# The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR

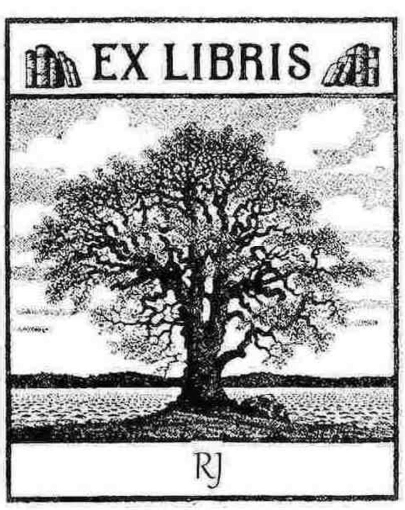
RADIO COMMUNICATION

$2.00

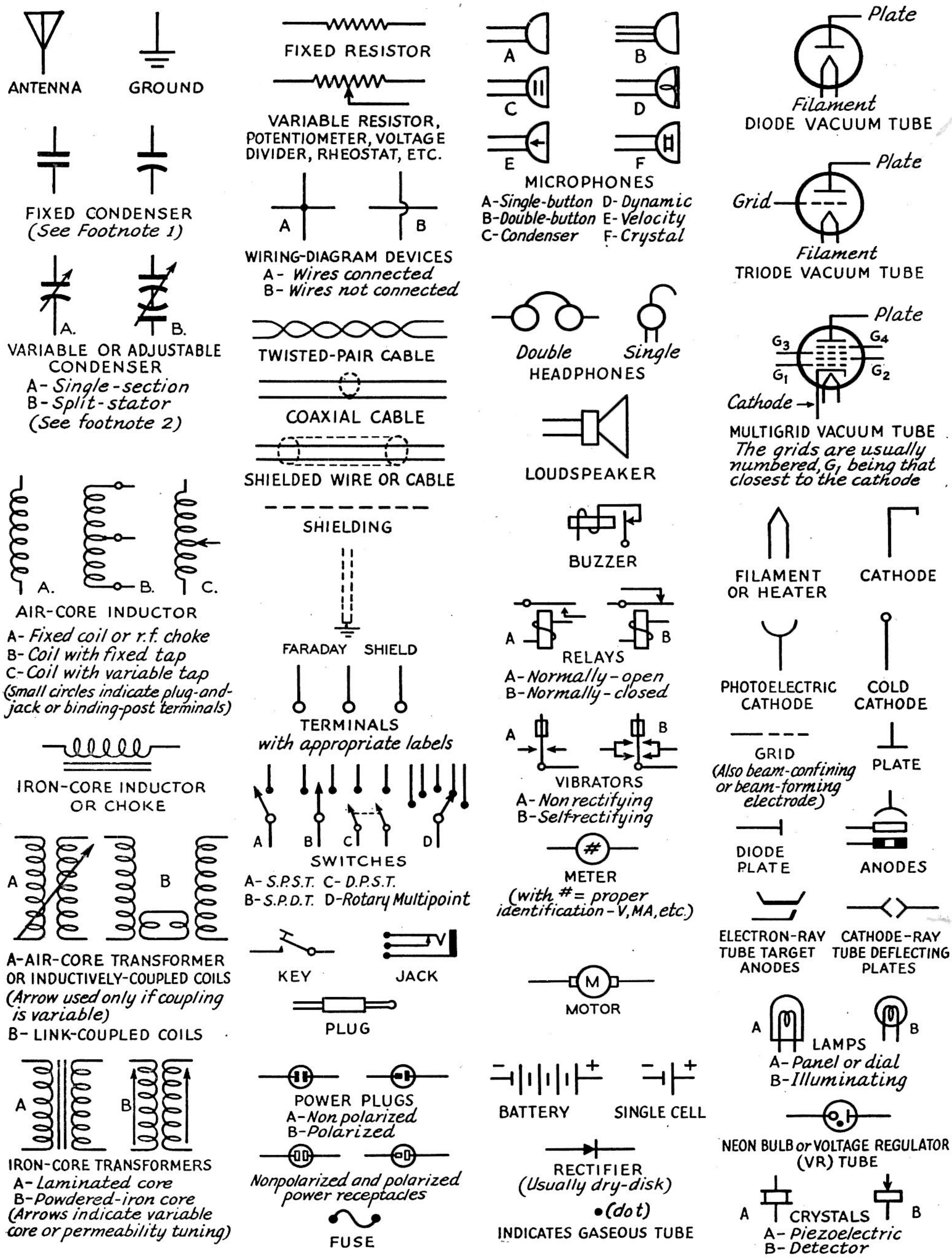
PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

25th edition 1948

# THE RADIO AMATEUR'S HANDBOOK



# SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS




[1] Where it is necessary or desirable to identify the electrodes, the curved element represents the *outside* electrode (marked "outside foil," "ground," etc.) in fixed paper- and ceramic-dielectric condensers, and the *negative* electrode in electrolytic condensers.

[2] In the modern symbol, the curved line indicates the moving element (rotor plates) in variable and adjustable air- or mica-dielectric condensers.

In the case of switches, jacks, relays, etc., only the basic combinations are shown. Any combination of these symbols may be assembled as required, following the elementary forms shown.

TWENTY-FIFTH EDITION

1948

# THE RADIO AMATEUR'S HANDBOOK

*by the*

HEADQUARTERS STAFF OF THE AMERICAN RADIO RELAY LEAGUE

*Published by*

THE AMERICAN RADIO RELAY LEAGUE, INC.
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# Foreword

In twenty-two years of continuous publication *The Radio Amateur's Handbook* has become as much of an institution as amateur radio itself. Produced by the amateur's own organization, the American Radio Relay League, and written with the needs of the practical amateur constantly in mind, it has earned universal acceptance not only by amateurs but by all segments of the technical radio world, from students to engineers, servicemen to operators. This wide dependence on the *Handbook* is founded on its practical utility, its treatment of radio communication problems in terms of how-to-do-it rather than by abstract discussion and abstruse formulas.

But there is another factor as well: dealing with a fast-moving and progressive science, sweeping and virtually continuous modification has been a feature of the *Handbook* — always with the objective of presenting the soundest and best aspects of current practice rather than the merely new and novel. Its annual rewriting is a major task of the headquarters group of the League, participated in by skilled and experienced amateurs well acquainted with the practical problems in the art.

In contrast to most publications of a comparable nature, the *Handbook* is printed in the format of the League's monthly magazine, *QST*. This, together with extensive and usefully-appropriate catalog advertising by manufacturers producing equipment for the radio amateur, makes it possible to distribute for a very modest charge a work which in volume of subject matter and profusion of illustration surpasses most available radio texts selling for several times its price.

This twenty-fifth edition of the *Handbook* is featured by the complete rewriting of most of the material. The textbook style that made the book so useful for training purposes during the war years has been replaced by a simpler and more understandable discussion of those basic facts that an amateur should know to get the most out of designing and using his apparatus. Owners of previous editions will recognize immediately that the over-all plan of the book has been changed — achieving, we believe, the object of segregating the material so that it can be most conveniently used. A great deal of new equipment has been constructed especially for this edition. As always, the object has been to show the best of current technique through equipment designs proved by thorough testing. As the art grows, the problem of presenting a representative selection of gear grows with it — a state of affairs that is reflected in an increase of well over a hundred pages in this edition. New chapters on ultrahigh frequencies, station assembly, and the elimination of interference to broadcasting have been added to round out the treatment of all phases of amateur radio. The material on operating has likewise been greatly expanded. Altogether, this revision is the most comprehensive of recent years.

The *Handbook* has long been considered an indispensable part of the amateur's equipment. We earnestly hope that the present edition will succeed in bringing as much assistance and inspiration to amateurs and would-be amateurs as have its predecessors.

Kenneth B. Warner
*Managing Secretary, A.R.R.L.*

West Hartford, Conn.
December, 1947

# Table of Contents

# THE AMATEUR'S CODE

**ONE** *The Amateur is Gentlemanly* . . . He never knowingly uses the air for his own amusement in such a way as to lessen the pleasure of others. He abides by the pledges given by the ARRL in his behalf to the public and the Government.

**TWO** *The Amateur is Loyal* . . . He owes his amateur radio to the American Radio Relay League, and he offers it his unswerving loyalty.

**THREE** *The Amateur is Progressive* . . . He keeps his station abreast of science. It is built well and efficiently. His operating practice is clean and regular.

**FOUR** *The Amateur is Friendly* . . . Slow and patient sending when requested, friendly advice and counsel to the beginner, kindly assistance and coöperation for the broadcast listener; these are marks of the amateur spirit.

**FIVE** *The Amateur is Balanced* . . . Radio is his hobby. He never allows it to interfere with any of the duties he owes to his home, his job, his school, or his community.

**SIX** *The Amateur is Patriotic* . . . His knowledge and his station are always ready for the service of his country and his community.

# Amateur Radio

Amateur radio is a scientific hobby, a means of gaining personal skill in the fascinating art of electronics and an opportunity to communicate with fellow citizens by private short-wave radio. Scattered over the globe are more than 100,000 amateur radio operators who perform a service defined in international law as one of "self training, intercommunication and technical investigations carried on by . . . duly authorized persons interested in radio technique solely with a personal aim and without pecuniary interest."

From a humble beginning at the turn of the century, amateur radio has grown to become an established institution. Today the American followers of amateur radio number 80,000, a group of trained communicators from whose ranks will come the professional communications specialists and executives of tomorrow — just as many of today's radio leaders were first attracted to radio by their early interest in amateur radio communication. A powerful and prosperous organization now provides a bond between amateurs and protects their interests; an internationally-respected magazine is published solely for their benefit. The Army and Navy seek the coöperation of the amateur in developing communications reserves. Amateur radio supports a manufacturing industry which, by the very demands of amateurs for the latest and best equipment, is always up-to-date in its designs and production techniques — in itself a national asset. Amateurs have won the gratitude of the nation for their heroic performances in times of natural disaster. Through their organization, amateurs have coöperative working agreements with such agencies as the United Nations and the Red Cross. Amateur radio is, indeed, a magnificently useful institution.

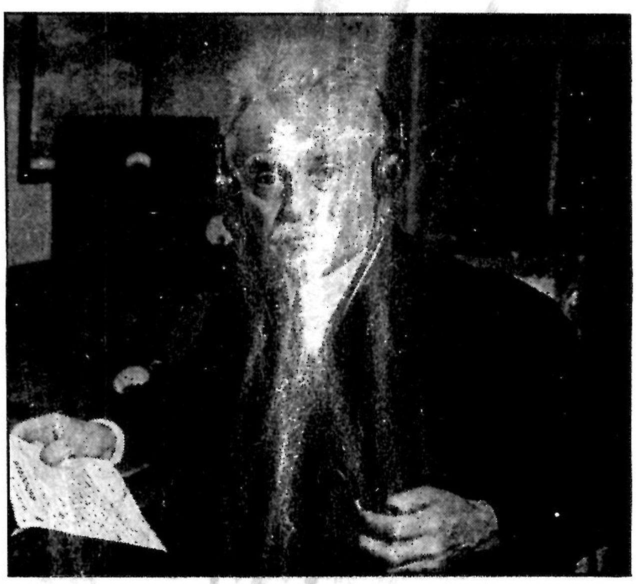
Although as old as the art of radio itself, amateur radio did not always enjoy such prestige. Its first enthusiasts were private citizens of an experimental turn of mind whose imaginations went wild when Marconi first proved that messages actually could be sent by wireless. They set about learning enough about the new scientific marvel to build home-made stations. By 1912 there were numerous Government and commercial stations, and hundreds of amateurs; regulation was needed, so laws, licenses and wavelength specifications for the various services appeared. There was then no amateur organization nor spokesman. The official viewpoint toward amateurs was something like this:

"Amateurs? . . . Oh, yes. . . . Well, stick 'em on 200 meters and below; they'll never get out of their backyards with that."

