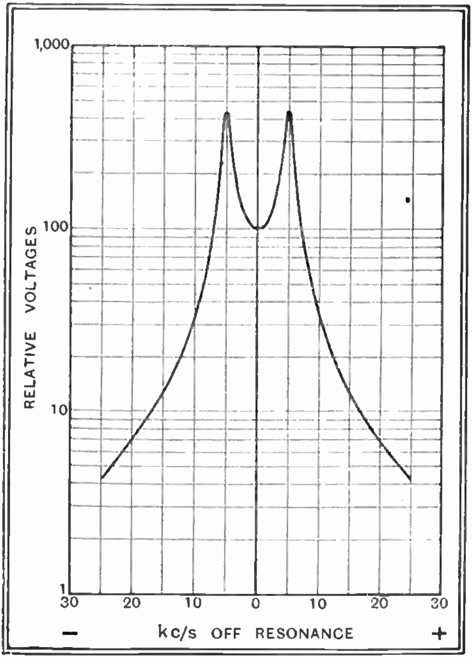

Fig. 173: Showing how "rabbit's ears" develop when an attempt is made to broaden the peak by closely coupling coils of high L/r. $L/r = 140$. $X/r = 79$ (formula 3)

as to put the I.F. carrier, not in the trough, where signals will be quietest, but on one of the peaks, where the output of sound will, in the case shown, be twenty times as great.

190. Designing the Amplifier

We will suppose, therefore, that in supplying coils to the amplifier of Fig. 168 we shall content ourselves with a low L/r ratio and a relative coupling little tighter than critical. This will give us a curve that is not too selective for acceptable quality while keeping away from practical difficulties in tuning.

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Suitable values are $L/r = 40$, $X/r = 1{\cdot}2$ to $1{\cdot}3$, which give us (for one filter) the curve of Fig. 174. This is practically flat to $2\frac{1}{4}$ kc/s off tune, after which it drops away to a little less than half-height at ± 5 kc/s. At ± 9 kc/s it is nearly ten times down. Two such filters in cascade will give a resonance curve typical of that of the I.F. amplifier of the average modern superheterodyne.

The gain to be expected from the I.F. stage is very readily calculated. Since it depends on the dynamic resistance $(2\pi fL)^2/r$ (Sec. 60) of the tuned circuits, it can (theoretically)

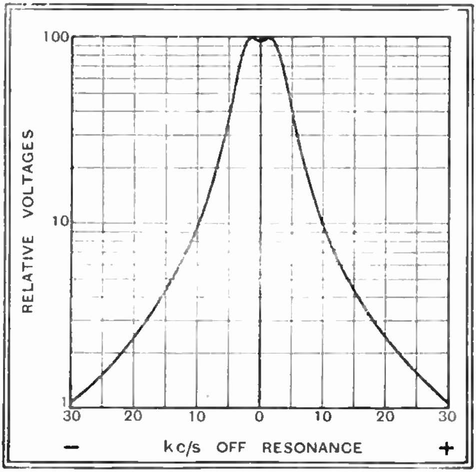

Fig. 174 : Resonance curve of filter suggested as suitable for I.F. amplifier of Fig. 168. $L/r = 40$. $X/r = 1{\cdot}25$

be raised to any desired value by choosing a sufficiently high value for L, of course keeping L/r constant at the chosen value. Let us suppose that the intermediate frequency is 450 kc/s, and that with an I.F. valve of slope $2{\cdot}5$ mA/V we want a gain of 250 times from grid of I.F. valve to grid of detector. Since the coupling is close to the critical value at which the voltage output from the secondary is half that which would appear across the primary is only one coil were used, the gain with a one-coil coupling will have to be almost exactly double this figure, making 500 times. Dividing this by the slope of the valve gives the dynamic resistance required for the anode coil, which is therefore 200,000 ohms. Knowing that $L/r =$

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40×10^{-6}, and $(2\pi f\mathrm{L})^2/r = 200{,}000$ ohms, we readily deduce* that L must be 625 μH, bearing in mind that $f = 450$ kc/s. This inductance we shall have to tune with 200 pF, including strays.

If the pentagrid has a slope of 3 mA/V, the conversion conductance will be at best 1·5 mA/V (Sec. 182), giving a gain of about $(200 \times 1{\cdot}5)/2 = 150$ times, reckoning from R.F. on modulator grid to I.F. on grid of I.F. valve. Since we have designed this stage to amplify 250 times, the overall gain from signal on grid of pentagrid to second detector will be 250×150, or about 37,500 times.

191. Appendix : Filter Formulæ

The *resonance curve* of a filter is given by :

$$\left(\frac{\mathrm{V}o}{\mathrm{V}}\right)^2 = \left(1 - \frac{158n^2p^2}{1+q^2}\right)^2 + \left(\frac{25{\cdot}2np}{1+q^2}\right)^2 \qquad \ldots \quad \ldots \quad \ldots \qquad (1)$$

where Vo = voltage at resonance
V = voltage at n cycles off tune
p = L/r (in henries and ohms)
q = relative coupling X/r, or ratio of coupling reactance to coil resistance.

Critical Coupling occurs when $q = 1$ (See Fig. 171).

This gives maximum voltage on second coil, this voltage being half that which would have appeared on the first coil had it been the only one used. Still closer coupling reduces the mean voltage of the modulated carrier but little.

The resonance curve of a critically-coupled filter can be plotted from :

$$\left(\frac{\mathrm{V}o}{\mathrm{V}}\right)^2 = 1 + 6245n^4p^4 \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (2)$$

The data curves of Fig. 172 are plotted from this, and provide a convenient short cut.

Peaked Curves. (q greater than 1.)

If peak is n cycles from resonance,

$$q^2 = 1 + 158p^2n^2 \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (3)$$

and height of peak is given by :

$$\frac{\mathrm{V}}{\mathrm{V}o} = \frac{1+q^2}{2q} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (4)$$

(Use by finding q, by formula (3), from known L/r and desired n ; then find V/Vo from formula (4).)

Approximate short cut in a single stage : height of peak n cycles out from resonance is given by :

$$\frac{\mathrm{V}}{\mathrm{V}o} = \frac{158n^2p^2 + 3}{4} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (5)$$

$$*\mathrm{L} = \frac{\mathrm{R}}{(2\pi f)^2\mathrm{L}/r}$$

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Overall Band-width

If it is desired that, at n cycles from resonance, the peak shall have been passed and the voltage shall have fallen again to the level of the trough at resonance, make :

$$q^2 = 1 + 79p^2n^2 \qquad \ldots \qquad (6)$$

The rest of the curve can then be sketched by finding n for peak from (3) and height of peak from (4).

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CHAPTER 18

AUTOMATIC CONTROLS

192. The Principle of A.V.C.

ALTHOUGH a few early superhets ended up with a grid detector and output stage, it is usual to take advantage of the high available pre-detector amplification to provide automatic volume control (A.V.C.) or, more correctly, automatic gain control. The purpose of this is, firstly, to cause all stations, strong or weak, to be reproduced at approximately equal volume, as set by the manual volume control knob ; and, secondly, to counteract the fluctuations in volume due to fading of the received signal. A.V.C. makes use of the D.C. that is a waste product of the detector (Sec. 112) to bias back the earlier amplifying valves, so reducing their gain. For this reduction in gain to be effective it is evident that the peak voltage of the signal reaching the detector must be able to rise, without producing distortion, to a value equal to the bias required to reduce the gain of preceding valves to a low figure. This voltage may amount to 15 volts or more, so the diode is the most suitable detector.

193. Simple A.V.C.

Fig. 175 gives a simple A.V.C. circuit, in which the diode V_2 serves both as second detector and as generator of the A.V.C. voltages. The signal applied from the secondary of the I.F. transformer T across anode and cathode of V_2 is rectified in the usual way with the aid of the condenser C and the leak R, the latter being in the form of a potentiometer from which any desired portion

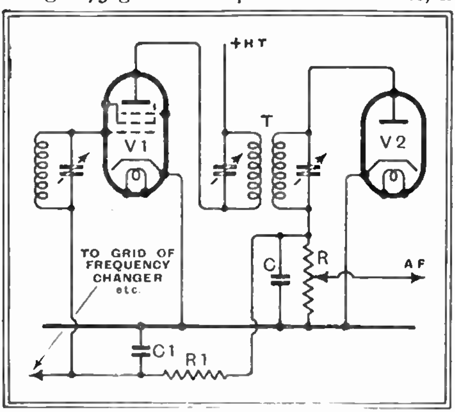

Fig. 175 : Skeleton circuit of simple A.V.C. arrangement. The D.C. voltage produced by signal rectification by V_2 is used for control bias

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of the total A.F. voltage across it can be conveyed to the A.F. amplifying valve. The flow of electrons through R on their way from anode to cathode of V_2 makes the "live" (unearthed) end of R negative to an extent substantially equal to the peak voltage of the applied I.F. signal that is driving the current. This voltage is fed back to the grid of V_1, the filter made up of R_1 and C_1 being interposed to prevent carrier-frequency and audio-frequency voltages from being also fed back along the same path.

If we make the assumption that a bias of 15 volts on V_1 (and other pre-detector valves not shown in the diagram) will be required to reduce their amplification sufficiently to enable them to handle local-station signals, it is evident that when the station is tuned in the peak I.F. voltage applied to V_2 must have this value. Further; any station inducing a lesser voltage in the aerial will give rise to some lower voltage at V_2.

If the degree of A.F. amplification following V_2 is such that 5 volts (peak) of signal is required at that valve to provide full output at the loud-speaker, it will be impossible to obtain full-strength signals without at the same time applying 5 volts of bias to all pre-detector valves. This means that all stations weaker than this are prevented from giving full output, even though the set would have adequate sensitivity to receive them properly if it were not for the A.V.C. system. If there are two controlled valves, and each has its slope reduced to one-tenth of its maximum value by the application of this bias, the sensitivity of the set will be one-hundredth of its maximum value.

Fig. 176 shows, graphically, the type of relationship between input signal and voltage at V_2 that would be given by a circuit like that of Fig. 175. As soon as the initial insensitivity of the detector is overcome, the rectified voltage applied as bias begins to reduce the sensitivity of the set, so that the climb in output with rising input becomes very slow. The dotted line shows how the output voltage would rise if, without the A.V.C, the amplification of the set remained constant irrespective of the signal applied.

It is fairly clear that the full useful sensitivity of the set could be regained if the A.F. amplification following the detector were raised until 1 volt at V_2 provided signal enough to load up the output valve, for at this voltage the A.V.C. has barely begun to reduce the sensitivity. But if this were done, we should find that at the other end of the scale the output would be excessive; for 15 volts bias, and with it 15 volts of signal, would still be produced by tuning in the local station.

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In spite of the A.V.C. system, drastic use of the volume-control would still be required on tuning from a near to a distant station, because the ratio of maximum to minimum power output would be 15^2, or 225 to 1.

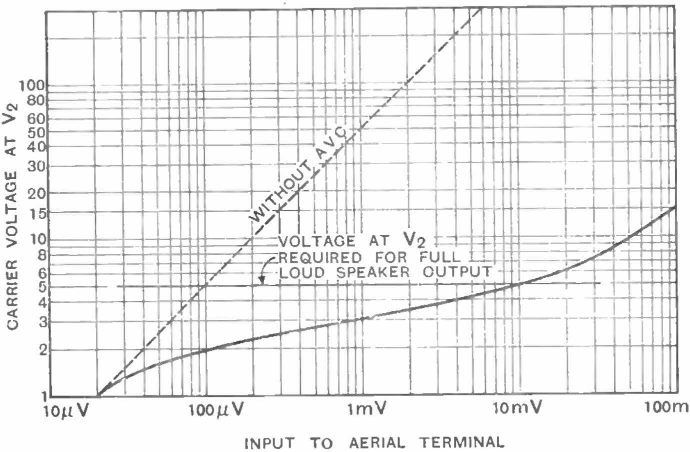

Fig. 176 : A.V.C. curve for system of Fig. 175. Note that if 5 V at V_2 is wanted for full output, the A.V.C. is unnecessarily limiting output on all inputs from 20μV. to 10 mV

194. Delayed A.V.C.

If we can arrange that the signal-voltage is always greater than the A.V.C. voltage, we can reduce this ratio very considerably. Suppose that the signal is allowed to rise to 5 volts before the A.V.C. system begins to operate; then, as 15 volts of bias will still be wanted for the local station, the signal it gives at the second detector must be 20 volts. On the assumption that the post-detector gain is so arranged that 5 volts at the detector fully loads the output valve, we now have a voltage ratio of 4 to 1 from loudest to faintest station within the range of A.V.C., or a power output ratio of 16 to 1, in place of the 225 to 1 of the circuit of Fig. 175.

This very considerable improvement can be realized in practice by the circuit of Fig. 177. So far as the signal circuits are concerned, this is identical with Fig. 175. Detection now takes place at one anode D_1, of a double-diode valve, the leak being returned, as before, to cathode. The signal is also

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applied, through the condenser C_2, to the second diode D_2, whose leak R_2 is returned to the earth line. By means of the battery shown, the cathode of V_2 is made positive with respect to earth, with the result that rectification at D_2 does not begin until the positive peaks of the I.F. signal run this electrode up to a voltage at least equal to that applied to the cathode.

If we make the cathode of V_2 positive by 5 volts and apply a 5-volt (peak) signal we can then adjust the post-detector gain until the rectified output just loads up the output valve. With this signal the A.V.C. diode D_2 is just about to rectify; the

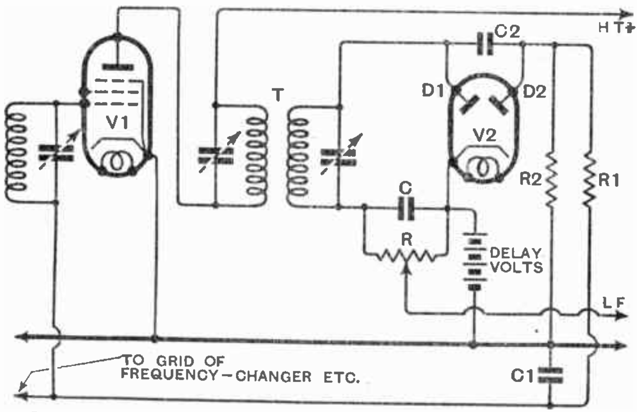

Fig 177 : Modification of Fig. 175 to produce delayed A.V.C. Until the peak voltage of the signal exceeds the positive bias on the cathode of V_2, the A.V.C. system does not begin to operate

ignal is therefore allowed to build up to full output without nterference from the A.V.C. system, which then immediately tarts work and tends to prevent any further rise. For a set so djusted, the A.V.C. curve, carried on to 15 volts bias (= 20 V ignal minus 5 V delay) will be of the type shown in the lowest urve of Fig. 178.

The two other curves represent the response of sets having elays of 10 V and 15 V respectively, and it will be clear that s the delay increases so does the perfection of the A.V.C. ystem. With 15 V delay the signal rises from 15 V to 30 V— 2 to 1 ratio only—for the required increase in A.V.C. bias ·om zero to 15 volts. Higher delay evidently implies that we ıall have to cut down the post-detector gain, so that the verall sensitivity of the set drops in proportion to the delay.

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The simple A.V.C. system of Fig. 175 is practically never used, owing to the disadvantages described, but delayed* A.V.C. produced as in Fig. 177 is used in the majority of modern sets. In place of using the battery shown, the cathode of the double diode is made positive by connecting it to some point of suitable potential elsewhere in the circuit—usually to the cathode of the output valve.

Also D_2 is usually connected through C_2 to the anode of the preceding valve. The object of this is to avoid what is called sideband shriek, caused by the great increase in amplification

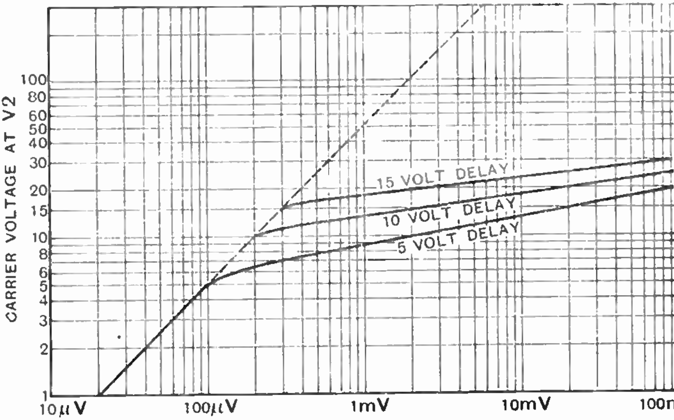

Fig. 178 : A.V.C. curves for circuit of Fig. 177. Note that the larger the delay the flatter the curve, as explained in the text

when A.V.C. voltage is derived from a highly selective circuit and the receiver is slightly mis-tuned. Taking the A.V.C. from the primary of the I.F. transformer, where the selectivity is less, ensures that the gain is kept low until the audible signal, derived from the secondary, has been tuned out.

Owing to the desirability of a large delay it is quite common

* This is another radio term that might have been more wisely chosen. It must be understood that the " delay " is in voltage, not in time. There is actually a time delay in *all* A.V.C. systems, due to the components C_1 and R_1 in Fig. 175. In view of the possible rapid fluctuations of signal strength, on short waves especially, due to fading, this delay ought to be reduced to a minimum consistent with adequate elimination of audio frequencies. About a tenth of a second is suitable.

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to allow the signal-rectifier to supply the output valve direct, without A.F. amplification. If a high-slope indirectly-heated pentode is used, requiring about $4\frac{1}{2}$ V peak signal, and a delay-voltage of 15 V is provided, the output valve will be fully loaded on a carrier 30 per cent. modulated. Alternatively, the delay may be decreased a little, and enough amplification provided after the detector to allow a low-slope pentode or even a triode to be used as output valve. In this case it is usual to employ a double-diode-triode, which as its name implies, combines a double-diode for detection and A.V.C. with a triode for subsequent amplification, all being built into the same bulb.

195. A.V.C. Distortion

Either simple or delayed A.V.C. is liable to lead to distortion if the circuit, both of the A.V.C. system itself and of the I.F. amplifier, is not properly proportioned. It can be shown that if the audio-frequency load of a detector is less than the D.C. load, distortion occurs when the modulation depth, reckoned as a percentage, exceeds a hundred times the ratio of the two loads. In Fig. 175 the D.C. load of the detector is R, while the speech-frequency load is more nearly equal to R and R_1 in parallel (Sec. 115). If R is 0·25 M Ω and R_1 is 1 M Ω, which are quite usual values, the audio-frequency load is 0·2 M Ω only, and distortion will occur if the modulation depth exceeds $\frac{0{\cdot}2}{0{\cdot}25} \times 100$, or 80 per cent.

A second source of distortion is found in the I.F. valve immediately before the detector, which in a set using simple A.V.C. is called upon to deliver a signal of the order of 10 to 15 volts when the local station is tuned in. With delayed A.V.C., the signal is even larger, being equal to the figure mentioned plus the delay voltage. To allow the last I.F. valve to pass on so large a signal it is not unusual to supply it with half only of the available A.V.C. voltage—which, on a strong signal, would bias the valve almost back to the bottom bend—but some risk of distortion still remains.

Both these sources of distortion can be avoided by using amplified A.V.C.