But as the years rolled on, amateurs found out how, and DX (distance) jumped from local to 500-mile and even occasional 1,000-mile two-way contacts. Because all long-distance messages had to be relayed, relaying developed into a fine art — an ability that was to prove invaluable when the Government suddenly called hundreds of skilled amateurs into war service in 1917. Meanwhile U. S. amateurs began to wonder if there were amateurs in other countries across the seas and if, some day, we might not span the Atlantic on 200 meters.

Most important of all, this period witnessed the birth of the American Radio Relay League, the amateur radio organization whose name was to be virtually synonymous with subsequent amateur progress and short-wave development. Conceived and formed by the famous inventor, the late Hiram Percy Maxim, ARRL was formally launched in early 1914. It had just begun to exert its full force in amateur activities when the United States declared war in 1917, and by that act sounded the knell for amateur radio for the next two and a half years. There were then over 6000 amateurs. Over 4000 of them served in the armed forces during that war.

Today, few amateurs realize that World



HIRAM PERCY MAXIM
President ARRL, 1914–1936

War I not only marked the close of the first phase of amateur development but came very near marking its end for all time. The fate of amateur radio was in the balance in the days immediately following the signing of the Armistice. The Government, having had a taste of supreme authority over communications in wartime, was more than half inclined to keep it. The war had not been ended a month before Congress was considering legislation that would have made it impossible for the amateur radio of old ever to be resumed. ARRL's President Maxim rushed to Washington, pleaded, argued, and the bill was defeated. But there was still no amateur radio; the war ban continued. Repeated representations to Washington met only with silence. The League's offices had been closed for a year and a half, its records stored away. Most of the former amateurs had gone into service; many of them would never come back. Would those returning be interested in such things as amateur radio? Mr. Maxim, determined to find out, called a meeting of the old Board of Directors. The situation was discouraging: amateur radio still banned by law, former members scattered, no organization, no membership, no funds. But those few determined men financed the publication of a notice to all the former amateurs that could be located, hired Kenneth B. Warner as the League's first paid secretary, floated a bond issue among old League members to obtain money for immediate running expenses, bought the magazine *QST* to be the League's official organ, started activities, and dunned officialdom until the wartime ban was lifted and amateur radio resumed again, on October 1, 1919. There was a headlong rush by amateurs to get back on the air. Gangway for King Spark! Manufacturers were hard put to supply radio apparatus fast enough. Each night saw additional dozens of stations crashing out over the air. Interference? It was bedlam!

But it was an era of progress. Wartime needs had stimulated technical development. Vacuum tubes were being used both for receiving and transmitting. Amateurs immediately adapted the new gear to 200-meter work. Ranges promptly increased and it became possible to bridge the continent with but one intermediate relay.

## ● TRANSATLANTICS

As DX became 1000, then 1500 and then 2000 miles, amateurs began to dream of trans-Atlantic work. Could they get across? In December, 1921, ARRL sent abroad an expert amateur, Paul F. Godley, 2ZE, with the best receiving equipment available. Tests were run, and *thirty* American stations were heard in Europe. In 1922 another trans-Atlantic test was carried out and 315 American calls were logged by European amateurs and one French and two British stations were heard on this side.

Everything now was centered on one objective: two-way amateur communication across the Atlantic! It must be possible — but somehow it couldn't quite be done. More power? Many already were using the legal maximum. Better receivers? They had superheterodynes. Another wavelength? What about those undisturbed wavelengths *below* 200 meters? The engineering world thought they were worthless — but they had said that about 200 meters. So, in 1922, tests between Hartford and Boston were made on 130 meters with encouraging results. Early in 1923, ARRL-sponsored tests on wavelengths down to 90 meters were successful. Reports indicated that *as the wavelength dropped the results were better.* A growing excitement began to spread through amateur ranks.

Finally, in November, 1923, after some months of careful preparation, two-way amateur trans-Atlantic communication was accomplished, when Schnell, 1MO, and Reinartz, 1XAM (now W9UZ and W3RB, respectively) worked, for several hours with Deloy, 8AB, in France, with all three stations on 110 meters! Additional stations dropped down to 100 meters and found that they, too, could easily work two-way across the Atlantic. The exodus from the 200-meter region had started. The "short-wave" era had begun!

By 1924 dozens of commercial companies had rushed stations into the 100-meter region. Chaos threatened, until the first of a series of national and international radio conferences partitioned off various bands of frequencies for the different services. Although thought still centered around 100 meters, League officials at the first of these frequency-determining conferences, in 1924, wisely obtained amateur bands not only at 80 meters but at 40, 20, 10 and even 5 meters.

Eighty meters proved so successful that "forty" was given a try, and QSOs with Australia, New Zealand and South Africa soon became commonplace. Then how about 20 meters? This new band revealed entirely unexpected possibilities when 1XAM worked 6TS on the West Coast, direct, at high noon. The dream of amateur radio — daylight DX! — was finally true.

From then until "Pearl Harbor," when U. S. amateurs were again closed down "for the duration," amateur radio thrilled with a series of unparalleled accomplishments. Countries all over the world came on the air, and the world total of amateurs passed the 100,000 mark. . . . ARRL representatives deliberated with the representatives of twenty-two other nations in Paris in 1925 where, on April 17th, the International Amateur Radio Union was formed — a federation of national amateur radio societies. . . . The League began issuing certificates to those who could prove they had worked all six continents. More than six thousand amateurs have been awarded WAC certificates.

## PUBLIC SERVICE

Amateur radio is a grand and glorious hobby but this fact alone would hardly merit such wholehearted support as is given it by our Government at international conferences. There are other reasons. One of these is a thorough appreciation by the Army and Navy of the value of the amateur as a source of skilled radio personnel in time of war. Another asset is best described as "public service."

About 4000 amateurs had contributed their skill and ability in '17–'18. After the war it was only natural that cordial relations should prevail between the Army and Navy and the amateur. These relations strengthened in the next few years and, in gradual steps, grew into coöperative activities which resulted, in 1925, in the establishment of the Naval Communications Reserve and the Army-Amateur Radio System (now the Military Amateur Radio System). In World War II thousands of amateurs in the Naval Reserve were called to active duty, where they served with distinction, while many other thousands served in the Army, Air Forces, Coast Guard and Marine Corps. Altogether, more than 25,000 radio amateurs served in the armed forces of the United States. Other thousands were engaged in vital civilian electronic research, development and manufacturing. They also organized and manned the War Emergency Radio Service, the communications section of OCD.

The "public-service" record of the amateur is a brilliant tribute to his work. These activities can be roughly divided into two classes, expeditions and emergencies. Amateur coöperation with expeditions began in 1923 when a League member, Don Mix, 1TS, of Bristol, Conn. (now assistant technical editor of *QST*), accompanied MacMillan to the Arctic on the schooner *Bowdoin* with an amateur station. Amateurs in Canada and the U.S. provided the home contacts. The success of this venture was such that other explorers followed suit. During subsequent years a total of perhaps two hundred voyages and expeditions were assisted by amateur radio, and for many years no expedition has taken the field without such plans.

Since 1913 amateur radio has been the principal, and in many cases the only, means of outside communication in several hundred storm, flood and earthquake emergencies in this country. The 1936 eastern states flood, the 1937 Ohio River Valley flood, the Southern California flood and Long Island-New England hurricane disaster in 1938, and the Florida-Gulf Coast hurricanes of 1947 called for the amateur's greatest emergency effort. In these disasters and many others — tornadoes, sleet storms, forest fires, blizzards — amateurs played a major rôle in the relief work and earned wide commendation for their resourcefulness in effecting communication where all other means had failed. During 1938 ARRL inaugurated a new emergency-preparedness program, registering personnel and equipment in its Emergency Corps and putting into effect a comprehensive program of coöperation with the Red Cross, and in 1947 a National Emergency Coördinator was appointed to full-time duty at League headquarters.

## TECHNICAL DEVELOPMENTS

Throughout these many years the amateur was careful not to slight experimental development in the enthusiasm incident to international DX. The experimenter was constantly at work on ever-higher frequencies, devising improved apparatus, and learning how to cram several stations where previously there was room for only one! In particular, the amateur pressed on to the development of the very high frequencies and his experience with five meters is especially representative of his initiative and resourcefulness and his ability to make the most of what is at hand. In 1924, first amateur experiments in the vicinity of 56 Mc. indicated that band to be practically worthless for DX. Nonetheless, great "short-haul" activity eventually came about in the band and new gear was developed to meet its special problems. Beginning in 1934 a series of investigations by the brilliant experimenter, Ross Hull (later *QST*'s editor), developed the theory of v.h.f. wave-bending in the lower atmosphere and led amateurs to the attainment of better distances; while occasional manifestations of ionospheric propagation, with still greater distances, gave the band uniquely-erratic performance. By Pearl Harbor thousands of amateurs were spending much of their time on this and the next higher band, many having worked hundreds of stations at distances up to several thousand miles. Transcontinental 6-meter DX is now a commonplace occurrence; even the oceans have been bridged! It is a tribute to these indefatigable amateurs that today's concept of v.h.f. propagation was developed largely through amateur research.

The amateur is constantly in the forefront of technical progress. His incessant curiosity, his eagerness to try anything new, are two reasons. Another is that ever-growing amateur radio continually overcrowds its frequency assignments, spurring amateurs to the development and adoption of new techniques to permit the

A corner of the ARRL laboratory.

accommodation of more stations. For examples, amateurs turned from spark to c.w., designed more selective receivers, adopted crystal control and pure d.c. power supplies. From the ARRL's own laboratory in 1932 came James Lamb's "single-signal" superheterodyne — the world's most advanced high-frequency radiotelegraph receiver — and, in 1936, the "noise-silencer" circuit. Amateurs are now turning to speech "clippers" to reduce bandwidths of 'phone transmissions and investigating "single-sideband suppressed-carrier" systems which promise to halve the spectrum space required by a voice-modulated signal.

During the recent war, thousands of skilled amateurs contributed their knowledge to the development of secret radio devices, both in Government and private laboratories. Equally as important, the prewar technical progress by amateurs provided the keystone for the development of modern military communications equipment. Perhaps more important today than individual contributions to the art is the mass coöperation of the amateur body in Government projects such as propagation studies; each participating amateur station is in reality a separate field laboratory from which reports are made for correlation and analysis.

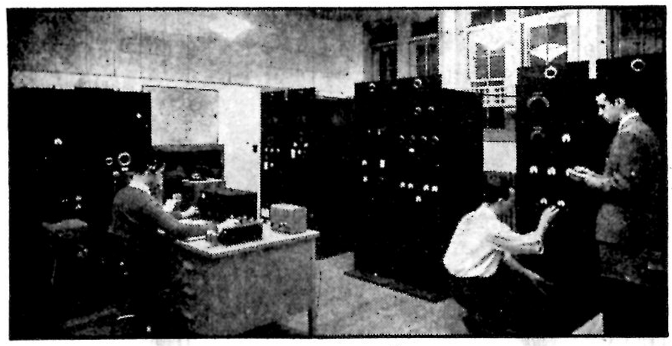
Emergency relief, expedition contact, experimental work and countless instances of other forms of public service — rendered, as they always have been and always will be, without hope or expectation of material reward — made amateur radio an integral part of our peacetime national life. The importance of amateur participation in the armed forces and in other aspects of national defense have emphasized more strongly than ever that amateur radio is vital to our national existence.

## THE AMERICAN RADIO RELAY LEAGUE

The ARRL is today not only the spokesman for amateur radio in this country but it is the largest amateur organization in the world. It is strictly of, by and for amateurs, is noncommercial and has no stockholders. The members of the League are the owners of the ARRL and *QST*.

The League is organized to represent the amateur in legislative matters. It is pledged to promote interest in two-way amateur communication and experimentation. It is interested in the relaying of messages by amateur radio. It is concerned with the advancement of the radio art. It stands for the maintenance of fraternalism and a high standard of conduct.

One of the League's principal purposes is to keep amateur activities so well conducted that the amateur will continue to justify his existence. Amateur radio offers its followers countless pleasures and unending satisfaction. It also calls for the shouldering of responsibilities — the maintenance of high standards, a coöperative loyalty to the traditions of amateur radio, a dedication to its ideals and principles, so that the institution of amateur radio may continue to operate "in the public interest, convenience and necessity."



The operating room at W1AW.

The operating territory of ARRL is divided into fifteen U. S. and five Canadian divisions. The affairs of the League are managed by a Board of Directors. One director is elected every two years by the membership of each U. S. division, and a Canadian General Manager is elected every two years by the Canadian membership. These directors then choose the president and vice-president, who are also members of the Board. The managing secretary, treasurer and communications manager are appointed by the Board. The directors, as representatives of the amateurs in their divisions, meet annually to examine current amateur problems and formulate ARRL policies thereon.

ARRL owns and publishes the monthly magazine, *QST*. Acting as a bulletin of the League's organized activities, *QST* also serves as a medium for the exchange of ideas and fosters amateur spirit. Its technical articles are renowned. It has grown to be the "amateur's bible," as well as one of the foremost radio magazines in the world. Membership dues include a subscription to *QST*.

ARRL maintains a model headquarters amateur station, known as the Hiram Percy Maxim Memorial Station, in Newington, Conn. Its call is W1AW, the call held by Mr. Maxim until his death and later transferred to the League station by a special FCC action. Separate transmitters of maximum legal power on each amateur band have permitted the station to be heard regularly all over the world. More important, W1AW transmits on regular schedules bulletins of general interest to amateurs, conducts code practice as a training feature, and engages in two-way work on all popular bands with as many amateurs as time permits.

At the headquarters of the League in West Hartford, Conn., is a well-equipped laboratory to assist staff members in preparation of technical material for *QST* and the *Radio Amateur's Handbook*. Among its other activities the League maintains a Communications Department concerned with the operating activities of League members. A large field organization is headed by a Section Communications Manager in each of the country's seventy-one sections. There are appointments for qualified members as Official Relay Station or Official 'Phone Station for traffic handling; as Official Observer for monitoring frequencies and the quality of signals; as Route Manager and 'Phone Activities Manager for the establishment of trunk lines and networks; as Emergency Coördinator for the promotion of amateur preparedness to cope with natural disasters; and as Official Experimental Station for those pioneering the frequencies above 50 Mc. Mimeographed bulletins keep appointees informed of the latest developments. Special activities and contests promote operating skill. A special section is reserved each month in *QST* for amateur news from every section of the country.

## AMATEUR LICENSING IN THE UNITED STATES

The Communications Act lodges in the Federal Communications Commission authority to classify and license radio stations and to prescribe regulations for their operation. Pursuant to the law, FCC has issued detailed regulations for the amateur service.

A radio amateur is a duly authorized person interested in radio technique solely with a personal aim and without pecuniary interest. Amateur operator licenses are given to U. S. citizens who pass an examination on operation and apparatus and on the provisions of law and regulations affecting amateurs, and who demonstrate ability to send and receive code at 13 words per minute. Station licenses are granted only to licensed operators and permit communication between such stations for amateur purposes, i.e., for personal noncommercial aims flowing from an interest in radio technique. An amateur station may not be used for material compensation of any sort nor for broadcasting. Narrow bands of frequencies are allocated exclusively for use by amateur stations. Transmissions may be on any frequency within the assigned bands. All the frequencies may be used for c.w. telegraphy and some are available for radiotelephony by any amateur, while others are reserved for radiotelephone use by persons having at least a year's experience and who pass the examination for a Class A license. The input to the final stage of amateur stations is limited to 1000 watts and on frequencies below 60 Mc. must be adequately-filtered direct current. Emissions must be free from spurious radiations. The licensee must provide for measurement of the transmitter frequency and establish a procedure for checking it regularly. A complete log of station operation must be maintained, with specified data. The station license also authorizes the holder to operate portable and mobile stations on certain frequencies, subject to further regulations. An amateur station may be operated only by an amateur operator licensee, but any licensed amateur operator may operate any amateur station. All radio licensees are subject to penalties for violation of regulations.

Amateur licenses are issued entirely free of charge. They can be issued only to citizens but that is the only limitation, and they are given without regard to age or physical condition to anyone who successfully completes the examination. When you are able to copy 13 words per minute, have studied basic transmitter theory and are familiar with the law and amateur regulations, you are ready to give serious thought to securing the Government amateur licenses which are issued you, after examination at a local district office or examining points in most of our larger cities, through FCC at Washington. A complete up-to-the-minute discussion of license requirements, and a study guide for those preparing for the examination, are to be found in an ARRL publication, *The Radio Amateur's License Manual*, available from the American Radio Relay League, West Hartford 7, Conn., for 25¢, postpaid.

## LEARNING THE CODE

In starting to learn the code, you should consider it simply another means of conveying information. The spoken word is one method,

| | | | |
|---|---|---|---|
| A | didah | N | dahdit |
| B | dahdididit | O | dahdahdah |
| C | dahdidahdit | P | didahdahdit |
| D | dahdidit | Q | dahdahdidah |
| E | dit | R | didahdit |
| F | dididahdit | S | dididit |
| G | dahdahdit | T | dah |
| H | didididit | U | dididah |
| I | didit | V | didididah |
| J | didahdahdah | W | didahdah |
| K | dahdidah | X | dahdididah |
| L | didahdidit | Y | dahdidahdah |
| M | dahdah | Z | dahdahdidit |
| 1 | didahdahdahdah | 6 | dahdididit |
| 2 | dididahdahdah | 7 | dahdahdididit |
| 3 | didididahdah | 8 | dahdahdahdidit |
| 4 | dididididah | 9 | dahdahdahdahdit |
| 5 | dididididit | 0 | dahdahdahdahdah |

Period: didahdidahdidah. Comma: dahdahdididahdah. Question mark: dididahdahdidit. Error: dididididididididit. Double dash: dahdididi-dah. Wait: didahdididit. End of message: didahdidahdit. Invitation to transmit: dahdidah. End of work: didididahdidah. Fraction bar: dahdididahdit.

*Fig. 1-1* — The Continental (International Morse) code.

the printed page another, and typewriting and shorthand are additional examples. Learning the code is as easy — or as difficult — as learning to type.

The important thing in beginning to study code is to think of it as a language of *sound*, never as combinations of dots and dashes. It is easy to "speak" code equivalents by using "dit" and "*dah*," so that A would be "di*dah*" (the "t" is dropped in such combinations). The sound "di" should be staccato; a code character such as "5" should sound like a machine-gun burst: dididididit! Stress each "*dah*" equally; they are underlined or italicized in this text because they should be slightly accented and drawn out.

Take a few characters at a time. Learn them thoroughly in di*dah* language before going on to new ones. If someone who is familiar with code can be found to "send" to you, either by whistling or by means of a buzzer or code oscillator, enlist his coöperation. Learn the code by *listening* to it. Don't think about speed to start; the first requirement is to learn the characters to the point where you can recognize each of them without hesitation. Concentrate on any difficult letters.

## ACQUIRING SPEED BY BUZZER PRACTICE

Regular practice periods will develop code proficiency. Two people can learn the code together, sending to each other by means of a buzzer-and-key outfit. An advantage of this system is that it develops sending ability, too, for the person doing the receiving will be quick to criticize uneven or indistinct sending. If possible get an experienced operator for the first few sessions to learn how well-sent characters should sound.

Either the buzzer set shown in Figs. 1-2 and 1-3 or the audio oscillator described will give satisfactory results as a practice set. The battery-operated audio oscillator in Figs. 1-4 and 1-5 is easy to construct and is effective. If nothing is heard in the headphones when the key is depressed, reverse the leads going to *either* transformer winding (do not reverse both windings).

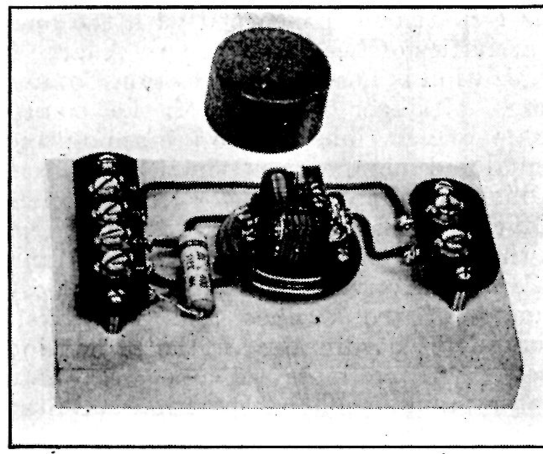


*Fig. 1-3* — **The cover of the buzzer unit has been removed in this view of the buzzer code-practice set.**

With a practice set ready, send single letters at first. When each character can be read quickly follow this by slow sending of complete words and sentences. Have the material sent at a rate slightly faster than you can copy easily; this speeds up your mind. *Write* down each letter you recognize. Do *not* try to write down the dots and dashes; write down letters. Don't stop to compare the sounds of different letters, or think too long about a letter or word that has been missed. Go right on to the next one, or each "miss" will cause you to lose several characters. If you exercise a little patience you will soon be getting every character. When you can receive 13 words a minute (65 letters a minute), have the sender transmit code groups rather than English text. This will prevent you from recognizing a word "on the way" and filling it in before you've really listened to the letters themselves.

After you have acquired reasonable proficiency, concentrate on the less common characters, as well as the numerals and punctuation. These prove the downfall of many applicants taking the code examination.