196. Amplified Delayed A.V.C.

When it is desired for any reason to work with a signal of the order of 1 volt at the detector, it is usual to provide amplified A.V.C., in which the rectified voltage is amplified before being

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fed back to earlier valves. This is done with a double-diode-triode in some such manner as shown in Fig. 179. As before, the signal is rectified by the diode D_1, with the leak R returned to cathode. The signal is passed for amplification to the grid of the triode, and as we want to amplify the D.C. as well, for A.V.C. purposes, we do *not* interpose a blocking condenser as in Fig. 107 (Sec. 112). The amplified signal is applied in the usual way to the grid of the output valve, while the amplified D.C. is developed also across a resistance R_3 connecting the cathode to a point some 100 volts negative with respect to the

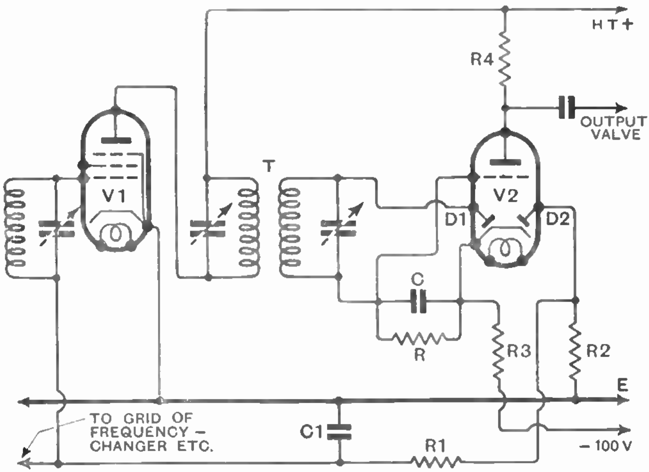

Fig. 179 : Amplified A.V.C., with delay. Initially positive, the cathode of V_2 is driven down to earth potential by the grid bias generated by rectification at D_1. At this point D_2 begins to draw current, and A.V.C. starts work

general earth-line of the set. R_3 and R_4 are so chosen that with no bias on V_2 other than that generated by grid-current through R the cathode is some 30 volts positive with respect to earth. The voltage drop across R_3 would therefore have to be about 130, requiring the resistance of R_3 to be fairly high. When a signal is rectified by D_1 the resulting steady negative voltage, as well as the A.F. signal, reaches the grid of the triode. This negative bias reduces the anode current of the valve, thereby reducing the voltage-drop across R_3 and tending to make the cathode negative. If the amplification is thirty times, a one-volt signal on D_1 will drive the cathode from + 30 V to

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earth potential. A further half-volt will drive it down to -15 V. The diode D_2, connected to earth through the high resistance R_2, takes no current so long as the cathode is positive, but as soon as the cathode reaches earth-potential current begins, the resistance between cathode and D_2 drops to a negligible value, and D_2 follows the cathode downward in potential. Signals up to 1 V on D_1 therefore generate no A.V.C. bias, and the set remains at full sensitivity, but by the time the signal reaches $1\frac{1}{2}$ V the full bias of 15 V is produced on D_2 and fed back to earlier grids in the usual way.

Thus, by this system, an even more level A.V.C. curve than that corresponding to a 15-volt delay in Fig. 178 can be produced from a 1-volt signal. Furthermore, the smallness of the signal ensures that the last I.F. valve shall never at any time be overloaded, while with the arrangements shown the audio-frequency and D.C. loads on the detector are identical, since the resistance R fulfils both functions. A.V.C. distortion is thus completely avoided.

By suitably increasing the positive potential of the cathode of the D.D.T., and increasing the signal voltage to correspond, almost perfect A.V.C. can be produced. It is possible to have a 10-volt delay (cathode at $+150$ V, amplification 15) followed by a rise in A.V.C. volts to the required 15 on increase of the signal from 10 to 11 volts. The A.V.C. curve for a system of this sort is a very close approach to the ideal, in which the dotted line of Figs. 176 or 178 would be followed up to the point at which full loud-speaker strength was reached, after which there would be no further rise in output, no matter how greatly the input were increased.

197. Push-Button Tuning

The demand for broadcast receivers that are simply and quickly operated led to a variety of methods by which a number of stations can be selected by pressing buttons instead of by turning a variable condenser. Of these methods, three are common:

(1) *Mechanical Systems.* A receiver in which the variable condenser is rotated to the tuning position for one of a number of selected stations by a mechanical link system from the push-button was marketed as long ago as 1928. The principal problems involved are ensuring that the exact tuning point is reached in spite of the magnification of mechanical movement, and enabling the selected station to be changed to suit local conditions. This method is confined generally to the cheaper models.

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(2) *Pre-set Systems*, in which a separate set of pre-selector and oscillator tuned circuits is provided for each desired station, and brought into circuit by push-button switches. There are two varieties, according to whether the tuning circuits are pre-set by a screw-type of variable condenser (such as those used for trimmers and padders (Sec. 183), with a fixed coil, or by screw-in iron core varying the inductance, with a fixed condenser. The latter is often called a permeability tuner.

It is difficult to control more than one pre-selector circuit in this way, so R.F. amplification is virtually ruled out; and the tuning circuits have to be very carefully designed if the frequency is to be free from drift. In particular, they must not be appreciably affected by changes in temperature.

(3) *Motor-driven Systems*. These are the most highly developed. The variable condenser is turned by a small electric motor, which is set into operation by the push-button and stops when it reaches the required position. As with the other systems, it is difficult to ensure that the tuning is sufficiently exact, so is usually supplemented by automatic frequency correction.

198. Automatic Frequency Correction

To obtain good quality of reproduction from a superhet, particularly if it is highly selective and the A.F. amplifier is designed to reinforce the high notes to compensate for their loss (Sec. 124), it is essential for the tuning to be accurate. It is difficult to ensure the necessary accuracy with any type of push-button tuning, and even with manual tuning the non-technical user cannot be relied upon to tune precisely enough. And even if properly tuned in when the receiver is first switched on, the frequency of the oscillator is likely to drift slightly as it warms up. A single kilocycle change is enough to upset the tone, and when tuned to a 10,000 kc/s station (30 metres) that is a change of only one-hundredth of 1 per cent.

The worst effects of the resulting sideband shriek are avoided, as already explained (Sec. 194), by suitable A.V.C. design; but that is intended merely to prevent unpleasant noises when tuning from one station to another. To ensure satisfactory results when listening to a station, especially in motor-tuned receivers, automatic frequency correction is employed in the more elaborate sets.

This is operated from the output of the I.F. amplifier, and is so designed that whenever the I.F. carrier passing through the set departs from the frequency to which the amplifier is

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tuned, the control makes the slight readjustment to the oscillator frequency that is required to bring the I.F. carrier back to its correct frequency.

The basis of the control is a *discriminator* and a *frequency-control valve.* In one system the discriminator consists of two sharply-tuned circuits arranged to peak one on either side of the nominal I.F. and at a separation of about 4 kc. from it. These receive the signal from the last I.F. amplifier, and are connected to two separate rectifiers in such a way that the rectified currents are in opposition, as shown in Fig. 180.

If the I.F. carrier F_c has exactly the correct frequency, and the two tuned circuits peak at frequencies equally spaced on

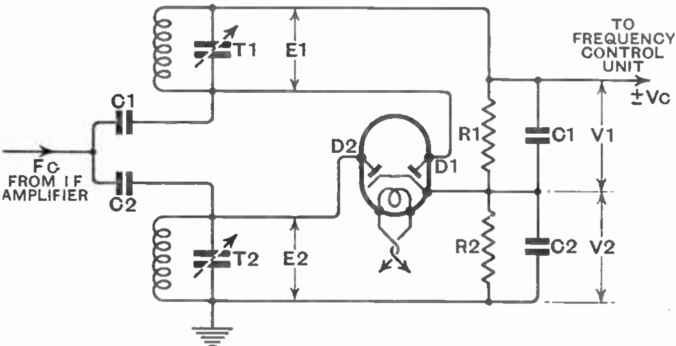

Fig. 180: The circuits T_1 and T_2, tuned one on either side of the correct intermediate frequency, pass to the double diode voltages which, when rectified, can be made to control oscillator-frequency

either side of it, E_1 and E_2 will be equal, and the rectified voltages V_1 and V_2 will also be equal. Being opposite in direction, the control-voltage developed will be zero. If F_c now approaches the frequency of the upper tuned circuit, E_1 will become greater than E_2, and V_1 will therefore exceed V_2, producing a resultant control voltage V_c that is negative in sign. Similarly, V_c will be positive if F_c drifts in the other direction, for E_2 will now exceed E_1, so that V_2 will be greater than V_1. The discriminator thus provides us with a voltage that depends for its sign on the direction in which the I.F. carrier departs from its correct frequency, becomes greater when mistuning is increased, and falls to zero when tuning is accurate.

A somewhat different arrangement, known as the phase discriminator, is now more commonly used.

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This voltage can be used to control oscillator frequency in any one of several ways, of which perhaps the simplest consists of connecting grid and cathode of the frequency control valve across the oscillator tuned circuit. If this valve has its gain controlled by the output from the discriminator, its input capacitance can be made to change (Sec. 120) sufficiently to provide the necessary small alteration in oscillator tuning.

In setting up a circuit of this kind care has to be taken that the control is in the right direction, so as to increase oscillator frequency when it is too low and *vice versa*. Incorrect connection, remedied by interchanging earthed and live sides of the output of Fig. 180, results in the slightest mistuning being automatically increased so that as soon as a station is found the control tunes it out again.

If a set fitted with A.F.C. is tuned to a fading station close in frequency to a powerful station, it may sometimes happen that the carrier wave of the former is not strong enough to prevent the latter affecting the A.F.C. and forcibly substituting itself for the desired programme. Imperfect alignment of the A.F.C. increases this tendency.

199. Automatic Selectivity Control

It is possible, by means of rather complex circuits, to devise schemes whereby the resonance curve of an I.F. amplifier can be broadened or narrowed by automatic means. In general a strong signal, received from a nearby station, is heard without much interference from other transmitters, while when receiving a weak station other transmitters on neighbouring wavelengths are liable to interfere. The A.V.C. system therefore may be used to control selectivity, broadening the tuning curve, initially of high selectivity, on receiving a strong signal. By this means a rough-and-ready automatic adjustment of the selectivity-quality compromise to suit changing conditions may be made with the limitation that for all strong stations, whether interference is present or not, high quality and low selectivity is provided, while for all weak stations, even if no interference is present, the opposite adjustment is made.

A still more complex, but at the same time more satisfactory solution to the problem may be made on the lines of Fig. 180, with the difference that the auxiliary circuits are tuned to the channels on either side of the required station—i.e., to frequencies 9 kc. higher and lower than the intermediate frequency. By reversing the connections of one detector so that the rectified voltages add, it becomes possible to narrow the selectivity curve of the receiver, initially made broad, whenever a signal is

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present on either of the channels adjacent to that being received. A disadvantage of this scheme is that interference is in any case weaker than the desired signal, so that it becomes necessary to provide two extra I.F. amplifiers, tuned 9 kc. on either side of the one dealing with the signal, to amplify the interference sufficiently to enable it to provide an adequate control-voltage from the interference-detecting system.

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CHAPTER 19

POWER SUPPLIES

200. The Power Required

MOST of the power that has to be fed to transmitters and receivers to make them work is needed for the valves, and this can be divided roughly into three classes: cathode-heating, anode (including screen-grid) supplies, and grid bias.

So far as cathode heating is concerned, it is purely a matter of convenience that it is done electrically. In principle there is no reason why it could not be done by a bunsen burner or by focusing the sun on it with a lens. But there are overwhelming practical advantages in electric heating.

Cathode heating power has to be supplied all the time the valves are required to work, whether or not they happen to be drawing any other power at the moment; and it is a substantial part of the total fed to a transmitter or receiver, so many attempts have been made to produce a "cold" valve, hitherto with very limited success.

Power supplied to the anode is more directly useful, in that a part of it is converted into the R.F. or A.F. output of the valve. This is generally not true of power taken by auxiliary electrodes, so valves are designed to reduce that to a minimum.

Not all valves need any grid bias; and, of those that do, not all have to be provided with it from a special supply. And only in the case of large transmitting valves is any substantial grid power consumed, the reason being that with this exception care is taken to keep grid current as nearly as possible down to nil (Sec. 73).

201. Batteries

A battery consists of a number of *cells*, and although a single cell is commonly called a battery it is no more correct to do so than to call a single gun a battery.

A cell consists of two different sorts of conducting plates or *electrodes* separated by an "exciting" fluid or *electrolyte*. The E.M.F. depends solely on the materials used, not on their size, and cannot greatly exceed 2 volts. The only way of obtaining higher voltage is to connect a number of cells in series to form a battery.

Other things being equal, a small cell has a higher internal resistance than a large one, so the amount of current that can be drawn from it without reducing its terminal voltage seriously

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is small. To obtain a large current a number of cells of equal voltage must be connected in parallel, or (preferably) larger cells used.

Many types of cells have been devised, but only two are commonly used for radio. They represent the two main classes of batteries—primary and secondary.

Primary cells cease to provide current when the chemical constituents are exhausted; whereas secondary cells, generally called *accumulators*, can be "recharged" by passing a current back through them by means of a source of greater E.M.F. The term "accumulator" is not very apt; and "recharging" is misleading, for it is simply a method of restoring the original chemical condition—quite a different process from recharging in the electrical sense (Sec. 23).

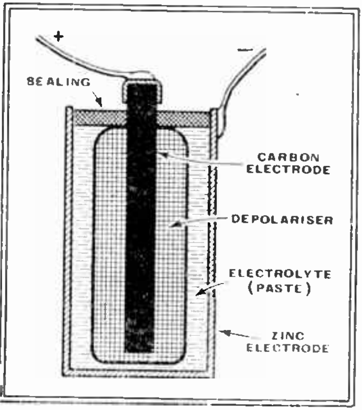

Fig. 181: Cross section of a dry cell showing the essential constituents

The most commonly used primary batteries are the so-called dry batteries—again, a badly chosen term, because any really dry battery would not work. The name refers to cells in which the electrolyte is in the form of a paste instead of a free liquid. The essential ingredient is ammonium chloride or "sal-ammoniac", and the plates are zinc (—) and carbon (+). The E.M.F. is about 1·4 volts per cell. When current is drawn, each coulomb causes a certain quantity of the zinc and electrolyte to be chemically changed. Towards the end of its life, the E.M.F. falls and internal resistance rises. Ideally, the cell would last indefinitely when not in use, but in practice a certain amount of "local action" goes on inside, causing it to deteriorate; and its "shelf life" is generally only a few months. Before buying dry batteries it is wise to have their E.M.F. per cell tested *when supplying full normal current*. If it is 1·3 or less it has been kept in stock too long.

If only the ingredients named above were included in a dry cell, its voltage would drop rapidly when supplying current, and recover after a period of rest, due to the formation and dispersal of a layer of hydrogen bubbles on the surface of the

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carbon. To reduce this layer a "depolariser" is used, consisting largely of manganese dioxide around the carbon electrode. Fig. 181 gives an idea of the usual construction of dry cells. The small ones used in torches and "H.T." batteries can supply several hundred milliamps for short periods, say a minute, with plenty of time to rest; but when used for hours on end, as in a radio set, a steady performance cannot be expected if the drain is allowed to exceed about 10 mA for the smallest sizes.

The most commonly used secondary cell is the lead-acid type, in which the plates consist of lead frames or grids filled with different compounds of lead, and the electrolyte is dilute sulphuric acid. Accumulators have an E.M.F. of 2 volts per cell and possess the great advantages of very low internal resistance and of being rechargeable. Against this there are certain drawbacks. The acid is very corrosive, and liable to cause much damage if it leaks out or "creeps" up to the terminals. If the cell is allowed to stand for many weeks, even if unused, it becomes discharged, and if left in that condition it "sulphates", that is to say, the plates become coated with lead sulphate which cannot readily be removed and which permanently reduces the number of ampere-hours the cell can yield on one charge—the so-called "capacity", which, however, means something quite different from the capacity of a condenser.

If the terminals are short-circuited, the resistance of the accumulator is so low that a very heavy current—possibly hundreds of amps.—flows and is liable to cause permanent damage. Some accumulators, such as those used in motor cars, are designed to supply heavy currents for short periods. Others, called "block" or "mass" batteries, are intended to give small currents for long periods, and are suitable for certain wireless purposes.

Accumulators are capable of very satisfactory service if they are properly looked after according to the makers' instructions for the particular type. The more important points are to keep the acid level to the mark by adding distilled water; to avoid long periods of disuse, maintaining a regular cycle of charge and discharge; to recharge whenever the voltage on load falls below 1·8 volts per cell; never to exceed the specified charge and discharge rates; and to keep acid off the terminals by greasing them with vaseline and making sure the vents are properly screwed down.

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202. Cathode Heating

As stated in Sec. 63, the cathode of a valve may be either directly or indirectly heated. The power for either type can be obtained from a battery, but in practice directly-heated valves are used because they require only about one-tenth of the power needed by the indirectly-heated types, and therefore a given battery will run them for much longer, or alternatively a much smaller battery can be used. These points are important, because battery power is far more expensive than that drawn from the mains, and high-power batteries are very heavy. Seeing then that the chief use of batteries is to enable the set to be carried about, it is desirable for the power required to be as small as possible.

Most battery valves are intended to have their filaments connected in parallel to a single 2-volt accumulator cell. This gives a more uniform E.M.F. than a dry cell; but for greater portability the latter has advantages, and dry-cell valves rated at 1·4 volts and 0·05 or 0·1 amp. are produced for this purpose. As the emission (and therefore the output the valve can give) depends on the filament heating power, these dry-cell valves are naturally more restricted than the 2-volt types, most of which are designed to take currents of 0·1 to 0·3 amp. An average "wireless" accumulator (a 20 amp.-hour cell) will supply one ampere for 20 hours on one charge; this is equivalent to running a 3- or 4-valve receiver for 30–40 hours, which may be a fortnight of ordinary use.

The vast majority of public electricity supplies are A.C. of about 230 volts at a frequency of 50 c/s, which can be stepped down to any convenient voltage for valve heating by means of a transformer (Sec. 48). If an attempt were made to run directly-heated valves in this way it would be unsuccessful, because the slight variation in temperature between peak and zero of the A.C. cycle, and also the variations of potential between grid and filament from the same cause, would produce a 50 c/s variation in anode current, which would be amplified by the following stages and cause a loud low-pitched hum. An occasional exception is a carefully arranged output stage, which of course is not followed by any amplification.

By heating the cathode indirectly (Sec. 63) its temperature changes too slowly to be affected by 50 c/s alternations, and as it carries none of the A.C. its potential is the same all over. Incidentally this improves the valve characteristics, and so does its larger area compared with a filament. Further, the greater rigidity of an indirectly-heated cathode allows the grid to be

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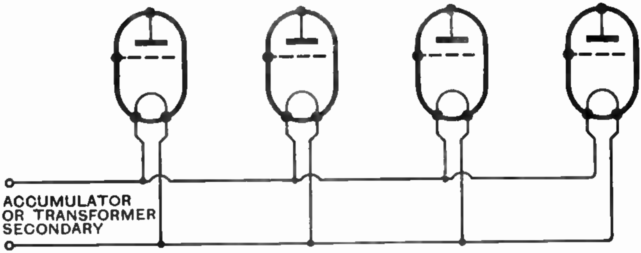
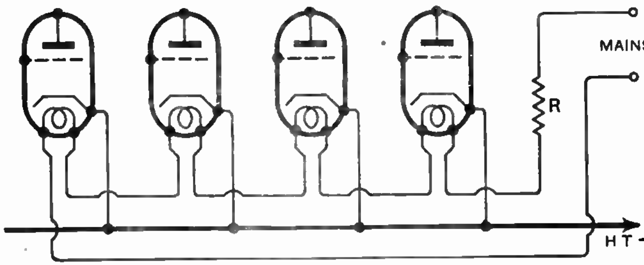
mounted closer to it; resulting in still greater mutual conductance. One may therefore expect a mains-driven set to be considerably more sensitive than a battery set with the same number of valves.

The most usual ratings for A.C. receiver valves are 4 volts 1 amp. (sometimes 2 amps. for output stages) or 6·3 volts 0·2 or 0·3 amp. This curious voltage is to enable these valves



(a)



(b)

Fig. 182 : Diagram *a* shows method of heating filaments of battery valves or heaters of A.C. valves in a 4-valve set. All valves require the same voltage. In diagram *b* which shows D.C. or universal valves with heaters in series, all take the same current. Note that in spite of different potentials of heaters all cathodes can be joined to H.T. –

to be used alternatively in cars, whose batteries average 6·3 volts when on charge. They are connected in parallel, like battery valves, as in Fig. 182*a*, but there are valves designed to have their heaters in series (Fig. 182 *b*), so the current is the same for all, but the voltage may be from 13 to 40 according to type. These are intended for D.C. mains (Sec. 210).

Practically all transmitter valves are directly heated, because the required emission is very great and indirect heating would be uneconomical. As the filaments are too large for their emperature to vary appreciably in one-hundredth of a second, and little or no amplification is involved, indirect heating is also unnecessary. The largest transmitter valve filaments take currents of 100 amps. or even more, at 10–30 volts.

203. Anode Current from the Mains

Supplied by an ordinary H.T. battery, one may reckon a unit (kilowatt-hour) of electricity to cost some thirty shillings at least. From the mains, even allowing for all the losses in conversion, one shilling would be a generous estimate. So one can afford in a mains-driven set to use plenty of anode current ; which means, in turn, a more ample output and less need to run the output stage on the verge of distortion.

The power is there ; the problem lies in making use of it. Obviously the A.C. cannot be used as it is, because during one half of each cycle the anodes would be negative and the valves would be inoperative. Before we can use it, we have to convert the current from alternating to direct.