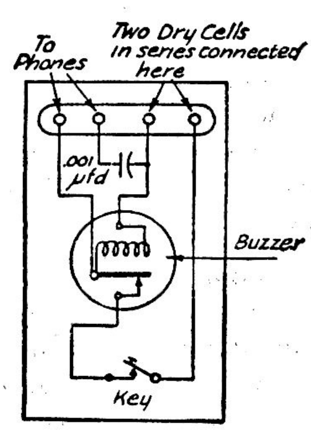



*Fig. 1-2* — **The headphones are connected across the coils of the buzzer, with a condenser in series. If the value shown gives an excessively loud signal, it may be reduced to 470 μμfd. or 220 μμfd.**

## LEARNING BY LISTENING

W1AW conducts practice transmissions nightly, Monday through Friday, at speeds from 9 to 35 w.p.m. Such practice tapes start at 10 P.M. EST (EDST in summer). In addition, the Official Bulletins, also sent from W1AW, give added practice at 15 and 25 w.p.m. See the Operating News section announcements of the W1AW operating schedule, and Code Proficiency Program notes, in the latest copy of *QST*. Practice until you can mail in what you have copied over the air on W1AW's monthly "qualifying run" to get a 15-word-per-minute Code Proficiency Certificate or a sticker for advanced speeds. As soon as you can, listen on a real communications receiver (with beat oscillator) and have the fun of learning by listening.

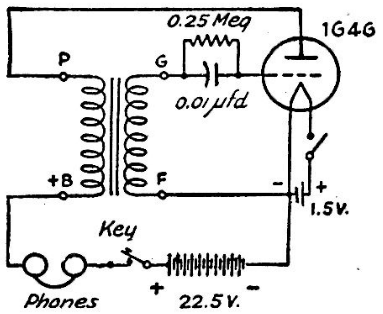



*Fig. 1-4* — Wiring diagram of a simple vacuum-tube audio-frequency oscillator for use as a code-practice set.

## ● USING A KEY

The correct way to grasp the key is important. The knob of the key should be about eighteen inches from the edge of the operating table and about on a line with the operator's right shoulder, allowing room for the elbow to rest on the table. A table about thirty inches in height is best. The spring tension of the key varies with different operators. A fairly heavy spring at the start is desirable. The back adjustment of the key should be changed until there is a vertical movement of about one-sixteenth inch at the knob. After an operator has mastered the use of the hand-key the tension should be changed and can be reduced to the minimum spring tension that will cause the key to open immediately when the pressure is released. More spring tension than necessary causes the expenditure of unnecessary energy. The contacts should be spaced by the rear screw on the key only and not by allowing play in the side screws, which are provided merely for aligning the contact points. These side screws should be screwed up to a setting which prevents appreciable side play, but not adjusted so tightly that binding is caused. The gap between the contacts should always be at least a thirty-second of an inch, since too-finely spaced contacts will cultivate a nervous style of sending which is highly undesirable. On the other hand, too-wide spacing (much over one-sixteenth inch), may result in unduly heavy or "muddy" sending.

Do not hold the key tightly. Let the hand rest lightly on the key. The thumb should be against the left side of the knob. The first and second fingers should be bent a little. They should hold the middle and right sides of the knob, respectively. The fingers are partly on top and partly over the side of the knob. The other two fingers should be free of the key. Fig. 1–6 shows the correct way to hold a key.

A wrist motion should be used in sending. The whole arm should not be used. One should not send "nervously" but with a steady flexing of the wrist. The grasp on the key should be firm, but not tight, or jerky sending will result. None of the muscles should be tense but they should all be under control. The arm should rest lightly on the operating table with the wrist held above the table. An up-and-down motion without any sideway action is best. The fingers should never leave the key knob.

Good sending may *seem* easier than receiving, but don't be deceived. A beginner should not attempt to send fast. Keep your transmitting speed down to your receiving speed, and bend your efforts to sending *well.* Do not try to speed things up too soon. A slow, even rate of sending is the mark of a good operator. Speed will come with time alone. Leave special types of keys alone until you have mastered the knack of handling the standard key. Because radio transmissions are seldom free from interference, a "heavier" style of sending is best to develop for radio work. A rugged, heavy key will help in developing this characteristic.

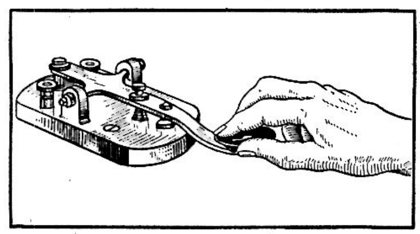


*Fig. 1-6* — This sketch illustrates the correct position of the hand and fingers for good sending with a telegraph key.

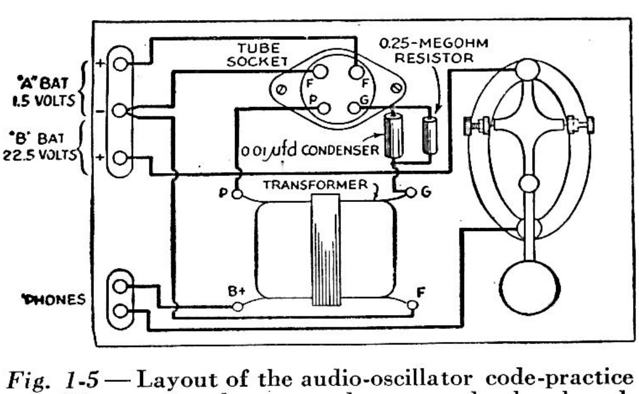



*Fig. 1-5* — Layout of the audio-oscillator code-practice set. All parts may be mounted on a wooden baseboard, approximately 5 × 7 inches in size.

To become expert in transmitting good code, after you have thoroughly learned each letter and numeral and can both send and copy letters without hesitation, your best practice is to listen to commercial automatic-tape stations. Perfectly-sent code can be accomplished only by a machine, and you want to get fixed in your mind, indelibly, the correct formation of each and every code character and in particular the associated spaces. One of the best methods for deriving this association is to find a commercial or other tape station sending at about your maximum receiving speed. Notice the formation of each letter, the spaces left between letters and words, and the proportion in length of dits to dahs. Listen to the transmissions as you would at a musical concert, concentrating on assimilating every detail. The spaces between words may seem exaggerated, simply because you have probably been running yours together. A score of other details where the automatic transmission is different than yours will very likely show up in the same text. From all this you will learn where your own faults lie and be able to correct them.

## THE AMATEUR BANDS

Amateurs are assigned bands of frequencies at approximate octave intervals throughout the spectrum. Like assignments to all services, they are subject to modification to fit the changing picture of world communications needs.

Below is a summary of the U. S. amateur bands on which operation is permitted as of our press date. Figures are megacycles. AØ means an unmodulated carrier, A1 means c.w. telegraphy, A2 is m.c.w., A3 is AM 'phone, A4 is facsimile, A5 is television; NBFM designates narrow-band frequency- or phase-modulated radiotelephony; and FM means frequency modulation, 'phone (including NBFM) or telegraphy.

| | |
|---|---|
| 3.500– 4.000 | A1 |
| 3.850– 4.000 | A3, Class A only |
| 3.850– 3.900 | NBFM, Class A only |
| 7.000– 7.300 | A1 |
| 14.000–14.400 | A1 |
| 14.200–14.300 | A3, Class A only |
| 14.200–14.250 | NBFM, Class A only |
| 27.160–27.430 | AØ, A1, A2, A3, A4, FM |
| 28.000–29.700 | A1 |
| 28.500–29.700 | A3 |
| 28.500–29.000 | NBFM |
| 29.000–29.700 | FM |
| 50.0 –54.0 | A1, A2, A3, A4 |
| 51.0 –52.5 | NBFM |
| 52.5 –54.0 | FM |
| 144 –148 | AØ, A1, A2, A3, A4, FM |
| 235 –240 | AØ, A1, A2, A3, A4, FM |
| 420* –450* | AØ, A1, A2, A3, A4, A5, FM |
| 1,215 – 1,295 | |
| 2,300 – 2,450 | |
| 3,300 – 3,500 | |
| 5,650 – 5,925 | AØ, A1, A2, A3, A4, A5, FM, Pulse |
| 10,000 –10,500 | |
| 21,000 –22,000 | |
| All above 30,000 | |

* Peak antenna power must not exceed 50 watts.

The future of the prewar amateur band at 1.75 Mc. has not been finally determined, it depending principally on the disposition of the navigational service now operating there. The 1947 International Radio Conference has resulted in certain changes in present bands expected to become effective late in 1949. They are: a reduction in the 20-meter band to make it thenceforth 14,000–14,350 kc.; a new band 21,000–21,450 kc.; and a shift in the 11-meter band to make it 26,960–27,230 kc. Additional changes, probably effective earlier than 1949, will shift the 1¼-meter band to its permanent location 220–225 Mc., and expand the 1215-Mc. band so that its top limit will be 1300 Mc. Because of the possibility of such changes, and because the portion of each band available for 'phone operation is customarily varied from time to time in accordance with changes in amateur operational habits, in such respects each amateur should keep himself currently informed by consulting *QST* or writing ARRL for latest information.

# Electrical Laws and Circuits

Everyone knows that radio is electrical in nature, and it is taken for granted that to know anything about the operation of radio equipment you have first to know something about electricity and electrical circuits. The amount of electrical knowledge you need in amateur radio depends on how far you delve into the technicalities of the various types of transmitters, receivers and measuring equipment that amateurs use. If you're just getting started you do not need very much, but as you progress you will find that you will acquire, more or less unconsciously, a great deal of basic information. That is, you will if you make a conscientious effort to understand and analyze the things that you observe in using radio gear.

The purpose of this chapter is to provide the answers to many questions about circuits that will come up in the course of building and operating an amateur station. It is intended as a *practical* reference section rather than a course in "theory." You can study it consecutively if you wish, of course. However, it should be even more valuable to you in showing how everyday problems can be solved when the occasion to solve them arises.

## Fundamentals

### ● ELECTRIC AND MAGNETIC FIELDS

At the bottom of everything in electricity and radio is a **field.** Although a field is not too easy to visualize, we need to have some appreciation of what it is if electrical effects are to be understood. When something occurs at one point in space because something else happened at another point, with no visible means by which the "cause" can be related to the "effect," we say the two events are connected by a "field." It does not matter whether or not the field is "real" — that is, whether it is something physical although, like air, invisible. The important point is that the distant effects are *predictable,* and it is convenient to attribute them to properties of a field. The fields with which we are concerned are the **electric** and **magnetic,** and the combination of the two called the **electromagnetic** field.

A field has two important properties, *intensity* (*magnitude*) and *direction.* That is, the field exerts a *force* on an object immersed in it; intensity measures the amount of force exerted while direction tells the direction in which the object on which the force is exerted will tend to move. An electrically-charged object in an electric field will be acted on by a force that will tend to move it in a direction determined by the direction of the field. Similarly, a magnet in a magnetic field will be subject to a force. Everyone has seen demonstrations of magnetic fields with pocket magnets, so intensity and direction are not hard to grasp.

A "static" field is one that is fixed in space. Such a field can be set up by a stationary electric charge **(electrostatic field)** or by a stationary magnet **(magnetostatic field).** But if either an electric or magnetic field is moving in space or changing in intensity, the motion or change sets up the other kind of field. That is, a changing electric field sets up a magnetic field, and a changing magnetic field generates an electric field. This interrelationship between magnetic and electric fields makes possible such things as the electromagnet and the electric motor. It also makes possible the **electromagnetic waves** by which radio communication is carried on, for such waves are simply traveling fields in which the energy is alternately handed back and forth between the electric and magnetic fields.

#### *Lines of Force*

We need, obviously, some way to compare the intensity and direction of different fields. This is done by picturing the field as made up of **lines of force,** or **flux lines.** These are purely imaginary threads that show, by the direction in which they lie, the direction the object on which the force is exerted will move. The *number* of lines in a chosen cross section of the field is a measure of the *intensity* of the force. The number of lines per square inch, or per square centimeter, is called the **flux density.**

## ELECTRICITY AND THE ELECTRIC CURRENT

Electrical effects are caused by extremely small particles of electricity called **electrons.** Everything physical is built up of atoms, particles so small that they cannot be seen even through the most powerful microscope. But the atom in turn consists of still smaller particles — several different kinds of them. One type of particle is the electron. An ordinary atom consists of a central core, called the **nucleus,** around which one or more electrons circulate somewhat as the earth and other planets circulate around the sun. Both the nucleus and the electrons are electrical, but the kind of electricity associated with the nucleus is called **positive** and that associated with the electrons is called **negative.**

The important fact about these two "opposite" kinds of electricity is that they are strongly attracted to each other. Also, there is a strong force of repulsion between two charges (a collection of electrified particles is called a **charge**) of the *same* kind. The positive nucleus and the negative electrons are attracted to each other, but two electrons will be repelled from each other and so will two nuclei. The fact that an atom contains both positive and negative charges makes it tend to stay together as a unit; in a normal atom the positive charge on the nucleus is exactly balanced by the total of the negative charges on the electrons. It *is* possible, though, for an atom to lose one of its electrons; when that happens the atom has a little less negative charge than it should — or, to put it another way, it has a net positive charge. Such an atom is said to be **ionized,** and in this case the atom is a **positive ion.** If an atom picks up an extra electron, as it sometimes does, it has a net negative charge and is called a **negative ion.** A positive ion will attract any stray electron in the vicinity, including the extra one that may be attached to a nearby negative ion. In this way it is conveniently possible for electrons to travel from atom to atom, and when such movement occurs on a measurable scale (millions or billions of electrons moving) we have a detectable **electric current.**

### *Conductors and Insulators*

The movement of electrons can take place in a solid, a liquid, or a gas. In liquids and gases, positive and negative ions, as well, are free to move when attracted electrically, but in solids only the electrons move. However, movement of electrons or ions is not possible in all substances. Atoms of some materials, notably metals and acids, will give up an electron readily, but atoms of other materials will not part with any of their electrons even when the electric force is extremely strong. Materials in which electrons or ions can be moved with relative ease are called **conductors,** while those that refuse to permit such movement are called **nonconductors** or **insulators.** The following listing shows how some common materials divide between the conductor and insulator classifications:

| *Conductors* | *Insulators* |
|---|---|
| Metals | Dry Air |
| Carbon | Wood |
| Acids | Porcelain |
| | Textiles |
| | Glass |
| | Rubber |
| | Resins |

### *Electromotive Force*

The electric force (called **electromotive force,** and abbreviated **e.m.f.**) that causes current flow may be developed in several ways. The action of certain chemical solutions on dissimilar metals sets up an e.m.f.; such a combination is called a **cell,** and a group of cells forms an electric **battery.** The amount of current that such cells can carry is limited, and in the course of current flow one of the metals is eaten away. The amount of electrical energy that can be taken from a battery consequently is rather small. Where a large amount of energy is needed it is usually furnished by an electric **generator,** which develops its e.m.f. by a combination of magnetic and mechanical means. Large generators in power houses supply the energy that is distributed to homes and factories.

In picturing current flow it is natural to think of a single, constant force causing the electrons to move. When this is so, the elec-

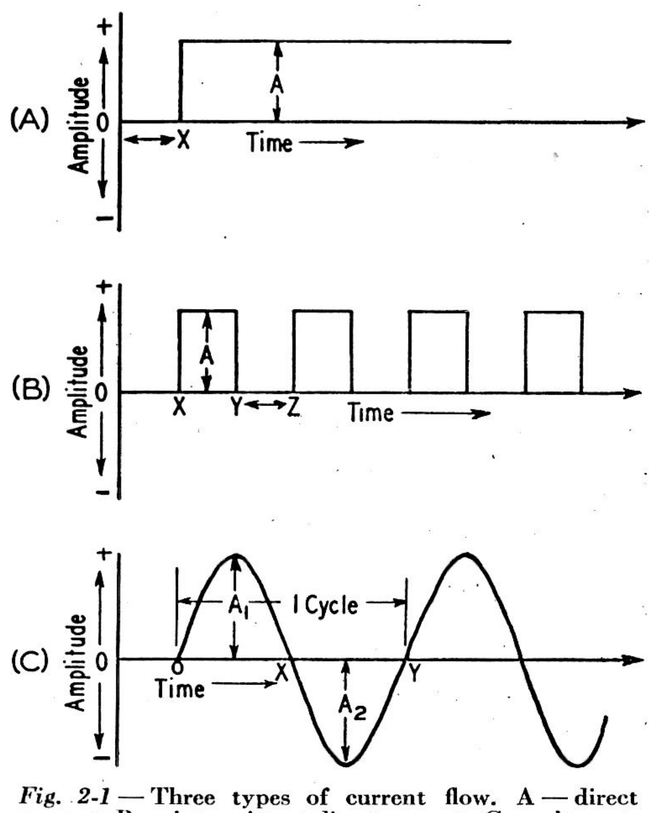



*Fig. 2-1* — Three types of current flow. A — direct current; B — intermittent direct current; C — alternating current.

trons always move in the same direction through a path or circuit made up of conductors connected together in a continuous chain. Such a current is called a **direct current**, abbreviated **d.c.** It is the type of current furnished by batteries and by certain types of generators. However, it is also possible — and desirable as well — to have an e.m.f. that periodically reverses. With this kind of e.m.f. the current flows first in one direction through the circuit and then in the other. Such an e.m.f. is called an **alternating** e.m.f., and the current is called an **alternating current** (abbreviated **a.c.**). The reversals **(alternations)** may occur at any rate from a few per second up to several billion per second. Two reversals make a **cycle**; in one cycle the force acts first in one direction, then in the other, and then returns to the first direction. The number of cycles in one second is called the **frequency** of the alternating current.

### *Direct and Alternating Currents*

The difference between direct current and alternating current is shown in Fig. 2-1. In these graphs the horizontal axis measures time, increasing toward the right away from the vertical axis. The vertical axis represents the amplitude or size of the current, increasing in either the up or down direction away from the horizontal axis. If the graph is *above* the horizontal axis the current is flowing in one direction through the circuit (indicated by the + sign) and if it is *below* the horizontal axis the current is flowing in the reverse direction through the circuit (indicated by the — sign). Fig. 2-1A shows that, if we close the circuit — that is, make the path for the current complete — at the time indicated by $X$, the current instantly takes the amplitude indicated by the height $A$. After that, the current continues at the same amplitude as time goes on. This is an ordinary *direct* current.

In Fig. 2-1B, the current starts flowing with the amplitude $A$ at time $X$, continues at that amplitude until time $Y$ and then instantly ceases. After an interval $YZ$ the current again begins to flow and the same sort of start-and-stop performance is repeated. This is an *intermittent* direct current. We could get it by alternately closing and opening a switch in the circuit. It is a *direct* current because the *direction* of current flow does not change; the graph is always on the + side of the horizontal axis.

In Fig. 2-1C the current starts at zero, increases in amplitude as time goes on until it reaches the amplitude $A_1$ while flowing in the + direction, then decreases until it drops to zero amplitude once more. At that time ($X$) the *direction* of the current flow reverses; this is indicated by the fact that the next part of the graph is below the axis. As time goes on the amplitude increases, with the current now flowing in the — direction, until it reaches amplitude $A_2$. Then the amplitude decreases until finally it drops to zero ($Y$) and the direc-

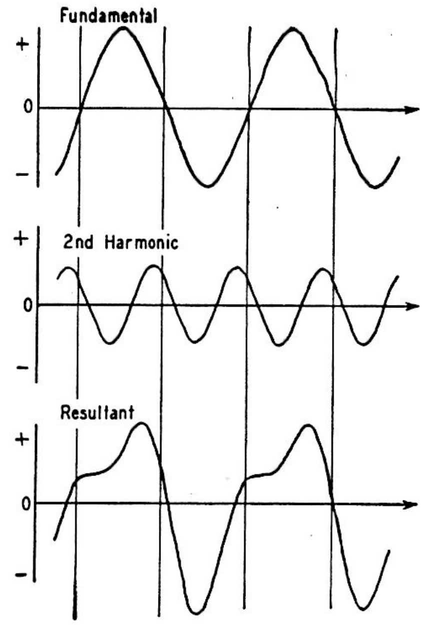



*Fig. 2-2* — A complex waveform. A fundamental (top) and second harmonic (center) added together, point by point at each instant, result in the waveform shown at the bottom. When the two components have the same polarity at a selected instant, the resultant is the simple sum of the two. When they have opposite polarities, the resultant is the *difference*; if the negative-polarity component is larger, the resultant is negative at that instant.

tion reverses once more. This is an *alternating* current.

### *Waveforms*

The graph of the alternating current is what is known as a **sine wave.** Sine-wave alternating current is the simplest — but not the only — kind. Notice that the variations in amplitude are quite regular and that the "negative" half-cycle or alternation is exactly like the "positive" half-cycle except for the reversal of direction. The variations in many a.c. waves are not so smooth, nor is one half-cycle necessarily just like the preceding one in shape. However, these more complex waves actually can be shown to be the sum of two or more sine waves of frequencies that are exact integral (whole-number) multiples of some lower frequency. The lowest frequency is called the **fundamental** frequency, and the higher frequencies (2 times, 3 times the fundamental frequency, and so on) are called **harmonics.**

Fig. 2-2 shows how a fundamental and a second harmonic (twice the fundamental) might add to form a complex wave. A little thought will show that simply by changing the relative amplitudes of the two waves, as well as the times at which they pass through zero amplitude, an infinite number of waveshapes can be constructed from just a fundamental and second harmonic. Waves that are still more complex can be constructed if more than two harmonics are used.

### *Electrical Units*

The unit of electromotive force is called the **volt.** An ordinary flashlight cell generates an e.m.f. of about 1.5 volts. The e.m.f. commonly supplied for domestic lighting and power is 115 volts, usually a.c. having a frequency of 60 cycles per second. The voltages used in radio receiving and transmitting circuits range from a few volts (usually a.c.) for filament heating to as high as a few thousand d.c. volts for the operation of power tubes.