204. Essential Parts of a Power Unit

We have already seen, in Chapter 10, how to convert A.C. into D.C. The first requirement is a *rectifier*, which is any device for allowing current to pass in one direction only. If the A.C. is of sine waveform it can be represented by Fig. 183 *a*. The result of connecting a rectifier in series is to abolish half of every cycle, as shown at *b*.

The *average* of this is only about one-third of the peak alternating current (Sec. 103), and in any case would be useless for running radio sets because it is not continuous. By using a *reservoir condenser* (Sec. 106) this defect can be removed and at the same time the average is brought almost up to the full peak voltage. The conditions are not quite the same as in a detector, because there the load resistance is usually quite high, whereas a typical receiver taking about 62 mA at 250 volts is a load resistance of 4,000 Ω. Unless the reservoir condenser is enormous, therefore, it has time for the discharge to cause an appreciable loss of voltage (Sec. 19) between one half-cycle and the next, causing the current supply to vary, as in Fig. 183*c*.

This is described as D.C. with a ripple, or unsmoothed D.C. ; and would cause a hum if used to feed a receiver. The third requirement is a smoothing circuit, or *filter*, designed to impede the A.C. ripple while permitting the D.C. to pass freely. If the filter is effective, the result is as in Fig. 183*d*, practically indistinguishable from current supplied by a battery.

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These three are the essential parts of most *power units* or "power packs". As we shall see, the reservoir condenser is not invariably used. But a transformer is generally necessary

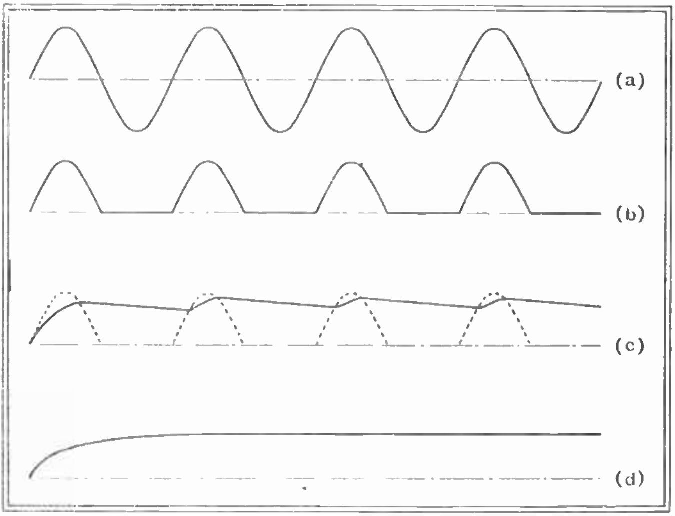

Fig. 183 : Stages in the conversion of A.C. to D.C. The original A.C. is shown as a. When rectified, every alternate half-cycle is suppressed (*b*). By using the rectified voltage (dotted in c) to charge a reservoir condenser, the resulting voltage across the condenser (full line) rises during the charging periods and falls in the intervening periods. By passing the current from the condenser through a suitable filter it is smoothed into practically pure D.C. (*d*)

to provide the required voltage, and has the incidental advantage of insulating the D.C. output from the mains.

205. Types of Rectifier : Soft Valves

The simple diode valve (Sec. 64) is an obvious rectifier, and is perhaps the most commonly used of any. Power rectifiers differ from diode detectors in having larger cathodes to supply the much greater emission needed, and larger anodes to dissipate the greater heat. A typical rectifier for a receiver has a 4-volt 2-amp. heater and is capable of supplying 120 mA at 350 volts.

Transmitter rectifiers are made for supplying up to several amps. at tens of thousands of volts. To stand the high reverse voltage when the diode is not passing current, the anode must be spaced some distance from the cathode. Then during the

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other half-cycle, when a very heavy current is flowing, the space-charge (Sec. 63) is bound to be large, requiring a large voltage between anode and cathode during the conductive half-cycle to neutralize it (Sec. 64). This voltage is deducted from the output and thereby lost, merely causing an undesirable amount of heat to be developed at the anode.

A very interesting way has been found of overcoming this disadvantage. You will remember that in discussing screen-grid tetrodes we came across the phenomenon of secondary emission, caused by the original electrons from the cathode adding to their numbers by knocking out others from the anode, or from anything else they might strike. Much the same thing

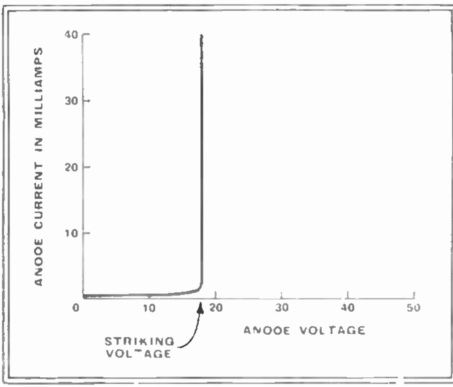

Fig. 184: Characteristic curve of a "soft" diode. When it is passing current the voltage across it is nearly constant, and R_a is therefore nil

happens in the space between if a small quantity of gas or vapour is present there. Provided that the electrons are travelling at a certain minimum speed, they are capable of knocking electrons out of the vapour molecules, thereby supplementing the emission of the cathode. That is not all; a still more valuable effect is that the molecules with electrons missing are thereby positive charges (Sec. 8) or *ions*. They therefore neutralize an equal number of electrons in the space, thus reducing the negative space charge. It might be thought that as each positive ion can neutralize only one electron, the original space charge, due to electrons emitted from the cathode, is untouched. This is not so, because although the ions are naturally attracted to the negative cathode, they are far heavier than electrons and move comparatively slowly. Each one remains in the space for a length of time sufficient for several electrons to cross it, and so neutralizes the space charge not only of its own electrons but also those from the cathode.

The anode-voltage/anode-current characteristic of this type of valve, known variously as a soft valve, or gas-filled valve, is therefore quite different from that of a high-vacuum or "hard" valve (compare Fig. 184 with Fig. 62). As the anode voltage is applied gradually, a very small anode current flows, until the voltage is sufficient to move the electrons fast enough to *ionize* the gas. When once that starts, the space charge is reduced, causing the anode voltage to be more potent, increasing the ionization ; and so on, until the internal resistance of the valve is nil and the current is limited only by the external circuit resistance. The point at which the current begins to increase without limit is called the *striking voltage*. This is the voltage drop across the valve, almost regardless of the current passed, and depends mainly on the type of gas or vapour. With mercury vapour it is only about 12–15 volts, even though several amps. of current may be flowing. The power lost in the rectifier is therefore very small compared with that available at the output, and a rectifier no bigger than a receiving valve can handle an output of a kilowatt.

Soft valves are slightly more tricky to use than hard—for example, they are likely to be damaged unless the cathode has had time to warm up before the anode voltage is applied—so are used mainly for heavier current purposes than ordinary receivers.

Quite a different sort of rectifier is made of alternate discs of copper oxide and copper, and called a metal rectifier. The discs are clamped together and require no " bottle " or heater. They are suitable for heavy currents, and although they are sometimes employed to rectify thousands of volts they are more outstandingly useful for low voltages such as are needed for charging batteries. Provided that they are not overheated by passing more than their rated current, they are extremely reliable and robust.

Very similar is the selenium rectifier, which is capable of standing a higher temperature.

Summarising : the hard diode is most suitable for moderate or small currents at moderate or large voltages ; the soft diode is better for heavy currents at all except very low voltages ; and the metal or selenium rectifier is best for any current at low voltages.

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206. Rectifier Circuits

There are various ways in which rectifiers can be connected in power unit circuits, and the simplest is indicated in Fig. 185*a*. A transformer T is shown for stepping the mains voltage up or down as required, and a single rectifier—the symbol denotes any type—is connected in series with it and a reservoir condenser C_1. A simple filter circuit consisting of a choke L and condenser C_2 completes the apparatus.

This is called a *half-wave* rectifier circuit, because only one-half of each A.C. cycle is utilised, the other being suppressed

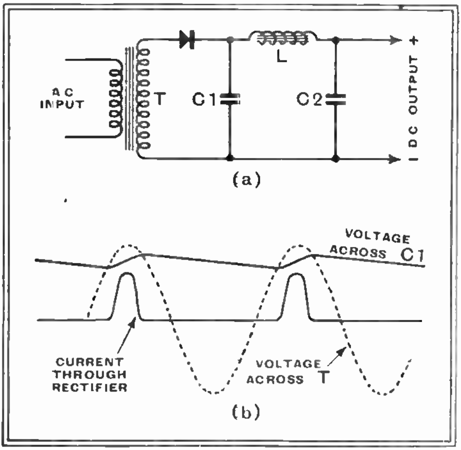

Fig. 185 : Simple half-wave rectifier circuit (*a*). Current can flow through the rectifier only when the voltage from the transformer T exceeds that across the reservoir condenser C1 (*b*). As this is during only a small part of each cycle, the rectifier current is several times greater than the output current, which flows continuously

(Fig. 183 *b*). If, as is desirable, C_1 is large enough for the voltage not to drop much between half-cycles, the proportion of each cycle during which the alternating voltage exceeds it is very small; and, as the recharging of C_1 has to take place during these brief moments, the rectifier is obliged to pass a current many times greater than that drawn off steadily at the output (Fig. 185 *b*). If a large output current is needed, then the reservoir condenser must be very large, the rectifier must be able to pass very heavy peak currents, and there are difficulties in designing the transformer to work under these conditions. So this circuit is confined mainly to low-current high-voltage applications.

Fig. 186 compares this half-wave circuit with the more elaborate arrangements; and for comparison the transformer is assumed to supply the same peak voltage in all of them.

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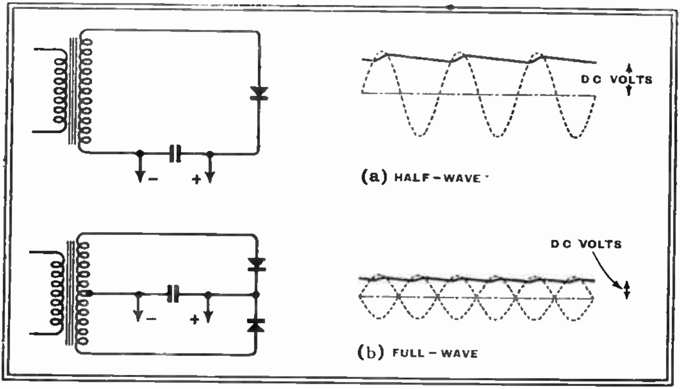

(*a*) *Half-wave.* The no-load output voltage (i.e., the voltage when no current is being drawn) is very nearly equal to the peak input voltage; but, for the reasons given above, the voltage tends to drop considerably on load.

(*b*) *Full-wave.* By centre-tapping the transformer secondary, two type *a* circuits can be arranged, in series with the A.C. source, to feed the load in parallel. For the same voltage from the whole secondary, the output voltage is halved; but for a given rectifier rating the current is doubled, and at the same time is steadier, because each rectifier takes it in turn to replenish the reservoir. The resulting ripple, being at twice the A.C. frequency, is easier for the filter to smooth out. The rectifier cathodes are "common", and, as one is always out of use while the other is working, a single cathode can serve both anodes without any increase in emission being needed.

This is sometimes described as a 2-phase circuit (the transformer gives two phases 180° apart); and, though less suitable than *a* for high voltage, is better for large current. It is commonly used with valve rectifiers in receivers.

(*c*) *Bridge.* Circuit *b* yields a terminal which is either positive or negative with respect to the centre tap by a voltage slightly less than half the total transformer peak voltage. By connecting a second pair of rectifiers in parallel with the same transformer winding, but in the opposite polarity, a terminal of opposite polarity is obtained. The total voltage between these two terminals is therefore twice that between one of them and the centre tap, i.e., it is nearly equal to the peak voltage across the whole transformer. The outputs of the two pairs of

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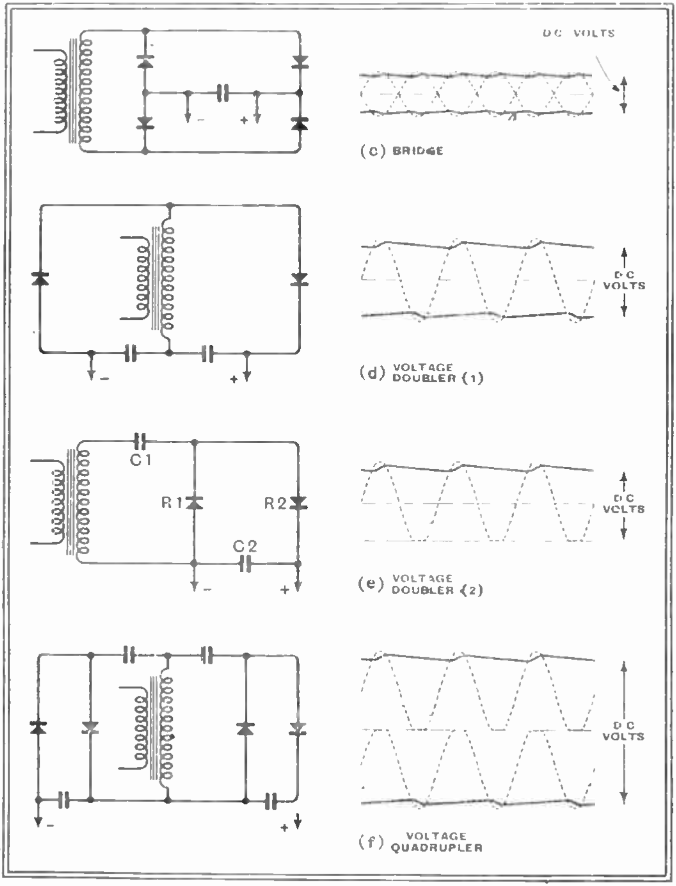

Fig. 186 : Comparison of various rectifier circuits, the voltage supplied from the transformer being the same in all. Note the comparative output voltages

rectifiers are effectively in series. Unless a half-voltage point is wanted, the centre-tap can be omitted.

This arrangement gives approximately the same no-load output voltage as *a*, but with the advantages of full-wave rectification. Unfortunately three separate cathodes are needed,

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so the circuit is not commonly used with valve rectifiers ; it is, however, quite usual with metal rectifiers, especially for low voltages.

(*d*) *Voltage Doubler : first method.* In contrast to *b*, the rectifiers are fed in parallel off the source to give outputs in series. As compared with *a*, the voltage is doubled and the ripple is twice the frequency. The voltage drop on load tends to be large, because each reservoir is replenished only once per cycle.

Like *c* (with centre tap), this circuit gives two supplies, one positive and one negative, from one transformer winding.

This circuit is the commonest for obtaining receiver H.T. with metal rectifiers.

(*e*) *Voltage Doubler : second method.* R_1 and C_1 act, as explained in connection with Fig. 102, to bring the negative peaks (say) to earth potential, so that the positive peaks are twice as great a voltage with respect to earth. R_2 and C_2 employ this as the input of a type *a* system, giving an output which, on no-load at least, rises to almost twice the transformer secondary peak voltage. Compared with *d*, one output terminal is common to the source, which may be convenient for some purposes. As R_2C_2 is a half-wave rectifier this system suffers from the disadvantages of that type, as well as the losses in R_1, so is confined to high-voltage low-current power units.

(*f*) *Voltage Quadrupler.* By connecting a second type *e* circuit in parallel with the transformer, to give an equal voltage output of opposite polarity, the total output voltage tends towards four times the peak of the supply. The conversion of *e* to *f* is analogous to the conversion of *a* to *d*.

207. Filters

We have already used a simple filter circuit (L and C in Fig. 107) to smooth out the unwanted R.F. left over from the rectification process. The same circuit is used in the vast majority of power units, but the components are designed for the very different frequency. To retain a high inductive reactance for impeding the ripple of only 50 or 100 c/s, without at the same time imposing an objectionable amount of D.C. resistance, an iron-cored coil is used. This has to be carefully designed if the core is not to be saturated by the D.C., which reduces the inductance and therefore the effectiveness of the choke coil. We encountered the same problem in Sec. 158, but the solution described there is inapplicable to a power

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smoothing filter, because the current is far greater and loss of D.C volts must be avoided.

A typical smoothing choke for a broadcast receiver might have an inductance of 20 henries at 60 mA. It is a common practice to save a component by making the magnetising coil of the loud speaker (Sec. 172) serve also as a smoothing choke. To prevent the ripple induced in the moving coil by the magnetising coil from causing a hum, a few turns of wire (called a hum-bucking coil) are connected in series with the moving coil and coupled to the magnetising coil in the correct direction to cancel out the voltage induced into the moving coil.

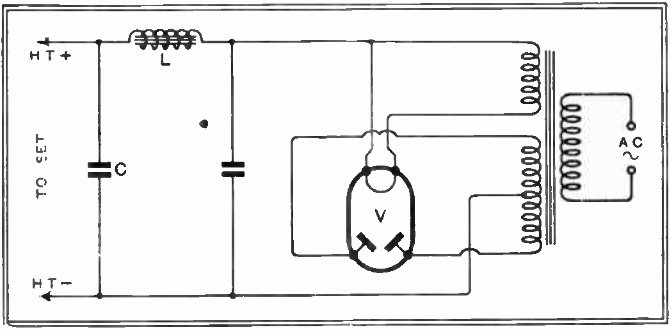

Fig. 187 : Full wave rectifier circuit (*b* in Fig. 186) fitted with smoothing filter LC

The condenser ought to have a very low reactance at ripple frequency, so is generally made large—8 to 16 μF, or even more ; usually of the electrolytic type.

Fig. 187 shows a complete H.T. power supply of the type that has become almost standardised for radio purposes. It consists of transformer, with one secondary to heat the cathode of the full-wave rectifier V, and the other to supply the 2-phase voltage to its anodes. One can recognize Fig. 186*b* here. The output feeds into a reservoir condenser and then through the filter LC.

The transformer is generally furnished with at least one other secondary, to heat the cathodes of the receiver valves. A separate one is needed, because those cathodes are at a greatly different potential from the rectifier cathode. Even if the cathodes are indirectly heated, the slender insulation between them and the heaters cannot stand high voltages as well as high temperatures.

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The same circuit is sometimes used for the power units of transmitters and other equipment.

It has already been noticed that as the current passing through the rectifier to replenish the reservoir has only a fraction of each cycle in which it can do so, it is a correspondingly heavier current, while it lasts, than that drawn steadily off. The voltage lost in the rectifier is therefore exaggerated, and causes the output voltage to fall rapidly with increasing current drain. Where it is particularly desired to minimise this effect, and obtain a voltage that remains reasonably constant under varying loads, the " choke input " filter is used. On paper the only difference is that the reservoir condenser is omitted ; but in practice the design of the choke (called a " swinging " choke) is different, and so is the theory, which is rather more involved. This system is the more usual one for transmitters and high-power amplifiers.

208. Decouplers

Unless the smoothing condenser is uneconomically large, its impedance to signal-frequency currents, looking from the load end, is a good deal higher than that of a battery in good condition. Therefore the varying (signal) currents in one valve stage set up across the H.T. supply a corresponding voltage, which is passed on to all the other valves fed from the same supply, and may seriously upse the working, possibly causing continuous oscillation. We have, in fact, a form of feedback.

To obviate such an undesirable state of things, *decoupling* is used. It is simply an individual filter for those valves liable to be seriously affected. As those valves are generally the preliminary stages in a receiver, taking only a small proportion of the total anode current, a resistor has sufficient impedance at the frequencies concerned. The loss in voltage may actually be desirable, as these stages are often required to be run at lower voltages than the power output stage ; in which case a single cheap component is made to serve the double purpose of a voltage-dropping resistance and (in conjunction with a condenser) a decoupler. Of such is the essence of good commercial design.

These decoupling components are shown as R and C in Fig. 188. The signal currents tend to take the easy path, through C, back to the cathode, rather than through R and the H.T. source. The larger the electrical values of C and R, the more complete the decoupling.

Similar arrangements are used for keeping screens and such auxiliary electrodes at the necessary constant potentials.

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It is the extensive use of decoupling and filtering that renders multi-valve circuit diagrams so alarming for the novice to

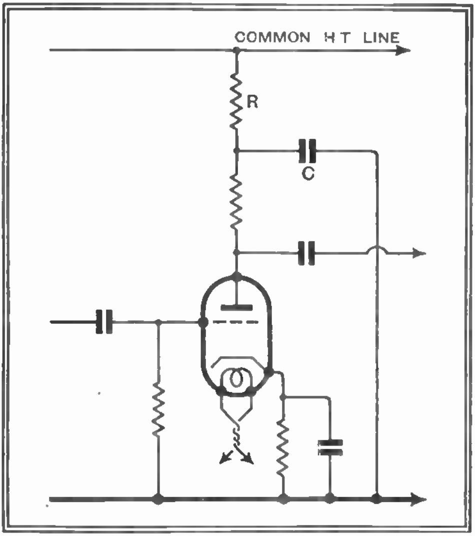

Fig. 188: Decoupling a valve from the H.T. line is performed by inserting R to block signal currents, and providing C to give them a path back to earth

contemplate. But once their purpose is grasped it is easy to sort them out from the main signal circuits.