The flow of electric current is measured in **amperes.** One ampere is equivalent to the movement of many billions of electrons past a point in the circuit in one second. Currents in the neighborhood of an ampere are required for heating the filaments of small power tubes. The *direct* currents used in amateur radio equipment usually are not so large, and it is customary to measure such currents in **milliamperes.** One milliampere is equal to one one-thousandth of an ampere, or 1000 milliamperes equals one ampere.

In assigning a value to an alternating current or voltage, it is necessary to take into account the difference between direct and alternating currents. A "d.c. ampere" is a measure of a *steady* current, but the "a.c. ampere" must measure a current that is continually varying in amplitude and periodically reversing direction. To put the two on the same basis, an a.c. ampere is defined as the amount of current that will cause the same heating effect (see later section) as one ampere of steady direct current. For a sine-wave alternating current, this **effective** (or **r.m.s.**) **value** is equal to the *maximum* amplitude of the current ($A_1$ or $A_2$ in Fig. 2-1C) multiplied by 0.707. The **instantaneous value** of an alternating current is the value that the current measures at any selected instant in the cycle.

If all the instantaneous values in a sine-wave alternating current are averaged over a *half*-cycle, the resulting figure is the **average value** of the alternating current. It is equal to 0.636 times the maximum amplitude. The average value is useful in connection with rectifier systems, as described in a later chapter.

These definitions of units apply to a.c. voltage as well as to current.

## FREQUENCY AND WAVELENGTH

### *Frequency Spectrum*

The electrical energy supplied for household use usually has a frequency of 60 cycles per second. Frequencies ranging from about 15 to 15,000 cycles per second are called **audio** frequencies, because the vibrations of air particles that our ears recognize as sounds occur at the same rate. Audio frequencies (abbreviated **a.f.**) are used to actuate loudspeakers and thus create sound waves.

Frequencies above about 15,000 cycles are called **radio** frequencies **(r.f.)** because they are useful in radio transmission. Frequencies all the way up to and beyond 10,000,000,000 cycles have been used for radio purposes. At radio frequencies the numbers become so large that it becomes convenient to use a larger unit than the cycle. Two such units in everyday use are the **kilocycle,** which is equal to 1000 cycles and is abbreviated **kc.,** and the **megacycle,** which is equal to 1,000,000 cycles or 1,000 kilocycles and is abbreviated **Mc.** The accompanying table shows how to convert frequencies expressed in one unit into frequencies in another unit.

The various radio frequencies are divided off into classifications for ready identification. These classifications, listed below, constitute the **frequency spectrum** so far as it extends for radio purposes at the present time.

| *Frequency* | *Classification* | *Abbreviation* |
|---|---|---|
| 10 to 30 kc. | Very-low frequencies | v.l.f. |
| 30 to 300 kc. | Low frequencies | l.f. |
| 300 to 3000 kc. | Medium frequencies | m.f. |
| 3 to 30 Mc. | High frequencies | h.f. |
| 30 to 300 Mc. | Very-high frequencies | v.h.f. |
| 300 to 3000 Mc. | Ultrahigh frequencies | u.h.f. |
| 3000 to 30,000 Mc. | Superhigh frequencies | s.h.f. |

### *Wavelength*

We said earlier that radio waves are traveling fields of electric and magnetic force. These fields travel at great speed — so great that, so far as we can observe, "cause" and "effect" are simultaneous. Nevertheless, it does take a definite amount of time for the effect of a field set up at one point to be felt at a point some distance away.

Radio waves travel at the same speed as light — 300,000,000 meters or about 186,000 miles a second. They are always set up by a radio-frequency current flowing in a circuit, because the rapidly-changing current sets up a magnetic field that changes in the same way, and the varying magnetic field in turn sets up a varying electric field. And whenever this happens, the two fields move outward at the speed of light.

Suppose our r.f. current has a frequency of 3,000,000 cycles per second. The fields, then, will go through complete reversals (one cycle) in 1/3,000,000 second. In that same period of time the fields — that is, the wave — will move 300,000,000/3,000,000 meters, or 100 meters. (The meter is the unit of length commonly used in all sciences. We could use miles, feet, or inches, though, if those units were more convenient.) By the time the wave has moved that distance the next cycle has begun and a new wave has started out. The first wave, in other words, covers a distance of 100 meters before the beginning of the next, and so on. This distance is the "length" of the wave, or **wavelength.**

The longer the time of one cycle — that is, the lower the frequency — the greater the distance occupied by each wave and hence the longer the wavelength. The relationship between wavelength and frequency is shown by the formula

$$\lambda = \frac{300{,}000}{f}$$

where $\lambda$ = Wavelength in meters
$f$ = Frequency in kilocycles

or

$$\lambda = \frac{300}{f}$$

where $\lambda$ = Wavelength in meters
$f$ = Frequency in megacycles

Example: The wavelength corresponding to a frequency of 3650 kilocycles is

$$\lambda = \frac{300{,}000}{3650} = 82.2 \text{ meters}$$

Most of our dealings are with frequency, if for no other reason than that it can be measured much more accurately than wavelength. However, we cannot ignore wavelength; it enters into the calculation of the size of "linear" circuits such as antennas.

# Resistance

The ease with which we can force an electric current through a conductor varies with the material, shape and dimensions of the conductor. Given two conductors of the same size and shape, but of different materials, the amount of current that will flow when a given e.m.f. is applied to the conductor will be found to vary with what is called the **resistance** of the material. The lower the resistance, the greater the current for a given value of e.m.f.

Resistance is measured in **ohms.** A circuit has a resistance of one ohm when an applied e.m.f. of one volt causes a current of one ampere to flow. The **resistivity** of a material is the resistance, in ohms, of a cube of the material measuring one centimeter on each edge. One of the best conductors is copper, which is why this metal is so widely used in electrical circuits. It is frequently convenient, in making resistance calculations, to compare the resistance of the material under consideration with that of a copper conductor of the same size and shape; Table 2-I gives the ratio of the resistivity of the material to that of copper.

The longer the path through which the current flows the higher the resistance of that conductor. For direct current and low-frequency alternating currents (up to a few thousand cycles per second) the resistance is *inversely* proportional to the cross-sectional area of the path the current must travel; that is, given two conductors of the same material and having the same length, but differing in cross-sectional area, the one with the larger area will have the lower resistance.

### *Resistance of Wires*

It is readily possible to combine all these statements about resistance in a single formula that would enable us to calculate the resistance of conductors of any size, shape and material. However, in most practical cases the problem will be to determine the resistance of a round wire of given diameter and length — or its opposite: finding a suitable size and length of wire to supply a desired amount of resistance. Such problems can be easily solved with the help of the information in the copper-wire table in Chapter Twenty-Four. This table gives the resistance, in ohms per thousand feet, of each standard wire size.

Example: Suppose a resistance of 3.5 ohms is needed and some No. 28 wire is on hand. The wire table in Chapter 24 shows that No. 28 has a resistance of 66.17 ohms per thousand feet. Since the desired resistance is 3.5 ohms, the length of wire required will be

$$\frac{3.5}{66.17} \times 1000 = 52.89 \text{ feet.}$$

Or, suppose that the resistance of the wire in the circuit must not exceed 0.05 ohm and that the length of wire required for making the connections totals 14 feet. Then

$$\frac{14}{1000} \times R = 0.05 \text{ ohm}$$

where $R$ is the maximum allowable resistance in ohms per thousand feet. Rearranging the formula gives

$$R = \frac{0.05 \times 1000}{14} = 3.57 \text{ ohms/1000 ft.}$$

Reference to the wire table shows that No. 15 is the smallest size having a resistance less than this value.

When the wire is not copper, the resistance values given in the wire table in Chapter Twenty-Four should be multiplied by the ratios given in Table 2-I to obtain the resistance.

Example: If the wire in the first example were iron instead of copper the length required for 3.5 ohms would be

$$\frac{3.5}{66.17 \times 5.65} \times 1000 = 9.35 \text{ feet.}$$

### *Temperature Effects*

The resistance of a conductor changes with its temperature. Although it is seldom necessary to consider temperature in making the

**TABLE 2-I**
**Relative Resistivity of Metals**

| *Material* | *Resistivity Compared to Copper* |
|---|---|
| Aluminum (pure) | 1.70 |
| Brass | 3.57 |
| Cadmium | 5.26 |
| Chromium | 1.82 |
| Copper (hard-drawn) | 1.12 |
| Copper (annealed) | 1.00 |
| Iron (pure) | 5.65 |
| Lead | 14.3 |
| Nickel | 6.25 to 8.33 |
| Phosphor Bronze | 2.78 |
| Silver | 0.94 |
| Tin | 7.70 |
| Zinc | 3.54 |

resistance calculations required in amateur work, it is well to know that the resistance of practically all metallic conductors increases with increasing temperature. Carbon, however, acts in the opposite way; its resistance *decreases* when its temperature rises. The temperature effect is important when it is necessary to maintain a constant resistance under all conditions. Special materials that have little or no change in resistance over a wide temperature range are used in that case.

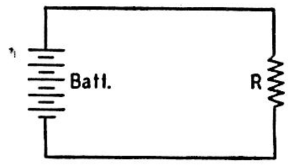
### *Resistors*

Resistance has important uses in electrical and radio circuits. A "package" of resistance made up into a single unit is called a **resistor.** Resistors having the same resistance value may be considerably different in size and construction. The flow of current through resistance causes the conductor to become heated; the higher the resistance and the larger the current, the greater the amount of heat developed. Consequently, high-resistance resistors intended for carrying large currents must be physically large so the heat can be radiated quickly to the surrounding air. If the resistor does not get rid of its heat quickly it might reach a temperature that would cause it to melt or burn. Types of resistors used in radio circuits are shown in the photograph.

### *Conductance*

The reciprocal of resistance (that is, $1/R$) is called **conductance.** It is usually represented by the symbol $G$, and the higher its value the greater the conductivity of the circuit. A circuit having large conductance has low resistance, and vice versa. In radio work the term is used chiefly in connection with vacuum-tube characteristics. The unit of conductance is the **mho.** A resistance of one ohm has a conductance of one mho, a resistance of 1000 ohms has a conductance of 0.001 mho, and so on. A unit frequently used in connection with vacuum tubes is the **micromho,** or one-millionth of a mho. It is the conductance of a resistance of one megohm.

*Fig. 2-3* — A simple circuit consisting of a battery and resistor.




## OHM'S LAW

The simplest form of electric circuit is a battery with a resistance connected to its terminals, as shown by the symbols in Fig. 2-3. A complete circuit must have an unbroken path so current can flow out of the battery, through the apparatus connected to it, and back into the battery. The circuit is **broken,** or **open,** if a connection is removed at any point. A **switch** is a device for making and breaking connections and thereby closing or opening the circuit, either allowing current to flow or preventing it from flowing.

The values of current, voltage and resistance in a circuit are by no means independent of each other. The relationship between them is known as **Ohm's Law.** It can be stated as follows: The current flowing in a circuit is directly proportional to the applied e.m.f. and inversely proportional to the resistance. Expressed as an equation, it is

$$I \text{ (amperes)} = \frac{E \text{ (volts)}}{R \text{ (ohms)}}$$

The equation above gives the value of current when the voltage and resistance are known. It may be transposed so that any of the three quantities may be found when the other two are known:

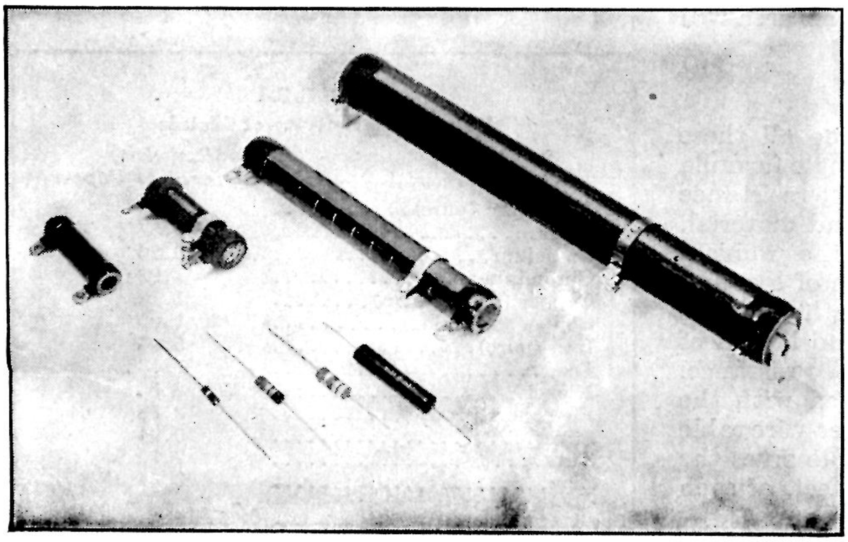
$$E = IR$$

(that is, the voltage acting is equal to the current in amperes multiplied by the resistance in ohms) and

$$R = \frac{E}{I}$$

(or, the resistance of the circuit is equal to the applied voltage divided by the current).

All three forms of the equation are used almost constantly in radio work. It must be



Types of resistors used in radio equipment. Those in the foreground with wire leads are carbon types, ranging in size from ½ watt at the left to 2 watts at the right. The larger resistors use resistance wire wound on ceramic tubes; sizes shown range from 5 watts to 100 watts. Three are the adjustable type, using a sliding contact on an exposed section of the resistance winding.

remembered that the quantities are in *volts*, *ohms* and *amperes*; other units cannot be used in the equations without first being converted. For example, if the current is in milliamperes it must be changed to the equivalent fraction of an ampere before the value can be substituted in the equations.

Table 2-II shows how to convert between the various units in common use. The prefixes attached to the basic-unit name indicate the nature of the unit. These prefixes are:

micro — one-millionth (abbreviated $\mu$)
milli — one-thousandth (abbreviated $m$)
kilo — one thousand (abbreviated $k$)
mega — one million (abbreviated $M$)

For example, one microvolt is one-millionth of a volt, and one megohm is 1,000,000 ohms. There are therefore 1,000,000 microvolts in one volt, and 0.000001 megohm in one ohm.

The following examples illustrate the use of Ohm's Law:

The current flowing in a resistance of 20,000 ohms is 150 milliamperes. What is the voltage? Since the voltage is to be found, the equation to use is $E = IR$. The current must first be converted from milliamperes to amperes, and reference to the table shows that to do so it is necessary to divide by 1000. Therefore,

$$E = \frac{150}{1000} \times 20{,}000 = 3000 \text{ volts}$$

When a voltage of 150 is applied to a circuit the current is measured at 2.5 amperes. What is the resistance of the circuit? In this case $R$ is the unknown, so

$$R = \frac{E}{I} = \frac{150}{2.5} = 60 \text{ ohms}$$

No conversion was necessary because the voltage and current were given in volts and amperes.

How much current will flow if 250 volts is applied to a 5000-ohm resistor? Since $I$ is unknown,

$$I = \frac{E}{R} = \frac{250}{5000} = 0.05 \text{ ampere}$$

Milliampere units would be more convenient for the current, and 0.05 amp. $\times$ 1000 = 50 milliamperes.

## SERIES AND PARALLEL RESISTANCES

Very few actual electric circuits are as simple as the illustration in the preceding section. Commonly, resistances are found connected in a variety of ways. The two fundamental methods of connecting resistances are shown in Fig. 2-4. In the upper drawing, the current flows from the source of e.m.f. (in the direction shown by the arrow, let us say) down through the first resistance, $R_1$, then through the second, $R_2$, and then back to the source. These resistors are connected in **series.** The current everywhere in the circuit has the same value.

In the lower drawing the current flows to the common connection point at the top of the two resistors and then divides, one part of it flowing through $R_1$ and the other through $R_2$. At the lower connection point these two currents again combine; the total is the same as the current that flowed into the upper common connection. In this case the two resistors are connected in **parallel.**

**TABLE 2-II**
**Conversion Values for Fractional and Multiple Units**

| *To change from* | *To* | *Divide by* | *Multiply by* |
|---|---|---|---|
| Units | Micro-units | | 1,000,000 |
| | Milli-units | | 1000 |
| | Kilo-units | 1000 | |
| | Mega-units | 1,000,000 | |
| Micro-units | Milli-units | 1000 | |
| | Units | 1,000,000 | |
| Milli-units | Micro-units | | 1000 |
| | Units | 1000 | |
| Kilo-units | Units | | 1000 |
| | Mega-units | 1000 | |
| Mega-units | Units | | 1,000,000 |
| | Milli-units | | 1000 |

### *Resistors in Series*

When a circuit has a number of resistances connected in series, the total resistance of the circuit is the sum of the individual resistances. If these are numbered $R_1$, $R_2$, $R_3$, etc., then

$$R \text{ (total)} = R_1 + R_2 + R_3 + R_4 + \ldots.$$

where the dots indicate that as many resistors as necessary may be added.

Example: Suppose that three resistors are connected to a source of e.m.f. as shown in Fig. 2-5. The e.m.f. is 250 volts, $R_1$ is 5000 ohms, $R_2$ is 20,000 ohms, and $R_3$ is 8000 ohms. The total resistance is then

$$R = R_1 + R_2 + R_3 = 5000 + 20{,}000 + 8000 = 33{,}000 \text{ ohms}$$

The current flowing in the circuit is then

$$I = \frac{E}{R} = \frac{250}{33{,}000} = 0.00757 \text{ amp.} = 7.57 \text{ ma.}$$

(We need not carry calculations beyond three significant figures, and often two will suffice because the accuracy of measurements is seldom better than a few per cent.)

### *Voltage Drop*

Ohm's Law applies to *any part* of a circuit as well as to the whole circuit. Although the current is the same in all three of the resistances in the example, the total voltage divides

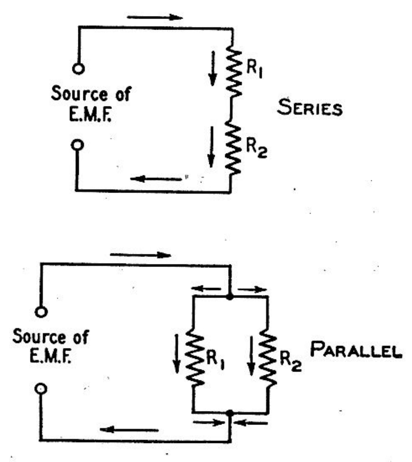



*Fig. 2-4* — Resistors connected in series and in parallel.

among them. The voltage appearing across each resistor can be found from Ohm's Law.

Example: If the voltage across $R_1$ (Fig. 2-5) is called $E_1$, that across $R_2$ is called $E_2$, and that across $R_3$ is called $E_3$, then

$$E_1 = IR_1 = 0.00757 \times 5000 = 37.9 \text{ volts}$$
$$E_2 = IR_2 = 0.00757 \times 20{,}000 = 151.4 \text{ volts}$$
$$E_3 = IR_3 = 0.00757 \times 8000 = 60.6 \text{ volts}$$

The total voltage must equal the sum of the individual voltage drops:

$$E = E_1 + E_2 + E_3 = 37.9 + 151.4 + 60.6 = 249.9 \text{ volts}$$

The answer would have been more nearly exact if the current had been calculated to more decimal places, but as explained above a very high order of accuracy is not necessary.

In a simple series circuit like that in Fig. 2-5, the voltage drop across each resistance can be calculated very simply, if only the drop and not the current is wanted. The drop across each resistor is proportional to the ratio of the individual resistance to the *total* resistance. Thus

$$E_1 = \frac{R_1}{R_1 + R_2 + R_3} \times 250$$
$$= \frac{5000}{5000 + 20{,}000 + 8000} = \frac{5000}{33{,}000} \times 250 = 37.8 \text{ volts}$$

$$E_2 = \frac{20{,}000}{33{,}000} \times 250 = 151.5 \text{ volts}$$

$$E_3 = \frac{8000}{33{,}000} \times 250 = 60.5 \text{ volts}$$

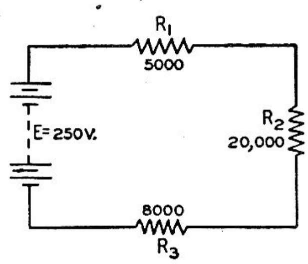



*Fig. 2-5* — **An example of resistors in series. The solution of the circuit is worked out in the text.**

In problems such as this considerable time and trouble can be saved, when the current is small enough to be expressed in milliamperes, if the resistance is expressed in kilohms rather than ohms. When resistance in kilohms is substituted directly in Ohm's Law the current will be in milliamperes if the e.m.f. is in volts.

Example: Since 5000 ohms = 5 kilohms, 20,000 ohms = 20 kilohms, and 8000 ohms = 8 kilohms. the equations above become

$$I = \frac{E}{R} = \frac{250}{33} = 7.57 \text{ ma.}$$

$$E_1 = IR_1 = 7.57 \times 5 = 37.9 \text{ volts}$$
$$E_2 = IR_2 = 7.57 \times 20 = 151.4 \text{ volts}$$
$$E_3 = IR_3 = 7.57 \times 8 = 60.6 \text{ volts}$$

### *Resistors in Parallel*

In a circuit with resistances in parallel, the total resistance is *less* than that of the *lowest* value of resistance present. This is because the total current is always greater than the current in any individual resistor. The formula for finding the total resistance of resistances in parallel is

$$R = \cfrac{1}{\cfrac{1}{R_1} + \cfrac{1}{R_2} + \cfrac{1}{R_3} + \cfrac{1}{R_4} + \cdots\cdots}$$

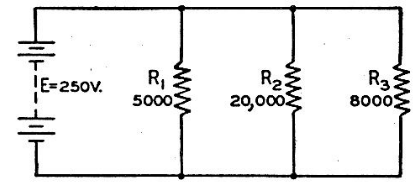



*Fig. 2-6* — **An example of resistors in parallel. The solution is worked out in the text.**

where the dots again indicate that any number of resistors can be combined by the same method. For only two resistances in parallel (a very common case) the formula is

$$R = \frac{R_1 R_2}{R_1 + R_2}$$

Example: If a 500-ohm resistor is paralleled with one of 1200 ohms, the total resistance is

$$R = \frac{R_1 R_2}{R_1 + R_2} = \frac{500 \times 1200}{500 + 1200} = \frac{600{,}000}{1700} = 353 \text{ ohms}$$

It is probably easier to solve practical problems by a different method than the "reciprocal of reciprocals" formula. Suppose the three resistors of the previous example are connected in parallel as shown in Fig. 2-6. The same e.m.f., 250 volts, is applied to all three of the resistors. The current in each can be found from Ohm's Law as shown below, $I_1$ being the current through $R_1$, $I_2$ the current through $R_2$ and $I_3$ the current through $R_3$.