209. Grid Bias

The most obvious way of applying grid bias is by means of a battery in series with the circuit between grid and cathode; exemplified in Fig. 74. A slight modification is the "parallel-feed" system, shown in Fig. 75, where R is a resistance so high that it causes negligible loss of the signal applied through Cg. A suitable choke coil is sometimes used instead of R; for example, if there is a likelihood of grid current and it is not desired to alter the effective bias by a substantial voltage drop across R.

Quite obviously it would be possible to replace a bias battery by a mains power unit designed to give the correct voltage. Except for high power apparatus such as transmitters this is

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hardly ever done, because the power involved is so small and there are more convenient alternatives, as we shall now see.

If a resistance, R in Fig. 189, is connected between the —H.T. terminal of the anode power unit (or battery) and the valves, the current flowing through it sets up a potential across it, negative at the —H.T. end, which can be applied, wholly or in part, as grid bias to any valves requiring it. R would, of course, be chosen so that when carrying the H.T. current for all the valves it would drop a voltage equal to the greatest grid bias needed. It is necessary to make sure that any de-

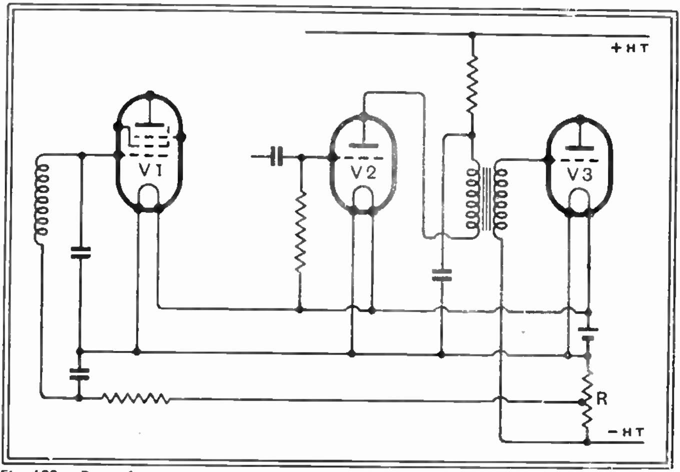

Fig. 189 : Part of a receiver circuit diagram, showing one method of obtaining grid bias, by using the drop in voltage caused by the H.T. current to all the valves passing through a resistor R

coupling condensers are connected to cathodes and not to —H.T., because in the latter case the signal currents going through R would be communicated to other valves, possibly causing instability. Sometimes the individual grid bias leads have little R.F. filters, as shown for V_1; and sometimes the whole bias is filtered before distribution.

It does not matter whether the grid of a valve is made negative with respect to the cathode, or the cathode is made positive with respect to the grid. If a battery set, the universal practice is to feed all the filaments from one battery as in Fig. 182*a*, and so the only way of applying individual bias to the valves is on the grid side, in one of the ways already described : but in a mains set the fact that the heaters are connected together in parallel does not rule out the possibility of the cathodes being connected at moderately different potentials ; up to, perhaps, 50 volts.

Instead of connecting a single resistor in series with all the valves, as in Fig. 189, one can connect separate resistors in series with each valve requiring bias, as in Fig. 190. The fact that more resistors are needed does not necessarily raise the cost, because they need not be nearly so large physically, nor need they be tapped. And the cathode bias system is far more

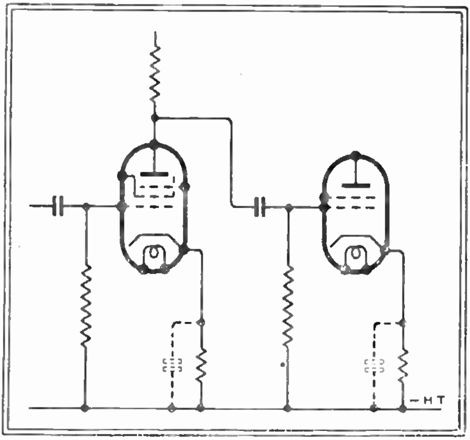

Fig. 190 : A modified bias system, in which each valve has a separate resistor. A large capacity shunt condenser must be used if negative feedback is not wanted

flexible, because valves can be added or removed without affecting the others. Any change in current in any of the valves in Fig. 189 affects the bias for all the others, and it is quite easy for unfavourable working conditions to result. On the other hand, the cathode bias system is self-adjusting. If the bias is too great, the anode current falls and causes less voltage drop in the resistor, reducing the bias ; and *vice versa*.

In certain cases this compensating action may be undesirable ; for example, it is impossible to work under Class B or C conditions, because when the anode current is cut off there is nothing to produce the bias. A convenient method then is to connect the cathode to a suitable point on a potential divider across the H.T. source.

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If only a simple cathode resistor is used, it carries not only the D.C. component of anode current but also the signal current, which produces a voltage in opposition to the grid input. We have, in fact, negative feedback (Secs. 167 and 168). If that is wanted, well and good; but if not, then something has to be done to provide a path of relatively negligible impedance for the signal currents, without disturbing the D.C. resistance. A shunt condenser, shown dotted in Fig. 190, fills the bill; for R.F. a value of 0·1 μF or even less is enough, but for the lowest audio frequencies nothing less than 25 or 50 μF will do. Low-voltage electrolytic condensers are produced for this purpose.

In still another method the input signal is made to provide its own bias. Fig. 191 is the circuit, and Secs. 106 and 107

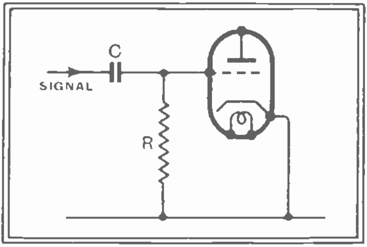

Fig. 191 : When a signal is applied to an otherwise unbiased valve in this manner, it developes its own bias equal to nearly the peak value of the signal, provided that the time constant CR is substantially greater than the period of one signal cycle

explain the theory. It is none other than the "grid detector". The essential conditions are a grid condenser C and "leak" R, such that their time constant (obtained by multiplying C in microfarads by R in megohms) is large compared with the duration of one signal cycle. Thus, with 0·01 μF and 0·5 M Ω, the time constant is 0·005 which is large compared with one cycle if the signal is 1 Mc/s, but not if it is 25 c/s.

The method is particularly useful in a Class C oscillator (Sec. 89) because the absence of bias at the start gives oscillation a powerful send-off, whereupon the resulting bias is proportional to the amplitude of oscillation and tends to keep the latter stable.

If the positive feedback in an oscillator is too large and the time constant is equal to many cycles, the first burst of oscillation may generate such a large bias that oscillation is quenched and cannot resume until the negative charge has leaked off C through R. Oscillation then restarts, is quenched, and so on, intermittently. This phenomenon is known as *squegging*, and is sometimes usefully employed for obtaining modulated

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oscillation. Elsewhere, it is a nuisance, which can be cured by loosening the back coupling and/or reducing the time constant CR.

210. D.C. and Universal Sets

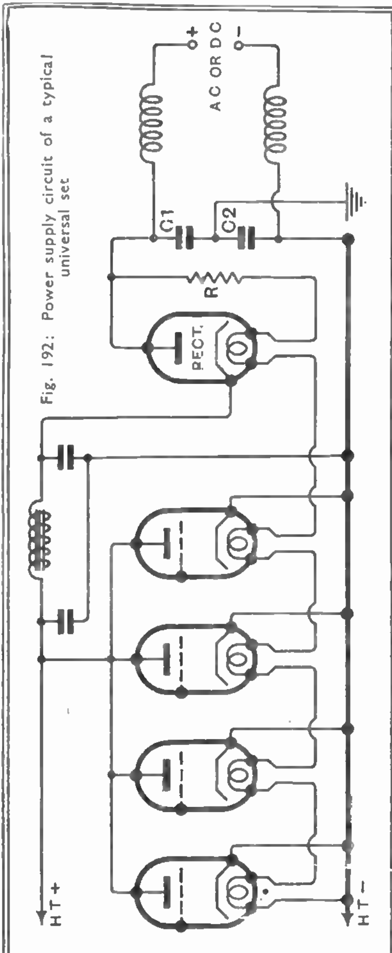

Fig. 192: Power supply circuit of a typical universal set

In receivers intended to be run on D.C. mains, rectification is unnecessary; but owing to the fact that the current is in effect mechanically rectified by the dynamo it contains a small alternating component. To prevent hum, the filter choke and condensers must be included. As D.C. cannot be stepped up by a transformer, the voltage of the supply is limited to that of the mains, generally about 230, so the choke ought to have a low resistance. Because the number of turns needed in a loudspeaker field coil is large, the resistance is too high to be run in series, as described in Sec. 207; in a D.C. set the field coil is connected directly across the mains and a separate choke used.

Sets for use on D.C. only are practically never made, because with slight modifications in design they can be made to work on either D.C. or A.C. and for this reason are called "universal". Their power systems are generally arranged somewhat as in Fig. 192. The heaters of the valves are in series (see also Fig. 182*b*), and in addition there is included in the circuit the heater of an indirectly heated half-wave rectifier. On A.C. mains this is of course essential, while on D.C. mains it is a "passenger", having no effect other than slightly increasing the resistance in the H.T. line. If the receiver is plugged in to D.C. mains in such a way as to make the rectifier anode negative it will not work, and the plug must be reversed in the socket.

The R.F. chokes and small condensers C_1 and C_2 in Fig. 192 are very desirable for preventing high-frequency disturbances due to electrical apparatus connected to the mains from reaching the receiver and causing noises. In an A.C. set the same purpose is served by an earthed screen between primary and secondaries of the transformer.

211. Vibrators

To generate an E.M.F. in a circuit electromagnetically, the magnetic field linked with the circuit must be varying. There is no difficulty about this if the field is produced by an alternating current. That is varying all the time. And so we have the transformer.

If a transformer is connected to a D.C. source, the current, being steady, produces an unvarying magnetic field, and no E.M.F. is generated in the secondary. Neither is any E.M.F. generated in the primary to oppose that of the supply, so there is nothing to restrict the flow of current except the coil's resistance, which is always made as small as possible to minimise losses. The result is likely to be a burnt-out coil or a blown fuse.

The only way of using a D.C. electromagnet to generate an E.M.F. is to get an effect of varying intensity by moving it about, as is done in a dynamo. That requires driving power, so is not nearly so convenient for small units (such as most radio sets) as a static transformer.

We have already noted that H.T. battery power is vastly more expensive than the mains. D.C. mains of 230 volts or so can be utilized at least for receivers; but often the only available supply is a 6-, 12- or 24-volt battery—for example, motor vehicles and other mobile units. If we could convert this low-voltage D.C. into A.C. we could step it up to any desired voltage and then convert it back into D.C., as described in the preceding sections.

It is difficult to convert D.C. into sinusoidal A.C. without the use of rotating machinery, but that is not essential. It can

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easily be converted into A.C. of a sort by interrupting it at high speed by a device working on a similar principle to the electric bell or buzzer. Fig. 193 illustrates this. ACB is the

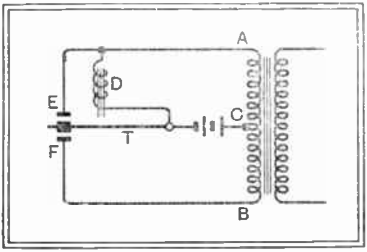

Fig. 193 : Principle of the vibrator method of obtaining A.C. from a battery

primary coil of a transformer. When the battery is switched on, current passes through the half-primary CA and the magnet coil D, attracting the vibrating tongue T towards it and towards the contact E which it strikes, short-circuiting D and allowing a much heavier current to grow in CA (Sec. 29). But meanwhile the current in D is dying away, and so is the magnetism, allowing T to spring back and make contact with F. This causes current to flow through the transformer in the opposite direction, CB, forming the negative half-cycle of current. At the same time current is flowing through CA, but is much less powerful because the impedance of D is in circuit. The whole

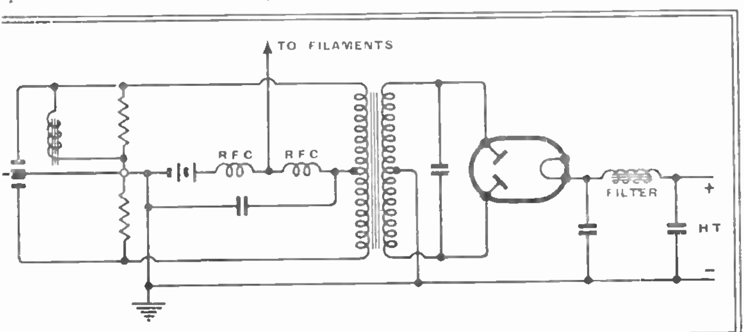

Fig. 194 : Showing how a vibrator is used to supply H.T. for a receiver from the L.T battery

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cycle than repeats, and so on. The frequency is determined chiefly by the mechanical resonance of the vibrating tongue.

Fig. 194 shows this principle adapted to a vibrator power unit; the stepped-up voltage is rectified and smoothed in the usual manner. The various resistors, condensers, and R.F. chokes are included to reduce sparking and absorb high-

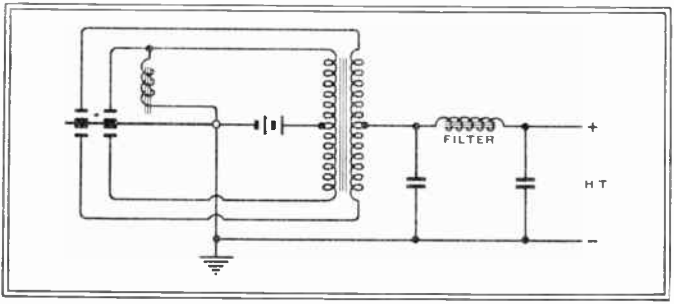

Fig. 195 : Principle of the synchronous or self-rectifying vibrator

frequency currents that would cause interference. It is by no means a simple matter to design a vibrator that is quiet, reliable and long-lived.

As the full-wave rectifier is simply an alternating switching device, some vibrators—the synchronous types—are provided with an extra pair of contacts for rectifying the high voltage output, as shown in Fig. 195. Obviously it has to be carefully designed so that the making and breaking of these contacts occur at the correct parts of the cycle as controlled by the primary contacts.

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CHAPTER 20

RADIATION AND AERIALS

212. Bridging Space—Radiation

IN the first chapter of this book the processes of radio communication were very briefly traced from start to finish; since then we have considered the transmitter and receiver in some detail. But the stages between—transmitter-aerial, space, receiver-aerial—involve ideas that so far have only been hinted at.

It has been implied that even when sound waves have been converted into electrical waves these electrical waves are not of themselves able to travel large distances in empty space, although they can do so along wires. And electric current, as we have seen (Sec. 9), consists of electrons in motion, and these electrons are available in immense numbers in metals but are comparatively absent in insulators such as air. It is difficult to force them to cross even a fraction of an inch of air, so obviously hopeless to expect them to go hundreds or thousands of miles.

But we have seen (Sec. 8) that the space surrounding any electric *charge* is pervaded by a mysterious influence called an *electric field*, which is able to attract or repel other electric charges separated by air, or even by empty space. Also (Sec. 24) an electric *current* causes another mysterious influence, a *magnetic field*, capable of inducing currents in circuits some little distance away.

A *fixed* electric charge, or charges in *steady* motion (an electric current), set up forces in the space around, but these forces diminish very rapidly as one moves away, and at a short distance are too small to detect. This is analogous to a perfectly steady current of air, which may cause a slight local increase in air pressure, but is noiseless. However, when even a slight pressure of air is interrupted, say, 1.000 times per second, it is audible perhaps miles away. It sets up air *waves*. In the same way, if an electric current is rapidly *varied* in strength it sets up waves that radiate outwards through space with the speed of light and die away much less with distance than the induction fields.

An electric current cannot be made to *increase* rapidly for ever, so to keep it varying continuously at a rapid rate it has to be alternately made to rise and fall, as in Fig. 29. The current supplied for lighting, heating and power does this, it is true; but only 50 times per second, and that is too slow to

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cause appreciable radiation. It corresponds to waving a fan slowly from side to side in the air—production of sound waves is negligible. But attach the fan to something that makes it vibrate 1,000 or so times per second and it will be painfully audible. A child's vocal organs can utter piercing sounds because they are high-pitched—many vibrations a second—and are much more audible than even a powerful singer's voice in the extreme bass.

The electric currents resulting from the action of sounds on a microphone are, of course, of the same frequencies as the sounds ; and even when as high as thousands per second they are too slow to stir up waves capable of travelling a useful distance. Hence the elaborate processes of modulation and detection, to make use of a carrier wave having a much higher frequency than the sounds. Although some radio communication is carried on by means of carrier waves alternating a few tens of thousands of cycles per second, radiation is much more complete when the frequency runs into millions

213. Electromagnetic Waves

The waves consist of a strange and complete union of electric and magnetic fields, so complete that each depends entirely on the other and would disappear if it were destroyed ; and are

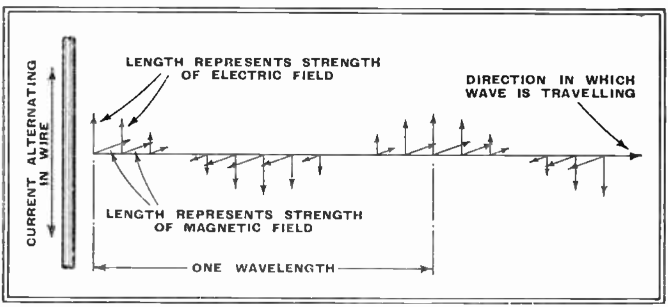

Fig. 196 : Representing one and a half cycles of radiated electromagnetic wave moving outwards from a wire carrying oscillatory current

therefore termed *electro-magnetic* waves. They break away from the circuit that gave them birth, and keep on travelling even if the current in the circuit stops.

If one end of a rope is waggled rapidly up and down, a *vertical* wave travels *along* the rope ; that is to say, at right angles to the direction of waggling. This analogy represents the electric half of an electromagnetic wave. To complete the picture the magnetic part has to be imagined at right angles to both, and therefore in a side-to-side direction. Fig. 196 shows part of a wire in which an electric current must be supposed to be alternating, up and down. Then the electric part of the wave is parallel to the wire and alternates as indicated by the varying lengths of the vertical arrows: the magnetic part is at right angles and is indicated by horizontal arrows (actually drawn as if viewed in perspective); and the combined wave moves outward from the wire and therefore its direction of motion is at right angles to both the fields. It is extraordinarily difficult to depict this three-dimensional phenomenon on paper, for the wave front actually forms a complete cylinder expanding around the wire.

The electric field is measured in volts per metre, if close to the radiating wire, or microvolts per metre if far off. This voltage actually exists in space, but naturally can cause a current to flow only if it impinges on a conductor, such as a wire. The voltage may be considered to be due to the cutting of the wire by the magnetic field (Sec. 27), just as the magnetic field may be considered to be due to the moving electric field (Sec. 24). But a *horizontal* wire will have no voltage in the direction of its length when struck by waves due to a vertical wire radiator; and vice versa. This distinction is known as *polarisation*; the waves due to a vertical wire radiator are conventionally called vertically polarised, that being the direction of their electric field. If a vertical wire half a metre long is held in vertically polarised radiation of strength 50 microvolts per metre, then an E.M.F. of 25 microvolts will be induced in it. But if either the wire or the radiation is horizontal, no E.M.F. is induced. At intermediate angles to the horizontal, the induced E.M.F. is proportional to the sine of the angle. To induce the maximum E.M.F. in a horizontal wire it is necessary for the radiation to be horizontally polarised, which can be arranged by placing the radiating wire horizontally. If radiation starts off with a certain polarisation it does not follow that it will arrive with the same. If it has travelled far, and especially if on its way it has been reflected from earth or sky, it is almost certain to have become slightly disarranged, and at least part of it may be picked up by a wire at almost any angle.

214. The Coil Radiator

It is time now to examine the radiating circuit a little more closely. Up to the present we have considered only a short

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section of wire in which a radio-frequency current is flowing, and have conveniently disregarded the rest of the circuit which is necessary for its existence. In Sec. 76 it was shown how oscillating currents of almost any frequency can be set up in a very simple circuit consisting of a coil and condenser; and, later on, how by suitably connecting a valve to it the oscillations can be kept going continuously. We can fulfil the necessary condition for radiating waves, namely, causing a current to vary extremely rapidly—millions of times a second if need be. Yet such a circuit does not turn out to be a very efficient radiator. What is the trouble?

The difficulty is that for every short section of circuit tending to radiate electromagnetic waves there is another carrying the current in the opposite direction and therefore tending to cancel it out. Each turn of wire in the coil has to go *up* one side and *down* the other in order to complete itself; and, as the same current flows through the whole turn, the wave-producing efforts of the two sides of the turn are in opposition and largely counteract one another. Largely, but not altogether Look at Fig. 197 and imagine it to be one turn of wire, for convenience square. Then you, looking at it, are an equa

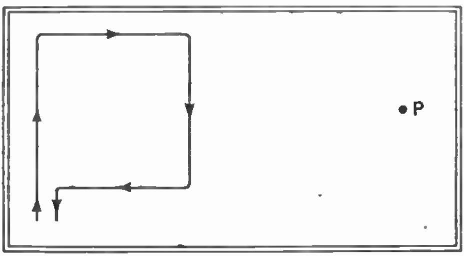

Fig. 197: From the point of view of the reader, the radiation due to different parts of a turn of wire carrying oscillatory current cancels out. Towards the point P, however, the different distances of the vertical sides of the square result in net radiation

distance from equal lengths of wire carrying equal currents and therefore radiating with equal strength; but as the current is always in opposite directions you never receive any net radiation at all. The same argument applies to the top and bottom sections. But now consider it from the point of view P. The down section is nearer than the up, and may therefore rather more than counteract the latter. Actually this disparity is quite negligible if P is miles away. But there is

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another thing to take into account. The radiation from the up section takes time—very little, it is true—to reach the down section on its way to P, and if the oscillations are very rapid indeed the current in the down section may have started to go up; in which case its radiation assists that coming from the previous up wire, and P will get the combined result. Even if the oscillation frequency is not high enough for a complete reversal, it may at least cause sufficient difference between the radiation from the two sections of wire to give a net result at P.

The best result is obtained when the diameter of the coil is half a wavelength (or any odd number of half wavelengths), because then the radiative efforts of both sections pull together in the direction P—and also in the opposite direction. If the coil is 2 inches in diameter—about 5 centimetres—the best wavelength is 10 (or $\frac{10}{3}$, $\frac{10}{5}$, etc.) centimetres, the frequency being 3,000 Mc/s. There are technical difficulties in producing such high frequencies, and even when they are produced the range is extremely limited. For most purposes, then, there are other reasons why the wavelength should be much longer. That being so, the obvious answer is to make the coil larger. As one sometimes wants to radiate waves thousands of metres

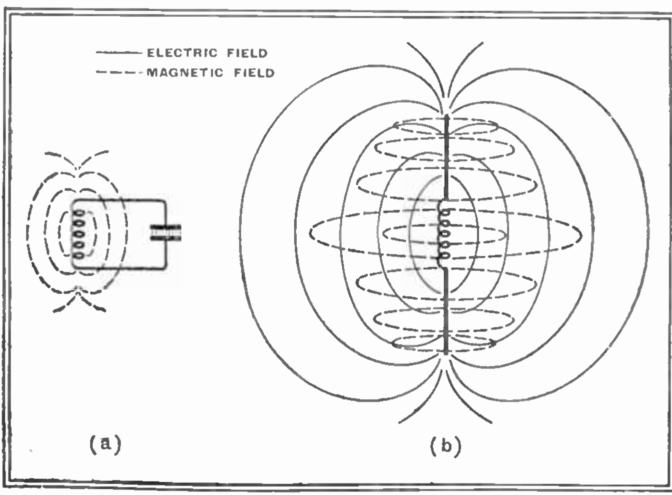

Fig. 198 : In a concentrated tuned circuit (*a*) the electric and magnetic fields are confined and radiation is slight. Opening out the circuit (*b*) extends the fields and also the radiation

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long, this suggestion has its difficulties too, even bearing in mind that it is not essential for the diameter of the coil to be as much as half a wavelength in order to radiate a useful amount. For these and other reasons radiators in the form of a coil are comparatively seldom used.

How about the other element in an oscillatory circuit—the condenser? As long as it is concentrated in a small space its radiating powers are almost negligible (Fig. 198*a*). But opening it out so that the lines of force form large loops, comparable with the wavelength, the vertical wires are carrying current in the same direction in all parts of them; and there is no return path to neutralise the resulting radiation. This arrangement (Fig. 198*b*) is a particularly efficient radiator; and it is actually used in large numbers in the form shown, although for convenience the coil is often separated from it as we shall see.

215. The Condenser Radiator

It is not essential for a circuit to be in resonance in order to radiate, but as the radiation is proportional to the current

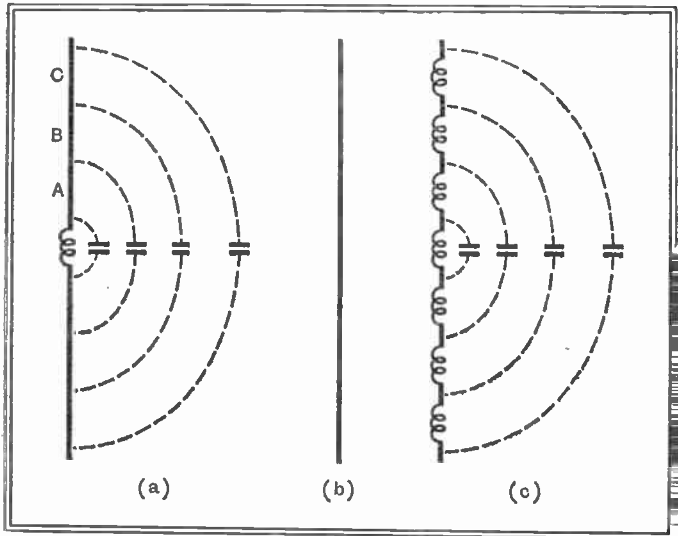

Fig. 199: The distributed capacitance of the radiator in Fig. 198*b* is approximately represented by 198a. A straight wire (*b*) actually has distributed capacitance and inductance, represented approximately by c

flowing, which is a maximum at resonance, the radiator is practically always tuned. So far we have considered tuned circuits made up of a coil and a condenser as in Fig. 198 *a*. These are devices embodying selected quantities of inductance and capacitance in concentrated form. But when the condenser is replaced by long vertical wires as at *b* they possess capacitance which is distributed along their length. It could be approximately represented by a large number of small condensers as in Fig. 199*a*. From this one can see that the current at the point A is greater than that at B, because some of it goes to charge up the section of wire between A and B as well as all that beyond B. Similarly the current at B is greater than that at C, where there is such a small bit of wire to charge up that it requires very little current indeed.

216. The Dipole

As the wires carry this charging current to and fro, pushing an excess of electrons alternately towards the upper and lower halves of the system, they set up a magnetic field; and the wires, as well as the coil, therefore possess inductance (Sec. 28), which is distributed along them. In fact, a tuned circuit can be constructed entirely of a straight wire (Fig. 199*b*), which can be approximately represented for electrical purposes as a number of tiny inductances and capacitances (*c*). There are, of course, moments in every part of the wire, when the current is zero, when it is just about to turn and come back again; but, whereas at the extreme ends of the wire it is *always* zero (because there is nothing beyond to charge), at the middle it alternates with the maximum intensity. This condition can be shown by the line marked I in Fig. 200, in which the distance it stands out from the wire at any point represents the R.M.S. or the peak value of the current at that point.

Each half of the wire becomes charged alternately positively and negatively; but the centre is always midway between these two charges and is therefore at zero potential. The maximum potentials are reached at the ends, for there the charges all along the wire are pushing up behind one another, and the two ends are always at opposite polarity (except instantaneously twice every cycle when they are zero), so the R.M.S. voltage distribution can be indicated by the line V.

As the simple straight wire has both inductance and capacitance, and normally the resistance is comparatively small, it possesses all the qualifications for a resonant circuit, and can oscillate at its own natural frequency. A resonant radiator of

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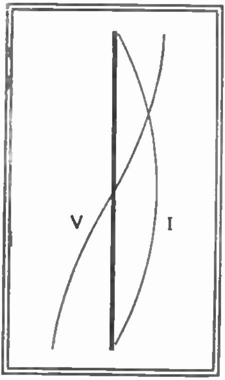

Fig. 200 : Distribution of current (I) and voltage (V) in a half-wave resonant radiator

this kind is known as a *dipole*. It turns out that the lengths of the resulting waves are almost exactly double the length of wire (actually a few per cent. longer than double), and therefore the arrangement is called a half-wave dipole. For short waves, especially those classified as ultra-short (less than 10 metres), the dimensions of a dipole are so compact that it is a very convenient and popular form of radiator, and can be used for either vertically or horizontally polarised waves, according to the way it is erected.

217. Aerial and Earth

The theory that we have just run through has assumed that the wire is suspended in space, far away from any material substances such as the earth. The conclusions are not appreciably upset by placing it in air, but substances having permeability or permittivity greater than 1 increase the inductance and capacitance respectively and cause an increase in wavelength. The effect due to the necessary supporting insulators is very slight if they are kept small. But the effect of the earth is generally considerable. If the desired wavelength is very short, there is no difficulty in suspending the dipole many wavelengths above the ground ; but as soon as radio began to be considered as a means of conveying messages it was found that the very short wavelengths, measured in centimetres rather than metres, have a very restricted range ; and a dipole for much longer waves cannot in practice be more than a wavelength or so above ground and the wire itself may be of unwieldy length. It was Marconi who thought of using the ground to form the lower half of a vertical dipole ; the remaining half sticking up out of it is therefore often known as the quarter-wave Marconi aerial. The current and voltage distribution are shown in Fig. 201, which may be compared with the top half of Fig. 200.

For the earth to be a perfect substitute for the lower half of the dipole it must be a perfect conductor ; which, of course, it never is, though sea-water is a good approximation to it. To overcome the loss due to semi-conducting earth forming a substantial part of the sphere of activity of the oscillating

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current and its resulting fields, it is a common practice to connect the lower end of the radiator—generally called an aerial at this stage of its development—to a radial system of copper wires, buried just below the surface to prevent people

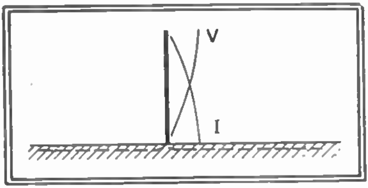

Fig. 201: The earth can be used as one half of a radiator, leaving a quarter-wavelength wire above earth. The dotted line represents buried wires sometimes used to improve the conductivity of the ground

from tripping over them, shown dotted in Fig. 201. This system is called an earth screen; an alternative consisting of an insulated set of wires stretched just over the ground is known as a counterpoise.

218. Aerial Coupling and Tuning

The next necessity is some means of maintaining oscillations in the aerial. Distributed inductance and capacitance, although satisfactory for composing the aerial itself, are not usually

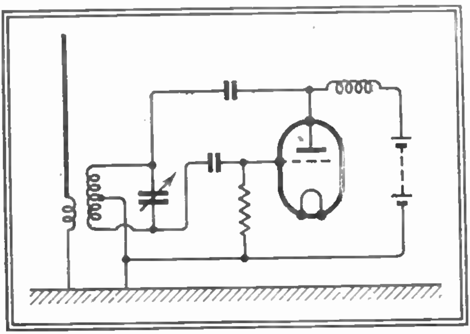

Fig. 202: Valve oscillator inductively coupled to a vertical aerial

convenient for coupling to a source of oscillations such as a valve oscillator. Oscillatory power can be supplied to the aerial by either capacitive or inductive coupling. The latter is

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generally the more convenient, and it is therefore necessary to put back some of the concentrated inductance that we removed when arriving at the simple dipole. The result is shown in Fig. 202. The most effective point at which to insert the coil is that at which the current is greatest. The Marconi aerial has a practical advantage here over the complete dipole, for the coil comes near the ground close to the other apparatus instead of awkwardly in mid-air.

The coil replaces some of the distributed inductance by concentrated inductance, leading to a shorter aerial for a given wavelength and therefore less radiation. To avoid this result, it is possible to neutralise the inductive reactance of the coil by an equal and opposite reactance furnished by a condenser in series with the coil. But in many cases it may actually be desirable to reduce the height of the aerial below its natural quarter of a wavelength: for example, it would be quite impracticable to erect a quarter-wave aerial to work on 10,000 metres—it would have to be about 8,000 feet high !—except by means of a barrage balloon, which would be a menace to aircraft and subject to weather conditions. Often, too, it is desired to adjust the tuning by some more handy means than adjusting the length of the aerial wire. It is quite normal practice, then, for a considerable proportion of the tuning reactance of an aerial system, especially for medium and long waves, to be in concentrated form ; but it must be remembered that this is at the expense of radiation efficiency.

Where there are limitations in height, due to restrictions imposed by flying, or by the resources of the owner, it is possible to increase radiation by adding a horizontal portion, giving the familiar T or inverted L aerials (Fig. 203). The effect of the horizontal extension is to localise the bulk of the capacitance in itself so that the current in the vertical part, instead of tailing off to zero, remains at nearly the maximum value and therefore radiates more (Fig. 203*b*). The top portion radiates, too, but with a different polarisation, so the addition due to it may not be very noticeable in a vertical receiving aerial.

219. Choice of Wavelength

Talking about receiving aerials, the whole of the foregoing principles apply, because the factors that make for efficient radiation are identical with those for the inverse process of deriving the maximum signal power from an electromagnetic wave.

Before considering receiving aerials in detail, however, it

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will be as well to know what factors lead to the choice of wavelength. The shorter the wavelength, the smaller and cheaper the aerial and the more efficiently it radiates. For example, a radio link was operated across the English Channel for several years before the War, on a wavelength of 17 centimetres, a half-wave aerial for which is only about 3½ inches long! Why, then, erect at vast expense aerials hundreds of feet high?

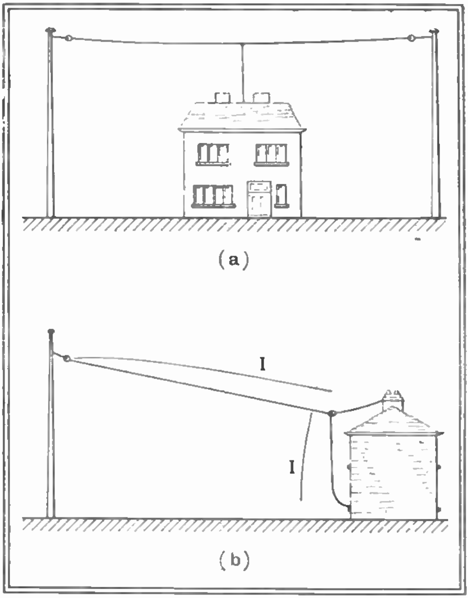

Fig. 203: To increase the radiation from a vertical aerial without increasing the height, a horizontal "top" is commonly added, forming the "T" (*a*) or "inverted L" (*b*)

One reason is that although great advances are being made in the generation of oscillations at very short waves corresponding to frequencies of 300 Mc/s and more, the dimensions of the valves and circuits employed are necessarily very small, and their heat dissipation is correspondingly restricted; there are also other factors that tend to reduce the efficiency; and so only a comparatively minute amount of power can be handled by them. The other main consideration is that even if plenty of power could be radiated it is fairly quickly absorbed, and not being reflected from the sky it fails to reach far beyond the horizon (Fig. 204*a*). So either the aerials have to be put at the tops of very high towers (in which case their cheapness dis-

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appears) or the range is restricted to a few miles. Wavelengths shorter than 10 metres are used for television, not because the shortness of wavelength in itself has any particular merit for

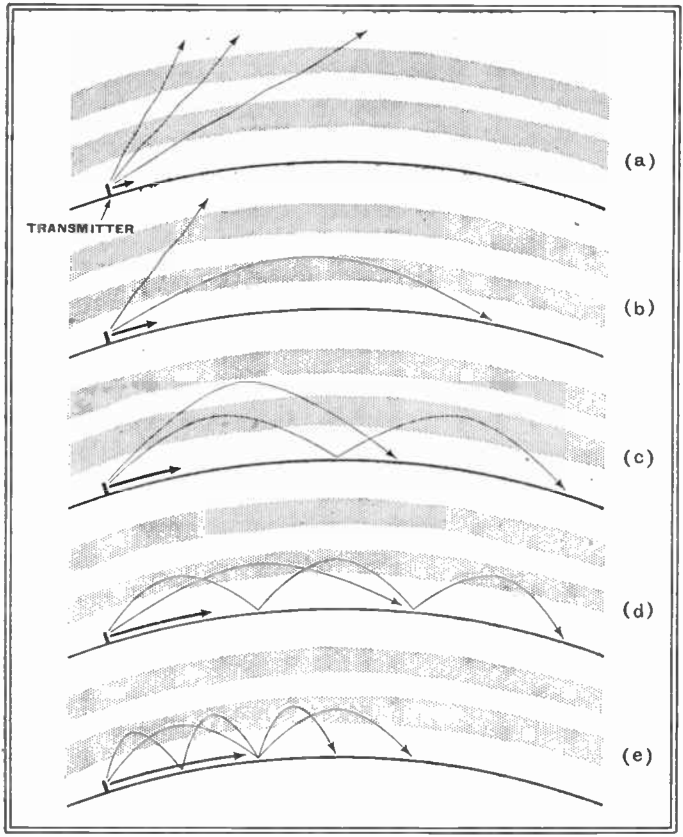

Fig. 204 : Showing (not to scale) the relative ranges of ground wave and reflected wave from very high frequencies (a) to very low (e)

that purpose, but because it corresponds to a very high frequency, which is necessary for a carrier wave that has to carry modulation frequencies of up to several Mc/s. They are also

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used for short-distance communication such as police cars, radiotelephone links between islands and mainland, and other specialised short-range purposes. For a range of even 50 miles a fairly high tower is necessary.

220. Influence of the Atmosphere

As the frequency is reduced below about 30 Mc/s (wavelength greater than 10 metres)—the exact dividing line depends on time of day, year, and solar activity cycle, and other conditions —the range of the wave front travelling along the surface of the earth, and therefore termed the ground wave, increases slowly, while the sky wave is returned to the earth at a very great distance, generally several thousands of miles (Fig. 204 *b*). Between the maximum range of the ground wave and the minimum range of the reflected wave there is an extensive gap, called the skip distance, which no appreciable radiation reaches.

As the frequency is further reduced this gap narrows, and earth reflections may cause the journey to be done in several hops (*c*). As the distances at which the sky waves return to earth vary according to time and other conditions as mentioned, it is rather a complicated business deciding which wavelength to adopt at any given moment to reach a certain distance. But a vast amount of data has been accumulated on this and enables fairly reliable communication to be maintained at all times by a judicious choice of wavelength. As waves usually arrive at the receiver by more than one path simultaneously, and tend to interfere with one another, fading and distortion are general unless elaborate methods are adopted for sorting the waves out. At a certain wavelength, of the order of 150 metres, the ground wave and reflected wave begin to overlap *at night*, while during daylight the reflected wave is more or less absent. Over the ranges at which there is overlap the two waves tend to interfere and cause fading and distortion, as they do with more than one reflected wave.

Finally, the range of the ground wave increases and becomes less affected by daylight or darkness, so that waves of 10,000 and more metres have a range of thousands of miles and are not at the mercy of various effects that make long-distance short wave communication unreliable. For this reason they were originally selected as the only feasible wavelengths for long ranges, and are still used for that purpose; but certain serious disadvantages have forced radio engineers to make the best of shorter waves, and now only a small fraction of long distance communication is borne by very long waves. The disadvantages of the latter are (1) the enormous size of aerial

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needed to radiate them; (2) the low efficiency of radiation even with a large and costly aerial system; (3) the higher power needed to cover long ranges, largely due to (4) the great intensity of atmospherics—interference due to thunderstorms and other atmospheric electrical phenomena; and (5) the very limited number of stations that can work without interfering with one another, because the waveband is so narrow in terms of frequency—which is what matters; see Sec. 148.

221. Classification of Frequencies

The following table, which summarises the foregoing, is due to Prof. F. E. Terman:

Class *	Frequency range kc/s	Wavelength range : metres	Outstanding characteristics	Principal uses
Low frequency.	Below 100	Over 3,000	Low attenuation at all times of day and year.	Long-distance trans-oceanic service requiring continuous operation.
Medium frequency.	100–1,500	3,000–200	Attenuation low at night and high in daytime; greater in summer than winter.	Range 100 to 500 kc/s used for marine communication, aeroplane radio, direction finding, etc. Range 150 to 300 and 550 to 1,500 kc/s employed for broadcasting.
Medium high frequency.	1,500–6,000	200–50	Attenuation low at night and moderate in daytime.	Moderate-distance communication of all types.
High frequency.	6,000–30,000	50–10	Transmission depends upon the ionisation in the upper atmosphere, and so varies greatly with the time of day and season. Attenuation extremely small under favourable conditions.	Long-distance communication of all kinds; aeroplane radio.
Very high frequency.	Above 30,000	Below 10	Waves travel in straight lines and are not reflected by ionised layers, so can only travel between points in sight of one another, or nearly so.	Short-distance communication; television; two-way police radio; portable equipment, aircraft landing beacons.

* Not to be confused with official classification of frequencies: see p. 332.

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222. Beams and Reflectors

Because of the disadvantages, already mentioned, of very long waves, short waves are used wherever possible. A further point in their favour is that it is practicable to concentrate the radiation in any desired direction instead of wasting a large part of the power by indiscriminate distribution. When

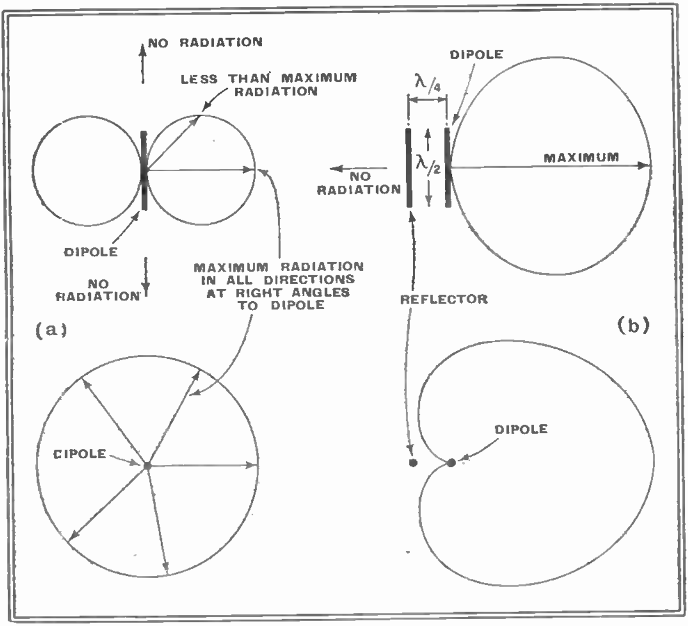

Fig. 205 : Polar diagrams of radiation from a simple dipole aerial (a) ; side view above, end view below, and dipole with reflector (*b*)

the wavelength is much greater than the dimensions of the aerial, radiation takes place fairly equally in all directions, and it is difficult to modify this. But when the aerial is half the wavelength, as in the simple dipole, the radiation varies from nil along its axis, to a maximum all round its "equator" as indicated by the line drawn around the dipole in Fig. 205*a*. The distance of any point on this line from the centre is an indication of the relative strength of radiation in that direction. If a second half-wave dipole, not fed with power, is placed quarter of a wavelength away, as shown at *b*, it is like a resonant

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coupled circuit and has oscillatory currents induced in it. These currents re-radiate, and the quarter wave spacing causes the re-radiation to be in such a phase as to cancel out the original radiation on the side where the second dipole is placed, and to reinforce it on the opposite side. The second dipole therefore acts as a *reflector*. The same argument applies to a receiving aerial; and by using reflectors at both transmitting and receiving ends there is a considerable gain in signal strength.

But this is only the beginning of what can be done. By placing a number of dipoles in a row, or end-on, or both, the radiation from them adds up in phase in certain directions, and cancels out in others, and in general the greater the number of

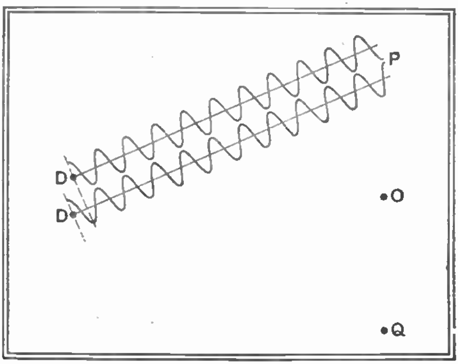

Fig. 206: Showing how the waves from two dipoles radiating in phase arrive at certain angles in completely opposite phase

dipoles in such an array the narrower and more intense the beam of radiation Obviously the size of such an array would be impracticable for long waves, whereas for very short waves it is reasonably compact. This compensates to a large extent for the difficulties in generating large amounts of power at very high frequencies; and in fact the short-wave beam system for Empire communications was just in time to prevent a large sum of money being spent on a grandiose scheme for high-power long-wave stations.

The principle on which most of the directional aerial arrays, however elaborate, depend can be illustrated by a pair of dipoles shown end-on at DD in Fig. 206. They are fed with R.F. power of the resonant frequency (i.e., wavelength equal to twice the length of the dipoles) *in phase*, so that at any point O equidistant from the two dipoles the radiation from both

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arrives in phase and gives stronger reception than from one dipole only. But points such as P or Q are at unequal distances from the dipoles, and if the difference in distance happens to equal half a wavelength—or any odd multiple of half a wavelength—the radiation from the two arrives in opposite phase and cancels out. By using a sufficiently large number of dipoles, suitably arranged, the radiation can be concentrated into a narrow angle.

It is necessary to consider, not only the distribution of radiation horizontally around the transmitter, but also the angle of elevation about the horizon. In this the ground has a large influence, and the height of the aerial above ground is an important factor. Obviously if the direct ground wave is being relied upon it is wasteful to direct most of the radiation upwards; while if reflected waves are necessary to reach the desired destination the upward angle should be adjusted accordingly. The design of an aerial array is therefore far from simple, especially as some of the data—conductivity of earth, and reflecting power of atmosphere, for example—are imperfectly known and subject to change.

223. Direction Finding : Frame Aerials

An important branch of directional wireless is direction finding, commonly abbreviated D.F. It is possible to make use of the fact that dipoles receive nothing when pointing towards the source of radiation to discover the direction in which the source lies. As Fig. 205*a* shows, it may be in either of two opposing directions; so to distinguish between these a reflector is generally used, as at *b*. If two or more receivers observe the bearing of the transmitter, its position on the map can be determined by drawing lines at the appropriate bearings from the receiving sites and noting where they intersect.

It is more usual, however, to use for directional receiving purposes a coil or frame aerial, which as we saw in Sec. 214 transmits and receives nothing broadside on, and in fact has a characteristic similar to that of Fig. 205*a* but turned through 90°. It has the advantage of being quite easy to tune over a considerable waveband.

The subject of direction finding is a large and complex one, as can be gathered from the fact that Keen's well-known book on it runs to 800 pages, so the details cannot be adequately discussed here; but it may be noted that directional receiving aerials—notably the frame type—are of value as a means of avoiding interference. No amount of selectivity in the receiver

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is able to separate two stations working on the *same* frequency; but if they are situated in different directions from the receiver a frame aerial can be set to cut out the unwanted one. This setting is fairly critical, whereas the strength of reception varies little over a wide range of angle each side of maximum; so it is possible to discriminate between waves arriving from not very different directions. Another glance at Fig. 205*a* will

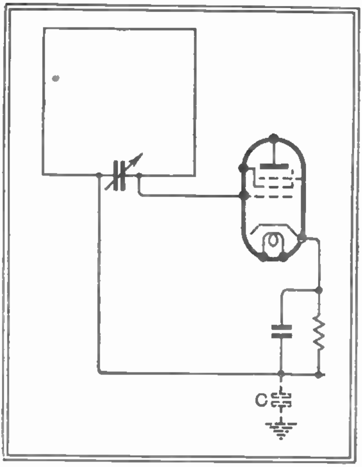

Fig. 207: Due to its capacitance to earth, a frame aerial is liable to act also as a vertical aerial, receiving a signal even when turned broadside to the source

make this clear. Most portable sets use frame aerials, and as some of this type have very poor selectivity it is just as well!

Unless special precautions are taken, such as electrically balancing the aerial with respect to earth, it will be found that a strong signal cannot be made to vanish at any setting of the frame. The reason for this is that besides acting as a coil it also acts as an ordinary vertical aerial, which has no directional properties. In Fig. 207 a wave striking the frame aerial broadside induces equal voltages in both vertical parts of it, and so far as sending a current *around* the tuned circuit is concerned they cancel out. But besides acting as a coil the frame has a certain amount of capacitance to earth just like any elevated aerial, as indicated by the imaginary condenser C; and the induced voltages are in parallel for causing currents to flow up and down between aerial and earth. Some of the current sets up a potential difference across the tuning condenser and this is amplified by the receiver. To avoid this the frame must be balanced to earth ; and Fig. 208 indicates one

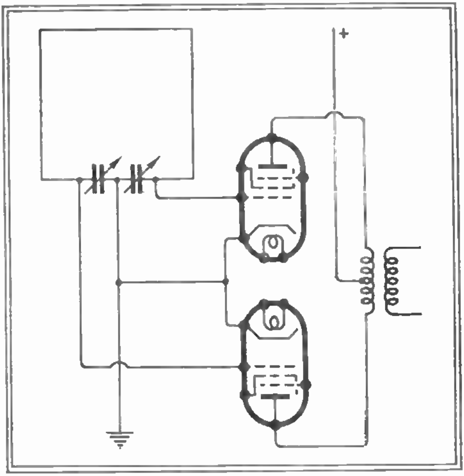

Fig. 208 : One method of avoiding "vertical" effect, by means of a balanced circuit

of several ways in which this can be done. Potentials to *earth* imposed on the grids of the two valves are equal and of the same polarity, and can be eliminated in the output by a push-pull connection. Potentials resulting from true coil action are in opposite phase and so are amplified.

224. Indoor Aerials and their Disadvantages

The frame aerial is not favoured for transmitting because unless it is inconveniently large its radiating properties are poor. Similarly, it is not very efficient as a receiving aerial, and can only be used when either the field strength of the radiation is very great, or a large amount of amplification is employed. This drawback is increased by the fact that it is generally inconvenient to mount a frame aerial out of doors, and indoors it is more or less screened off from signals, while being in an excellent position for picking up undesired noises due to electrical equipment of many kinds—fans, cleaners, razors, etc.—used in houses. The receiver itself also inevitably contributes to noise when amplification is pressed to its limit.

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The same applies even more forcibly to what is known as the mains aerial. This consists of a condenser of about 100 pF connected from the R.F. input of the receiver to one side of the twin mains lead used to supply power to the set. This lead, and the wiring of the house, act as an aerial of a rather inefficient kind. It is most effective, however, in conveying electrical noise to the receiver, originating perhaps from many premises in the district.

A short piece of wire suspended around the room, usually concealed along the picture rail, is little better, for it is closely coupled to the electrical wiring. These indoor aerials, unless used merely as temporary expedients or in situations where an outdoor aerial is impossible, are generally a sign of (*a*) laziness, (*b*) dislike of the unsightliness of a more conspicuous aerial, or (*c*) a feeling that use of an outdoor aerial indicates possession of a cheap insensitive receiver and hence social inferiority. Many high-class receivers give very poor results because they have such a reserve of amplification that it is considered that "any aerial or no aerial will do". Actually the important thing is not the strength to which the incoming signal can be amplified—that is generally quite easy—but the ratio between it and "noise". Use of a large efficient outdoor aerial enables the amplification to be cut down and noise reduced.

It is a common fallacy that use of a large aerial reduces selectivity. Actually it enables the coupling between aerial and first tuned circuit to be reduced, which reduces the damping of the circuit and so improves its selectivity, besides making it easier to gang it with other tuned circuits in the receiver (Sec. 118).

225. Anti-Interference Aerials

The increasing use of electrical equipment, much of which causes noise in radio receivers, has led to widespread adoption of anti-interference aerials, the principle of which is to erect the part of the system which is effective for receiving as far away as possible from domestic and industrial sources of electrical noise, and to screen the lead-in and indoors portion, which otherwise is relatively less good for station reception but contributes the greater part of noise pick-up. The higher the aerial proper, the greater the desired signal and the less noise it is likely to pick up. When the level is reached below which any benefit from increased signal reception is likely to be more than outweighed by increase in noise, the whole of the wire from here to the first tuned circuit in the receiver should be

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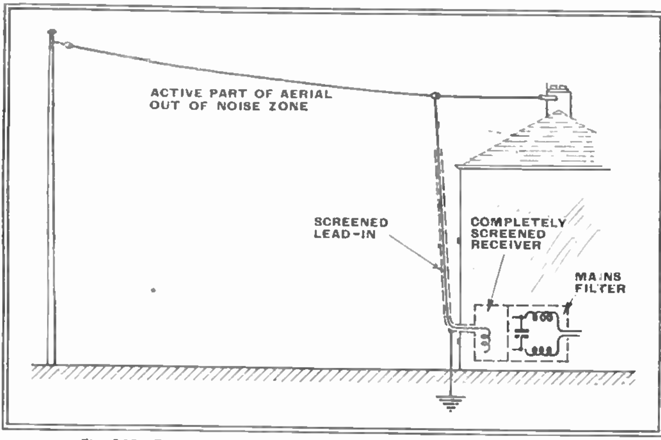

Fig. 209: Diagram of a screened-downlead anti-interference aerial

completely screened. The receiver circuits themselves are assumed to be screened, and it is necessary to stop the ingress of noise from the mains lead by interposing a R.F. filter—usually a choke coil in series with each wire, and a condenser shunted between them. Whether it is better for the condenser to be on the mains side or receiver side of the chokes depends on local conditions.

Schematically the foregoing arrangement is as shown in Fig. 209. The aerial downlead and lead-in are surrounded by a braided metal sheath, which is earthed. Insulation between wire and sheath is, of course, essential, and consequently the capacitance to earth of the screened portion is very large—far larger than that of the unscreened part. It is like connecting a high-capacitance solid-dielectric condenser across aerial and earth terminals, as in Fig. 210. The dielectric, unless very carefully selected, causes serious R.F. loss (Sec. 62); but, even if it did not, the capacitance shunts a large part of the precious signal current

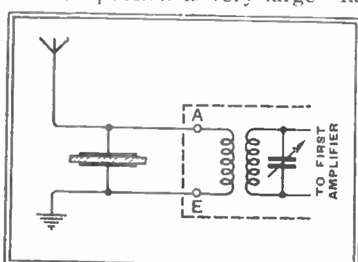

Fig. 210: Approximate electrical equivalent of the screened downlead, causing loss of signal

away from the receiver. The higher the frequency, the lower the reactance and the greater the loss. And, of course, the transfer of so much capacitance to the tuned circuit upsets the tuning if, as is usual, it is ganged.

226. Matching the Impedances

The answer to this problem is in Sec. 48, where the rule was given for calculating the effect, on one winding of a transformer, caused by connecting a resistance across the other. The same holds good for reactance. Suppose, for example, the capacitance of the screened downlead is 1,600 *p*F. At 1 Mc/s that is a reactance of about 100 ohms (Sec. 36). If the transformer between aerial and first tuned circuit were a close-coupled one with a 1 : 1 ratio the effect would be to add 1,600 *p*F to the tuning condenser, which would be perhaps 10 times the normal tuning capacitance for that frequency, and the voltage developed across it would be very much reduced. But if it were a 1 : 10 ratio transformer the equivalent reactance across the secondary would be 10^2 or 100 times as much—10,000 ohms, which is the reactance (at 1 Mc/s) of 16 *p*F, or only one hundredth part of the capacitance of the aerial downlead. Actually it is quite unnecessary to work out the reactance, for the secondary capacitance is reduced in the same proportion as the reactance is increased—the square of the transformer turns ratio—and, of course, is the same at all frequencies, apart from certain effects neglected by the assumption of an ideal transformer.

There is a limit to the transformer ratio that can be advantageously used in the endeavour to minimise the effect of the large shunt capacitance of the aerial downlead. The primary current increases as the step-up ratio is increased ; and there comes a point where the loss due to this current having to pass through the series impedance of the aerial more than counterbalances any gain due to step-up ratio. It is another case of *matching* the aerial impedance to that of the tuned circuit, as in Sec. 117. The impedance of the type of screened lead generally used is about 80 ohms, regardless of frequency ; whereas that of the tuned circuit varies considerably with frequency, being possibly well over 100,000 ohms at low radio frequencies and 5,000 ohms or less at the very high. So the transformer ratio is a compromise, as we saw in Sec. 118 ; a ratio of from 1 : 10 to 1 : 25 is of the right order except for very high frequencies.

How about the connection between aerial and screened downlead ? The impedance at medium and low frequencies

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of an ordinary non-resonant aerial (as distinct from the dipole or the quarter-wave earthed aerial, which have to be a different length for each wavelength) is also very variable and indeterminate, but it is almost certain to be very much greater than 80 ohms; and unless there is to be another bad impedance match with consequent loss of signal strength a step-*down* transformer is needed at this point.

The design of such a transformer to cover low, medium, and high radio frequencies is very involved indeed, and compromise reigns almost supreme; but manufacturers have managed to produce transformers suitable for both top and bottom ends of the screened section (the latter for use where the receiver aerial circuit is not already specially designed for the purpose) in which the whole of the broadcast wavebands are covered; and, although some loss of signal is inevitable, it is not as great as the reduction in noise is likely to be, and so a net gain in signal-noise ratio is shown.

227. Radiation Resistance

The reason why the impedance of an aerial changes widely as the frequency is varied is that the reactance of its inductance and capacitance both depend on frequency; and it is only when it resonates that the two are equal and opposite, leaving only the resistance to take account of (Sec. 51). In this respect it is like a circuit in which the tuning is not varied to resonate with the applied frequency. If an aerial is required to work on only one frequency, then it is possible to adjust its length to resonate. We have already seen that the dipole type of aerial resonates when it is very nearly half a wavelength long, and the vertical earthed aerial when it is approximately quarter of a wavelength, assuming no "lumped" inductance or capacitance is added. The latter enable aerials of the wrong length to be adjusted to resonance.

When the aerial does resonate, its impedance consists of resistance only, just as in an ordinary tuned circuit. This resistance, when measured at the resonant frequency, is always found to be greater than one would expect from the gauge of wire and known losses. The reason is that, unlike the concentrated form of tuned circuit, it radiates an appreciable part—perhaps a very large part—of the applied radio-frequency power. Although the power is not used up in heating the aerial it is removed from it, and the number of watts radiated as electromagnetic waves is proportional to the square of the current just as is the number of watts lost in heat. The heat loss, W watts, in a resistance R, is I^2R; so by this the resistance

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is $\frac{W}{I^2}$. If the radiated power is W_R watts, then $\frac{W_R}{I^2}$ can be looked upon as a quantity similar to a resistance, which, when multiplied by the square of the current, gives the number of watts radiated. It is actually called the *radiation resistance*. Obviously an efficient aerial is one in which the radiation resistance is a large proportion of the total resistance, so that most of the applied power is radiated and not much is wasted as heat. Such an aerial is generally good for receiving as well as for transmitting.

A dipole, for example, if made of copper rod or tubing, has an extremely low *loss resistance* ; but owing to its open fields the radiation resistance is quite large, and quickly damps out any oscillation that is started in it without means for maintenance, just like a highly damped oscillatory circuit of the concentrated kind, except that most of the oscillatory power leaves the circuit as radiation and not as heat.

228. Aerials as Resonant Circuits

The resistance of a concentrated tuned circuit is that which would be found by opening it at *any* point and measuring (at the resonant frequency, of course), between the terminals as in Fig. 211 *a*. The two reactances—inductive and capacitive—cancel one another out at resonance ; and the resistance normally being small, a small voltage injected may cause quite a large current to flow. Similarly a dipole can be opened at the centre, where the ratio of current to voltage is large ; and when voltage at the resonant frequency is applied the distributed inductance and capacitance cancel one another out. The loss resistance is relatively small, and the radiation resistance is found to be nearly the same for all resonant dipoles kept far from earth or other bodies and supported by a minimum of insulation. The figure is between 70 and 80 ohms. It may be remembered, incidentally, that this is a normal impedance for a screened aerial downlead, so it is unnecessary to use a transformer for matching one to the other.

When a resonant circuit is connected in parallel (Fig. 211 *c*), the currents through the two branches tend to cancel out ; in fact if it were not for resistance it would be unnecessary to feed any current in from outside to keep oscillation going, once started. A normal tuned circuit—resistance small compared with reactance—behaves as a high resistance when connected in parallel (Secs. 56–60), in many practical cases 100,000 ohms or more. The lower the series resistance, the higher the

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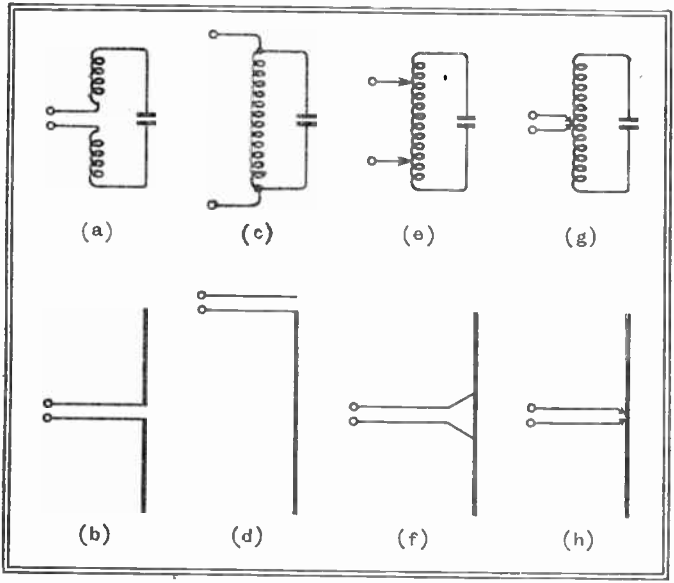

Fig. 211: The various methods of connecting a dipole, shown in the lower row, are equivalents of the connections, shown in the upper row, to a concentrated tuned circuit

dynamic resistance. Similarly, a dipole, being really a tuned circuit, presents a high resistance across its ends, but, owing to its substantial radiation resistance, its end resistance is not nearly so high as a low-loss circuit; and, in practice, it is not fed across its two ends, for the leads spaced so far apart would themselves radiate substantially and modify the whole action of the system. A practical alternative is shown at *d*, which looks absurd until it is realised that owing to the distributed nature of the dipole's fields its centre is at zero potential and therefore can be regarded as "earth" even when not so connected. The second wire of the feeder carries a current in opposite phase which partly cancels out the radiation from the first. Used in this way the dipole presents a dynamic resistance of about 3,000 ohms.

Just as the dynamic resistance of a tuned circuit between the connecting leads can be reduced by "tapping down" (*e*), so too can a dipole (*f*). Points can be found presenting a resistance of several hundred ohms, suitable for matching a feeder line of that amount. As we saw in the preceding Section, and Sec. 48, *e* is equivalent to a step-up transformer. In either case—tuned circuit or dipole—the reactances in both branches between the tapping points cancel out, but they are less in amount and the dynamic resistance is less accordingly. Going to extremes, when the tappings are brought closer together the reactance between them dwindles and finally reaches zero when they coincide (*g* and *h*).

Except when used in push-pull, one end of the coil and of the condenser are at earth potential, and so it is more usual to connect one of the tappings there; but as the centre of a dipole is the zero potential point it is normally treated as a push-pull circuit, with balanced connecting lines.

When a dipole, or any aerial, is used at other than the resonant frequency, the impedance is not just resistance; it includes either inductive or capacitive reactance, so in general is larger; and besides the loss of efficiency due to being out of tune there is a mismatch between it and the connecting feeder.

Note on Section 221. Radio frequencies have been officially classified so as to give standard meanings to the terms "low," "high," etc., in that connection, as follows:—

Class	Abbrevn.	Frequencies	Wavelengths
Very Low	V.L.F.	Below 30 kc/s	Over 10,000 metres
Low	L.F.	30–300 kc/s	1,000–10,000 metres
Medium	M.F.	300–3,000 kc/s	100–1,000 metres
High	H.F.	3–30 Mc/s	10–100 metres
Very High	V.H.F.	30–300 Mc/s	1–10 metres
Ultra High	U.H.F.	300–3,000 Mc/s	10–100 cms
Super High	S.H.F.	3,000–30,000 Mc/s	1–10 cms

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CHAPTER 21

TRANSMISSION LINES

229. Feeders

ANTI-INTERFERENCE aerials have already given us an example of having to provide some means of connecting an aerial which is up in the air to a receiver indoors and close to the ground. Dipole aerials are another; and so are the more elaborate systems used for directional purposes. The connecting links are called feeders or transmission lines. It is generally desirable that they shall not themselves radiate or respond to radiation, because that would modify the designed directional effect of the aerial itself, or introduce undesired interference. This object is easily achieved, in principle at least, by placing the go and return wires very close together—the parallel wire feeder—so that the two radiations or receptions nearly cancel out; or, better still, enclosing one lead completely within the other—the coaxial feeder.

In either case, but especially the latter, the closeness of the two leads means a high capacitance between them; and hence a low impedance. It might appear at first sight that as the impedance of a condenser decreases steadily as the frequency increases, so would the impedance of a feeder. But this line of argument (which we temporarily adopted in Sec. 226) ignores the *inductance* of the feeder, which behaves in the opposite manner; and it happens that *under certain conditions* the two effects exactly balance one another, and the impedance of a feeder or transmission line is practically the same at all frequencies, and dependent only on the diameters and spacing of the wires or tubes employed. Moreover, under the same conditions this impedance is entirely resistance, and is equal to the resistance of the load connected to the far end.

Beginners usually have difficulty in seeing how the resistance depends on these things and not on the *length*, which one is accustomed to regard as the most important factor in reckoning the resistance of a piece of wire. It seems quite absurd to say now that whether the line is long or short has nothing to do with its resistance. The reason for this paradox is that at very high frequencies the actual resistance of the wire or tube, which is what depends on its length, is almost or entirely negligible in comparison with its inductance and capacitance. The effect of length is generally taken into account in a different way, by saying that it causes a loss of such and such a proportion of the signal voltage per 100 feet (or other unit of length). Obviously

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a very thin wire, of comparatively high resistance, will cause a greater loss of this kind than low-resistance wire.

To understand exactly what is meant by this new sort of resistance or impedance, which is independent of length, involves mathematics of a higher order than is assumed in this book. But the following explanation, although somewhat crude, may perhaps succeed in conveying some idea of the meaning.

230. Waves along a Line

The inductance and capacitance of a parallel or a coaxial line are distributed uniformly along it, and so cannot be exactly represented by conventional symbols. But Fig. 212, repre-

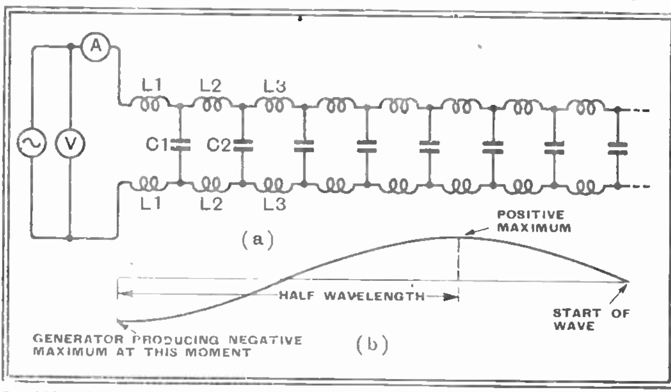

Fig. 212 : A parallel wire line is approximately equivalent to the system shown at *a*, in which L and C are very small and very numerous. After the generator has turned out three quarters of a cycle the situation is as shown at *b*

senting a short section of the beginning of a line, is an approximation to it. A high-frequency generator supplies power, and a voltmeter and ammeter measure the R.M.S. voltage and current. Suppose the line is infinitely long, or, if that is too great a stretch of the imagination, so long that we have a little while to examine conditions at the starting end before the wave has had time to travel the whole way to the distant end.

What happens during the first positive half-cycle is that current starts flowing through L_1 and charging up C_1. The inductance prevents the current from rising to its maximum exactly at the same time as the voltage maximum, and it keeps it flowing a little while after the first half cycle is over. In the meantime, the inertia of L_2 to the growth of current through

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it allows a charge to build up in C_1; but gradually current gets moving in L_2, and so on, passing the positive half-cycle along the line from one section to another. Meanwhile, the generator has gone on to its negative half-cycle, and this follows its predecessor down the line. Assuming the line and its surroundings are of non-magnetic material and are free from solid (or liquid) insulating material (in other words, the permeability and permittivity are both equal to 1) the wave will pass down the line with the speed of light or electromagnetic waves in space. So if, for example, the generator frequency is 50 Mc/s,

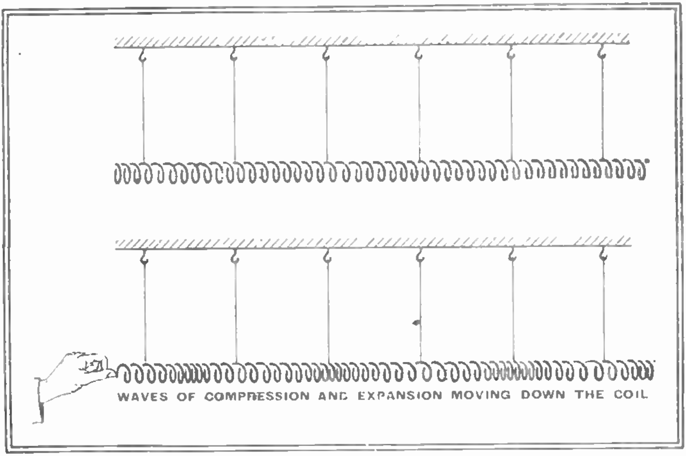

Fig. 213 : Mechanical equivalent of an electrical transmission line. When the "input" is moved rapidly in and out, waves travel along the coiled spring

each cycle lasts for one fifty millionth of a second, and as the wave travels at 300 million metres per second the single wave has gone 300,000,000/50,000,000 or 6 metres down the line. That is, of course, equal to the wavelength. When the current is a maximum at one point along the line it is a maximum in the opposite direction 3 metres either side (Fig. 212 *b*).

A mechanical analogy is provided by a long coil of springy wire suspended horizontally. If one end is moved to and fro along the axis, visible waves of compression and expansion of the spring are set in motion along the coil to its far end (Fig. 213). At the moment that one part of the spring is being compressed, parts either side are expanding.

To supply to the starting end of the line the alternating charges that are passed down its length demands from the generator a certain current. Note that the *length* of the line has nothing to do with the strength of current needed, so long at least as the front of the wave has not had time to reach the distant end. Both this current and the voltage that drives it are indicated on the meters, and as the ratio of voltage to current is impedance, we arrive at a figure of impedance which is independent of length, and depends only on distributed capacitance and inductance, which in turn depend on spacing and diameter of the wires.

Let us now go, very quickly, to some point along the line and wait for the wave to arrive. If a suitable voltmeter and ammeter are provided at this point we can measure the voltage and current; and, assuming that the resistance of the wires composing the line, and the leakage between them and the radiation, are all negligible, then the current and voltage are the same as at the start. If there has been no loss on the way, there can have been no reduction in the power sent by the generator.

231. Surge Resistance

Hurrying along to the distant end of the line, we wait for the wave there. Here it comes, the same voltage and current; until it gets right to the end. Then, if the end is open circuited, there can be no current there. And if it is short circuited there can be no voltage. But suppose we connect a resistance equal to the already measured line impedance, then the ratio of voltage to current exactly satisfies Ohm's Law for that particular resistance, and it absorbs the whole power just as fast as it arrives. As the current going into it and the voltage across it are the same as at the beginning, or indeed any other point on the line, the resistance is equivalent to an infinitely long line of the same impedance. The generator would not know whether the wave started by it was still going on for ever and ever, trying to reach the end, or was being quietly dissipated by a resistance quite a short distance away, *provided that that resistance is equal to the impedance of the line.*

If when the generator supplies 500 volts a current of 1 amp. (alternating, of course) flows into the line, then it is reasonable to say that the impedance of the line is 500 ohms.

In arriving at this result the *length* of the line has not been taken into account at all, for the current starts flowing before the resulting wave has reached the far end. If now a resistance of 500 ohms is connected at that end, when the wave does reach it 1 amp. will flow due to the 500 volts, just as if the

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resistance were an infinite length of line. This being so, the impedance of the line acts towards the generator as a resistance of 500 ohms, regardless of frequency; and 500 ohms is called the *characteristic or surge resistance* (or impedance) of the line. The line is said to be properly matched or terminated, and the whole of the power supplied by the generator is delivered to the load resistance at the far end.

In practice, as already stated, a certain percentage of the voltage or power are lost for every 100 feet of line, due to losses by heat or radiation on the way. Apart from this, however, the current and voltage would be exactly the same all along the line, and in fact the line would be a perfect link of any desired length between generator and load resistance, enabling them to be installed in different places without loss on the way. The load being purely resistive, in spite of the reactance of the line, current and voltage are everywhere in phase. There is no tendency for the line to resonate at any frequency, as line inductive and capacitive reactances cancel out.

232. Wave Reflection

If the load is reactive, or is not equal to the characteristic impedance of the line, matters are more complicated. Suppose that in the example considered a load of 2,000 ohms is connected, instead of 500 ohms. According to Ohm's Law it is impossible for 500 volts to be applied across 2,000 ohms and cause a current of 1 amp. to flow. Yet 1 amp. is arriving. What does it do? Part of it, having nowhere to go, starts back for home. In other words, it is *reflected* by the mismatch or improper termination. The reflected current, travelling in the opposite direction, can be regarded as being opposite in phase to that which is arriving, giving a resultant which is less than 1 amp. In this process a reflected voltage is produced which is *in* phase with that arriving, giving an augmented voltage at the end of the line. If 50 per cent. of the current is reflected, leaving 0·5 amp. to go into the load resistance, then the reflected voltage is also equal to half that arriving, giving in our example a total of 750. A voltage of 750 and current 0·5 amp. would fit a 1,500 ohm load, but not 2,000 ohms; so the reflected proportions have to be slightly higher—actually 60 per cent., giving 800 volts and 0·4 amp.

We now have 500 volts 1 amp. travelling from generator to load, and 300 volts 0·6 amp. returning, opposite in phase to one another, to the generator. At a distance back along the line equal to quarter of a wavelength the arriving wave and the returning wave are half a wavelength apart (because a return journey has to be made over the quarter wavelength distance). Whatever the voltage or current may be at any point on a wave, half a wavelength farther on it is at all times equal but opposite. So at a point quarter of a wavelength from the load end the reflected current is *in* phase with that arriving, while it is the voltages' turn to oppose one another. The voltage is therefore 200, and the current 1·6 amp. At a point one-eighth of a wavelength from the load and the arriving and reflected currents and voltages are all quarter of a wavelength, or 90°, out of phase, giving 582 volts and 1·16 amps. The same for any odd number of eighth-wavelengths. Half a wavelength along, the arriving and reflected waves are one whole wavelength apart, which brings them into step once more; so the conditions are the same as at the load.

233. Standing Waves

The current and voltages can be traced out point by point and when plotted are as in Fig. 214. It is important to realise that this is not, as it were, a flashlight photograph of the waves travelling along the line; these are R.M.S. values permanently set up at the points indicated, and would be indicated by meters connected in or across the lines at those points (assuming the meters did not appreciably affect the impedance of the line). Because the arriving and reflected waves, travelling in opposite directions, combine to cause this stationary wavelike distribution of current and voltage along the line, the effect is called *standing waves*. For comparison, the uniform distribution of current and voltage, resulting when the load resistance equals the surge impedance of the line, is shown dotted.

What happens when the reflected wave reaches the generator? It depends on the generator impedance. If it is 500 ohms, and resistive, then the 300 volt 0·6 amp. reflected wave is completely accepted by it, and if the line is an exact whole number of half-wavelengths long the voltage at the terminals of the generator is 500 (outgoing) + 300 (returning) or 800 volts, and the current 1 — 0·6, or 0·4 amp.; in fact, the conditions are the same as at the load end, and the line presents a resistance to the generator given by 800/0·4 = 2,000 ohms, the same as that of the load.

Suppose, however, that the generator resistance is something different, say 2,000 ohms; then part of the returning wave would be reflected back towards the load, and so on, the reflected quantities being smaller on each successive journey,

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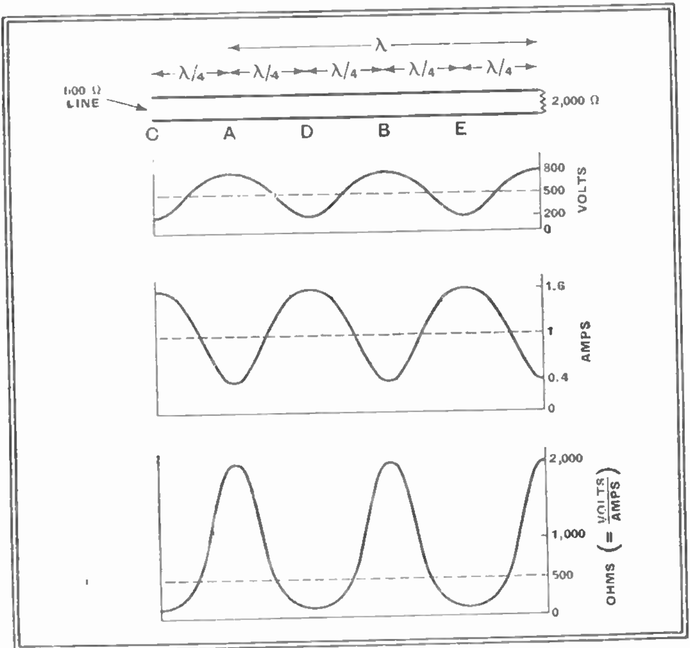

Fig. 214 : The load end of a 500-ohm line, with a 2,000-ohm load, showing the distribution of voltage, current, and (as a result), impedance

until finally becoming negligible. The standing waves are the resultant of all these travelling waves, and if a large proportion is reflected at each end—as happens if the impedances there are very much greater or less than the line impedance—it is possible for the voltages and currents at certain intervals along the line to build up to very large amounts. In the case of a transmitter feeder, the voltage may rise high enough to spark across. Even though the resistance of the line may be small enough to be negligible when properly matched, the loss due to heating and radiation is proportional to I^2, so may be substantial when standing waves are produced. Quite apart from this, however, the power delivered to a load by a generator—it may be an aerial fed by a valve oscillator—is a maximum when load and generator impedances are matched ; the same principle holds as with load and generator directly connected ; compare Fig. 68 (dotted curve). Standing waves are generally an indication that they are mismatched and, consequently, less than the maximum possible power is being delivered. In any case the line losses are greater than if standing waves are absent.

234. Load-to-Line Mismatch

In the example we have been following through, when the load resistance was equal to the line impedance—500 ohms—the power delivered was $500 \times 1 = 500$ watts. But with a 2,000-ohm load the voltage increased to 800 and the current dropped to 0·4 amp., giving 320 watts. The difference, 180 watts, is represented by the reflected wave, 300 volts 0·6 amp.

It has been stated that the loaded line presents a resistance of 2,000 ohms to the generator *if it is a whole number of half-wavelengths long*. But what if it isn't? To bring it within the scope of Fig. 214, suppose it is $1\frac{1}{4}$ wavelengths long. Then, assuming that the generator resistance is still 500 ohms and therefore the right amount for accepting the whole of the reflected wave on its first return, the voltage at the generator end (as seen from Fig. 214) is 200 and the current 1·6 amp.—again 320 watts. The impedance of the line to the generator is therefore $200/1 \cdot 6 = 125$ ohms.

By selecting this point at which to connect the generator, then, the effect of making the load 4 times the surge impedance of the line is to reduce the load on the generator to one-fourth of the surge impedance. The same applies at any point where the voltage and current have the values stated, that is to say at any odd number of quarter-wavelengths from the load. Although the power delivered by the generator is the same when connected to either a 2,000-ohm point or a 125-ohm point, the power wasted in the generator (I^2R) in the latter event is $1 \cdot 6^2 \times 500 = 1{,}280$ watts, but in the former only $0 \cdot 4^2 \times 500 = 80$ watts. When the load is matched to the line and generator, the power lost in the generator is the same as that delivered to the load—500 watts—so it *may* be considered advantageous to employ a deliberate mismatch in order to obtain an output of 320 watts with a loss of only 80 rather than get 500 and have to lose 500 in the process. In any case, it is obviously better to have too high a load resistance rather than too low (compare Secs. 99 and 118).

From the current and voltage curves it is easy to work out the impedance of a line of any length, and an impedance curve is shown in Fig. 214. It may be noted that at odd numbers of eighth-wavelengths the impedance is equal to the line surge impedance. From a consideration of the travelling currents and voltages in the line it can be shown that only at quarter-wavelength intervals are they exactly in phase, giving a resistive impedance ; at all other points they are reactive as well. To specify the impedance completely, therefore, it would be necessary to analyse it by giving curves of resistance and reactance.

One thing that emerges from all this is that it is possible to do a certain amount of impedance matching by selecting an appropriate point at which to connect. For example, suppose the generator resistance is 2,000 ohms ; then, by connecting it to a point on the mismatched 500-ohm line at which it also presents an impedance of 2,000 ohms (such as A or B), the generator is perfectly matched to the 2,000-ohm load, just as if it were connected straight to it. The only difference resulting from the fact that the surge impedance of the line does not match generator and load is that standing waves are set up on it, which slightly increase the power lost in the line itself. It must be remembered again that although we may be entitled to neglect the loss resistance of the line when calculating its surge impedance, it is usually enough to cause an appreciable loss of power. But if the generator is mismatched to the load it causes the power delivered to the load to be less, even if the loss due to line resistance is absolutely nil.

235. The Quarter-Wave Transformer

A more interesting result is that by shifting the point of connection to one at which the line impedance is 125 ohms (such as C, D, or E), a generator having an internal resistance of 125 ohms can be perfectly matched to a load of 2,000 ohms. The line behaves, in fact, as a 1 : 4 transformer. Similarly, points can be selected giving any ratio between 1 : 4 and 4 : 1, but those marked by letters A to E are preferable because they avoid complications due to reactance.

It is not necessary to use the whole of a long line as a matching transformer ; in fact, owing to enhanced losses it is generally undesirable to do so. It can be seen from Fig. 214 that the maximum ratio of transformation, combined with non-reactive impedance at both ends, is given by a section of line only quarter of a wavelength long. It can be seen that the mismatch ratio to the line itself is the same at both ends. In our example the ratio 125 ohms to 500 ohms (generator line) is equal to 500 ohms to 2,000 ohms (line to load), each being 1 : 4, and, incidentally, equal to the voltage ratio of the whole transformer. If Z_G is the generator impedance, Z_0 the surge impedance of the

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line and Z_L the load impedance, Z_G is to Z_O as Z_O is to Z_L; put otherwise, $\frac{Z_G}{Z_O} = \frac{Z_O}{Z_L}$, so $Z_O^2 = Z_G Z_L$, and $Z_O = \sqrt{Z_G Z_L}$. This formula enables us to work out the surge resistance of the quarter-wave line necessary for matching two unequal impedances. It is often used for linking lines of unequal impedance without causing reflections. In practice it is limited to the fairly narrow ranges of impedance over which it is practicable to construct lines.

That brings us to the calculation of Z_o, the surge impedance. Neglecting the loss resistance of the line, it is equal to $\sqrt{\frac{l}{c}}$, where l and c are the inductance and capacitance per centimetre of the line; and formulæ have been worked out for these in terms of the spacing and diameter of the wires or tubes used. When they are substituted in $Z_o = \sqrt{\frac{l}{c}}$ the result for a parallel

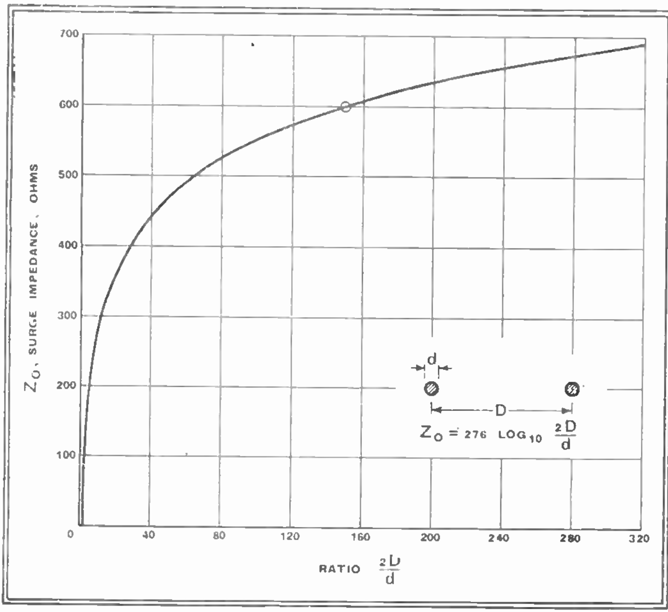

Fig. 215: Curve giving the surge impedance of a parallel wire line in terms of diameter and spacing of the wires

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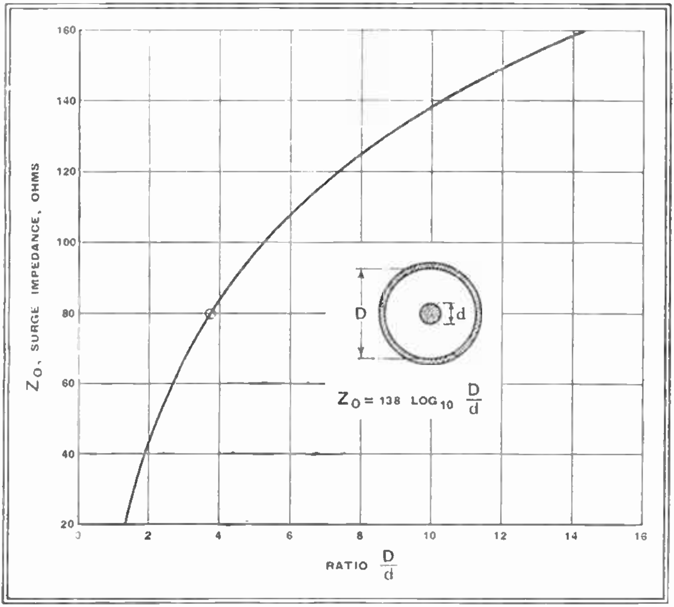

Fig. 216 : Curve similar to Fig. 215, but for coaxial lines

wire line (Fig. 215) is 276 $\log_{10} \frac{2D}{d}$, and for a coaxial line (Fig. 216) it is 138 $\log_{10} \frac{D}{d}$. For reasonable practical values of $\frac{D}{d}$, the surge impedance of a parallel wire line may be 300 to 650 ohms and a coaxial line 60 to 100 ohms. The most efficient proportions give *Zo* equal to 600 and 80 respectively. The accompanying curves enable the surge impedance of either type of line to be found from its dimensions, provided that it is air-spaced.

An example of the use of a quarter-wave transformer would occur if it were wished to connect a centre-fed dipole—say 80 ohms—to a 320-ohm parallel-wire feeder. These could be matched by joining them up by a section of line a quarter wavelength long and spaced to give a *Zo* equal to $\sqrt{80 \times 320} =$ 160 ohms. A parallel wire line would have to be excessively

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close to give this, but the problem could be solved by using a length of 80-ohm coaxial cable for each limb and joining the metal sheaths together as in Fig. 217, putting the impedances in series across the ends of the 320-ohm line.

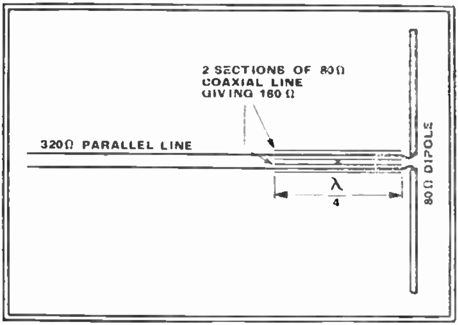

Fig. 217 : One method of matching a dipole to a line of higher impedance

236. Fully Resonant Lines

Going now to extremes of mismatch, it is of interest to inquire what happens when the " load " resistance is either infinite or zero ; in other words, when the end of the line is open-circuited or short-circuited. Take the open circuit first. If this were done to our original 500-ohm example (Fig. 214), the current at the end would obviously be nil, and the voltage would rise to 1,000—double its amount across the matched load. The same would apply at points A and B ; while at C, D, and E the voltage would be nil and the current 2 amps. The impedance curve would fluctuate between zero and infinity.

With a short-circuited line, there could be no volts across the end, but the current would be 2 amps. ; in fact, exactly as at E with the open line. A shorted line, then, is the same as an open line shifted quarter of a wavelength along. Reflection in both cases is complete, because there is no load resistance to absorb any of the power.

If the generator resistance is very large or very small, *nearly* all the reflected wave will itself be reflected back, and so on, so that if the line is of such a length that the voltage and current maximum points coincide with every reflection, the voltages and currents will build up to high values at those maximum points—dangerously high with a powerful transmitter. This

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reminds us of the behaviour of a high-Q resonant circuit (Sec. 51). The maximum current or voltage points are called *antinodes*, and the points where current or voltage are zero are *nodes*.

When the length of a short-circuited or open-circuited line is a whole number of quarter-wavelengths, the input impedance is approximately zero or infinity. An *odd* number of quarter wavelengths gives opposites at the ends—infinite resistance if the other end is shorted, and vice versa. An *even* number of quarter wavelengths gives the same at each end.

In between, as there is now no load resistance, the impedance

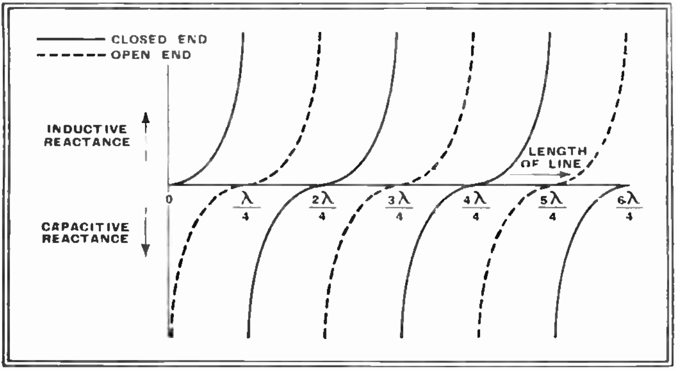

Fig. 218 : Showing how the reactance of a line varies with its length

is a pure reactance. At each side of a node or antinode there are opposite reactances—inductive and capacitive. If it is a current node, the reactance at a short distance each side is very large; if a voltage node, very low. Fig. 218 shows how it varies. It is clear from this that a short length of line—less than quarter of a wavelength—can be used to provide any value of inductance or capacitance. For very short wavelengths, this form is generally more convenient than the usual coil or condenser.

237. Stub Matching

One purpose for which this form of reactance is very useful is as an alternative to the quarter-wave transformer method of matching lines.

The quarter-wave transformer generally involves inserting a specially constructed section, with appropriate connecting and mounting arrangements. Going back to Fig. 214, the 500-ohm line terminated by a load of 2,000 ohms could be matched by inserting between them a quarter-wave section of $\sqrt{500 \times 2,000} = 1,000$ ohms. As Figs. 215 and 216 show, this is undesirable. It would have to be made of wires so thin that they would cause resistance loss, or so far apart that they would cause radiation loss.

But looking at the lowest curve of Fig. 214, we see that there is a point on the line, one-eighth of a wavelength from the load, at which the impedance is 500 ohms. The reason why this does not match the 500-ohm line is that it is

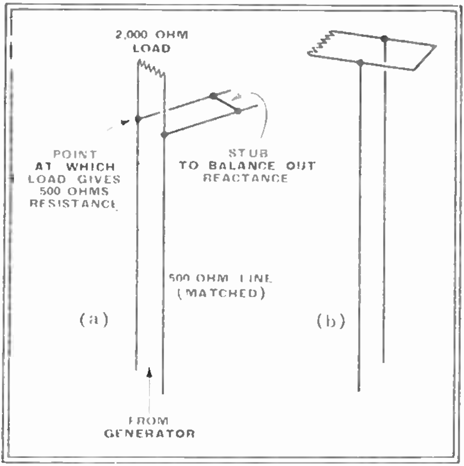

Fig. 219: Example of stub-matching a load—which might consist of an aerial system—to a line of unequal surge impedance. The reactance of the stub is adjusted by moving the shorting bar along it, and the resistance match is adjusted by moving the stub along the feeder line
***b* is electrically the same as a, the difference being mechanical**

only at quarter-wave intervals that the phase relationships between the original and reflected waves are such as to give a purely resistive impedance. Elsewhere there is reactance, which upsets the matching.

Any reactance can be balanced out or neutralised by an equal reactance of the opposite kind. So all we have to do with the system in Fig. 214 is to find, by calculation or experiment, the point nearest the load at which the *resistance* matches the line—500 ohms in this case—and then balance out the reactance by connecting in parallel at this point a short length of open or shorted line, called a stub. This can easily be made on the spot, out of the same wire as the rest of the line.

The load in Fig. 214 is higher than the line impedance, so the reactance in the quarter-wave section nearest the load is of

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the same kind as for an open-circuited line ; that is (from Fig. 218), capacitive.

So the stub must be inductive, and Fig. 218 shows that if it is to be less than a quarter-wave long it must be short-circuited at the end.

Fig. 219 shows the arrangement. There are two variables—position of stub (i.e., position on the feeder at which the load at the end " looks like " a resistance equal to the feeder's Z_0, together with a reactance in parallel), and length of stub, which must be adjusted to balance out the line reactance.

Insulators used for supporting or spacing the line inevitably introduce a certain amount of capacitance in parallel with it, thereby unmatching it. As Fig. 218 shows, a capacitive reactance at one point on a line is transformed to an equal inductive reactance farther along. If, therefore, this point is selected for the next insulator it neutralises the effect of the first, and prevents the mismatching growing from insulator to insulator.

238. " Metallic Insulators "

Fig 218 also shows that a short-circuited quarter-wave line

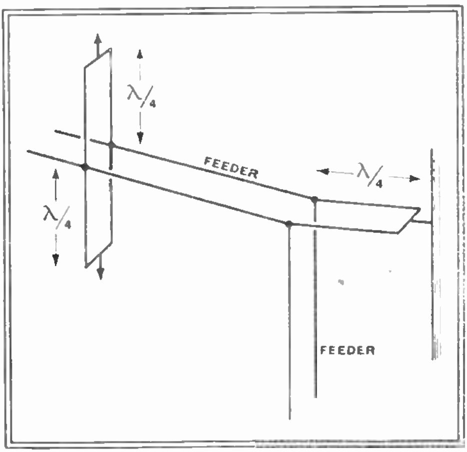

Fig. 220 : Showing how " metallic insulators " can be used to support and direct a feeder line

(shaped like a hairpin) presents at the open end an infinite impedance, or rather would do so it if had no resistance loss

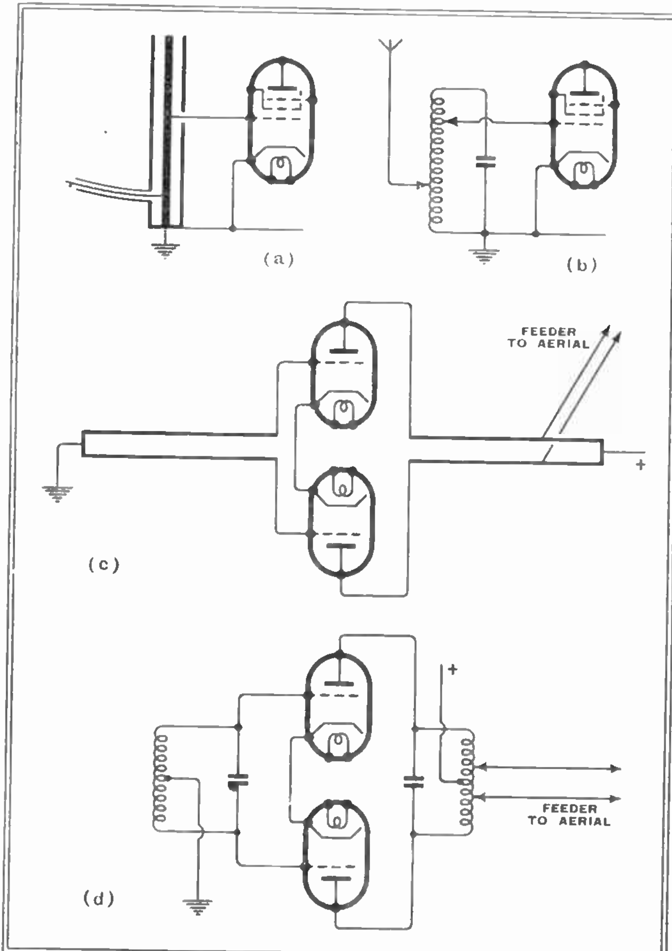

Fig. 221 : Two circuits in which distributed or "line" tuned circuits are substituted for the well-known concentrated type. *a* and *b* are unbalanced, and *c* and *d* balanced or symmetrical

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(compare dynamic resistance again). If made of the same wire as a transmission line, the impedance is sure to be higher at very high frequencies than that of an insulator, and it is cheaper, stronger and easier to instal. So parallel wire lines are generally anchored by these " metallic insulators ", as shown in Fig. 220. They are usually earthed at the centre.

Metallic insulators differ from the conventional kind in acting as insulators only at the particular frequency for which they are quarter of a wavelength—or any odd number of quarter waves—long. So they act also as filters, removing waves of other lengths or frequencies. Metallic insulators cannot be used in systems that have to function at more than one fixed frequency.

239. Lines as Tuned Circuits

A quarter-wave line shorted at one end, such as can be used as a metallic insulator is, as we have seen, one in which the inductive and capacitive reactances are equal. It is equivalent, in fact, to a resonant tuned circuit ; and at wavelengths less than about 2 metres is generally more efficient and easily constructed than the conventional sort.

The lower the resistance of the wire, the higher the dynamic resistance across the open ends ; and to match lower impedances all that is necessary is to tap it down, just as if it were any other sort of tuned circuit.

The parallel wire or bar type lends itself to push-pull connection, and the coaxial type to single-ended circuits. Owing to their high efficiency, it is easier to obtain oscillation than with coils and condensers. Fig. 221*a* is an example of a coaxial tuning circuit, such as might be used in a receiver, and *b* is the conventional equivalent. A coaxial aerial feeder is used, and being normally about 80 ohms, is tapped low down, near the earthed end. The impedance at the top end is normally many thousands of ohms, and may be too high for the input of a valve, which is quite low at very high frequencies.

Fig. 221*c* is an example of a push-pull oscillator, with its equivalent at *d*.

These are some of the increasingly common applications of transmission lines.

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APPENDIX

The reader of various books and articles on radio is very liable to be confused by different terms used to mean the same thing. The following list has therefore been compiled to help clear up the matter. In most cases the first to be mentioned is the one most commonly used in this book. The associated terms are not necessarily *exact* equivalents. Terms distinctively American are printed in italics.

Radio—Wireless

Radar—Radiolocation

Radio frequency (R.F.)—High frequency (H.F.)

Audio frequency (A.F.)—Low frequency (L.F.)—Speech frequency—Voice frequency

Intermediate frequency (I.F.)—Supersonic frequency

Ultra-high frequency (U.H.F.)—Very high frequency (V.H.F.)

Harmonic—Overtone

Frequency—Periodicity

Root-mean-square (R.M.S.)—Effective—Virtual

Capacitance—Capacity

Permittivity—Dielectric constant—Specific inductive capacity

Q—Magnification

Dynamic resistance—*Antiresonant impedance*

Surge impedance—Characteristic impedance

Feeder—Transmission line

Aerial—*Antenna*

Earth—*Ground*

Frame aerial—*Loop antenna*

Valve—*Vacuum tube*—*Tube*—*Audion*

Anode—*Plate*

Anode A.C. resistance—Anode resistance—Valve impedance—*Plate impedance*

Anode battery—H.T.—" *B* " *battery*

Filament battery—L.T.—" *A* " *battery*

Grid battery—Bias battery—G.B.—" *C* " *battery*

Mutual conductance—*Transconductance*

High-vacuum—Hard

Gas-filled—Soft—Low-vacuum

Tetrode—Screen-grid valve (but not all tetrodes are screen-grid valves)

Heptode—Pentagrid

Frequency Changer—Mixer

Auto-bias—Self-bias

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Detection—Rectification—*Demodulation* (in U.S.A.: British usage reserves this term for a different phenomenon)

Reaction—Retroaction—Feedback—*Regeneration*

Quality (of reproduced sound)—*Fidelity*

Automatic gain control (A.G.C.)—Automatic volume control (A.V.C.)

Frequency changer—First detector

Reaction coil—*Tickler*

Tuned circuit—*Tank circuit*—LC circuit

Condenser—Capacitor

Coil—Inductor

Screen—*Shield*

Moving coil loud speaker—*Dynamic loud speaker*

Telephones—Phones—Earphones—Headphones—Headset

Radiogramophone—*Combination*

Interference—Jamming

Atmospherics—Strays—X's—*Static*

Noise—Machine interference—*Man-made static*

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Reference should also be made to the list of Contents on page ix, and alternative terms on pages 350 & 351

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