For convenience, the resistance will be expressed in kilohms so the current will be in milliamperes.

$$I_1 = \frac{E}{R_1} = \frac{250}{5} = 50 \text{ ma.}$$

$$I_2 = \frac{E}{R_2} = \frac{250}{20} = 12.5 \text{ ma.}$$

$$I_3 = \frac{E}{R_3} = \frac{250}{8} = 31.25 \text{ ma.}$$

The total current is

$$I = I_1 + I_2 + I_3 = 50 + 12.5 + 31.25 = 93.75 \text{ ma.}$$

The total resistance of the circuit is therefore

$$R = \frac{E}{I} = \frac{250}{93.75} = 2.66 \text{ kilohms } (= 2660 \text{ ohms})$$

### *Resistors in Series-Parallel*

An actual circuit may have resistances both in parallel and in series. To illustrate, we use the same three resistances again, but now connected as in Fig. 2-7. The method of solving such a circuit is as follows: Consider $R_2$ and $R_3$ in parallel as though they formed a single resistor. Find their equivalent resistance. Then this resistance in series with $R_1$ forms a simple

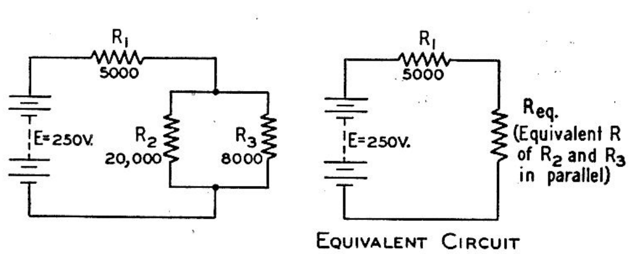



*Fig. 2-7* — **An example of resistors in series-parallel. The solution is worked out in the text.**

series circuit, as shown at the right in Fig. 2-7.

Example: The first step is to find the equivalent resistance of $R_2$ and $R_3$. From the formula for two resistances in parallel,

$$R_{eq.} = \frac{R_2R_3}{R_2 + R_3} = \frac{20 \times 8}{20 + 8} = \frac{160}{28} = 5.71 \text{ kilohms}$$

The total resistance in the circuit is then

$$R = R_1 + R_{eq.} = 5 + 5.71 \text{ kilohms} = 10.71 \text{ kilohms}$$

The current is

$$I = \frac{E}{R} = \frac{250}{10.71} = 23.4 \text{ ma.}$$

The voltage drops across $R_1$ and $R_{eq.}$ are

$$E_1 = IR_1 = 23.4 \times 5 = 117 \text{ volts}$$
$$E_2 = IR_{eq.} = 23.4 \times 5.71 = 133 \text{ volts}$$

with sufficient accuracy. These total 250 volts, thus checking the calculations so far, because the sum of the voltage drops must equal the total voltage. Since $E_2$ appears across both $R_2$ and $R_3$,

$$I_2 = \frac{E_2}{R_2} = \frac{133}{20} = 6.75 \text{ ma.}$$

$$I_3 = \frac{E_2}{R_3} = \frac{133}{8} = 16.6 \text{ ma.}$$

where $I_2$ = Current through $R_2$
$I_3$ = Current through $R_3$

The total is 23.35 ma., which checks sufficiently close with 23.4 ma., the current through the whole circuit.

There is a general rule for handling such complex circuits: Reduce the various resistances in parallel or series in *parts* of the circuit to *equivalent* resistances that then can be handled as *single* resistances in a simpler circuit. Eventually this process will lead to a simple series or parallel circuit from which the current and voltage drops can be calculated. Once these are known, Ohm's Law can be applied to each part of the circuit to determine currents and voltage drops in individual resistances.

## POWER AND ENERGY

Power — the rate of doing work — is equal to voltage multiplied by current. The unit of electrical power, called the *watt*, is equal to one volt multiplied by one ampere. The equation for power therefore is

$$P = EI$$

where $P$ = Power in watts
$E$ = E.m.f. in volts
$I$ = Current in amperes

Common fractional and multiple units for power are the *milliwatt*, one one-thousandth of a watt, and the *kilowatt*, or one thousand watts.

Example: The plate voltage on a transmitting vacuum tube is 2000 volts and the plate current is 350 milliamperes. (The current must be changed to amperes before substitution in the formula, and so is 0.35 amp.) Then

$$P = EI = 2000 \times 0.35 = 700 \text{ watts}$$

By substituting the Ohm's Law equivalents for $E$ and $I$, the following formulas are obtained for power:

$$P = \frac{E^2}{R}$$

$$P = I^2R$$

These formulas are useful in power calculations when the resistance and either the current or voltage (but not both) are known.

Example: How much power will be used up in a 4000-ohm resistor if the voltage applied to it is 200 volts? From the equation

$$P = \frac{E^2}{R} = \frac{(200)^2}{4000} = \frac{40{,}000}{4000} = 10 \text{ watts}$$

Or, suppose a current of 20 milliamperes flows through a 300-ohm resistor. Then

$$P = I^2R = (0.02)^2 \times 300 = 0.0004 \times 300 = 0.12 \text{ watt}$$

Note that the current was changed from milliamperes to amperes before substitution in the formula.

Electrical power in a resistance is turned into heat. The greater the power the more rapidly the heat is generated. We said earlier that if a resistor is to handle considerable power it must be large in size and must be constructed in such a way that the heat will be carried off rapidly by the surrounding air. This prevents the temperature of the resistor from rising to a dangerous point. Resistors for radio work are made in many sizes, the smallest being rated to "dissipate" (or carry safely) about ¼ watt. The largest resistors used in amateur equipment will dissipate about 100 watts.

However, electrical power is not always turned into heat. The power used in running a motor, for example, is converted to mechanical motion. The power supplied to a radio transmitter is largely converted into radio waves. Power applied to a loudspeaker is changed into sound waves. Nevertheless, every electrical device has some resistance, so a part of the power supplied to it is dissipated in that resistance and hence appears as heat even though the major part of the power may be converted to another form.

### *Efficiency*

In devices such as motors and vacuum tubes, the object is to obtain power in some other form than heat. Therefore power used in heating is considered to be a loss, because it is not the *useful* power. The **efficiency** of a device is the *useful* power output (in its converted form) divided by the power input to the device. In a vacuum-tube transmitter, for example, the object is to convert power from a d.c. source into a.c. power at some radio frequency. The ratio of the r.f. power output to the d.c. input is the efficiency of the tube. That is,

$$Eff. = \frac{P_o}{P_i}$$

where $Eff.$ = Efficiency (as a decimal)
$P_o$ = Power output (watts)
$P_i$ = Power input (watts)

Example: If the d.c. input to the tube is 100 watts and the r.f. power output is 60 watts, the efficiency is

$$Eff. = \frac{P_o}{P_i} = \frac{60}{100} = 0.6$$

Efficiency is usually expressed as a percentage; that is, it tells what per cent of the input power will be available as useful output. The efficiency in the above example is 60 per cent.

If a resistor is used purely for generating heat — as in an electric heater or cooker — its efficiency is practically 100 per cent, because all of the power input is converted into the desired form of power output. However, generating heat is usually not the desired end when resistors are used in radio equipment. The power losses in them are tolerated because very often a resistor performs a function that could not be conveniently or economically performed by any other device.

### *Energy*

In residences, the power company's bill is for electric **energy**, not for power. What you pay for is the *work* that electricity does for you, not the *rate* at which that work is done. Electrical work is equal to power multiplied by time; the common unit is the **watt-hour,** which means that a power of one watt has been used for one hour. That is,

$$W = PT$$

where $W$ = Energy in watt-hours
$P$ = Power in watts
$T$ = Time in hours

Other energy units are the **kilowatt-hour** and the **watt-second.** These units should be self-explanatory.

Energy units are seldom used in amateur practice, but it is obvious that a small amount of power used for a long time can eventually result in a "power" bill that is just as large as though a large amount of power had been used for a very short time.

## Capacitance and Condensers

Suppose two flat metal plates are placed close to each other (but not touching) as shown in Fig. 2-8. Normally, the plates will be electrically "neutral"; that is, the number of electrons in each plate will just balance the number of atomic nuclei and there will be no electric charge.

Now suppose that the plates are connected to a battery through a switch, as shown. At the instant the switch is closed, electrons will be attracted from the upper plate to the positive terminal of the battery, and the same number will be repelled into the lower plate from the negative battery terminal. This electron movement will continue until enough electrons move into one plate and out of the other to make the e.m.f. between them the same as the e.m.f. of the battery. (That this must be so should be fairly obvious. The plates are conductors, and when they are connected to the battery, the battery voltage must appear between them.)

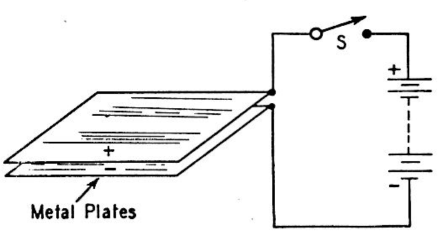



*Fig. 2-8* — A simple condenser.

If the switch is opened after the plates have been charged, the top plate is left with a deficiency of electrons and the bottom plate with an excess. In other words, the plates remain charged despite the fact that the battery no longer is connected. They remain charged because with the switch open there is nowhere for the electrons to go. However, if a wire is touched between the two plates (**short-circuiting** them) the excess electrons on the bottom plate will flow through the wire to the upper plate, thus restoring electrical neutrality to both plates. The plates have then been **discharged.**

The two plates constitute an electrical **condenser,** and from the discussion above it should be clear that a condenser possesses the property of storing electricity. It should also be clear that during the time the electrons are moving — that is, while the condenser is being charged or discharged — a *current* is flowing in the circuit even though the circuit is "broken" by the gap between the condenser plates. However, the current flows *only* during the time of charge and discharge, and this time is usually very short. There can be no *continuous* flow of direct current through a condenser.

The charge or quantity of electricity that can be placed on a condenser when a given voltage is applied depends on its **capacitance** or **capacity.** The larger the plate area and the smaller the spacing between the plates the

**TABLE 2-III**
**Dielectric Constants and Breakdown Voltages**

| *Material* | *Dielectric Constant* | *Puncture Voltage** |
|---|---|---|
| Air | 1.0 | 19.8–22.8 |
| Alsimag A196 | 5.7 | 240 |
| Bakelite (paper-base) | 3.8–5.5 | 650–750 |
| Bakelite (mica-filled) | 5–6 | 475–600 |
| Celluloid | 4–16 | |
| Cellulose acetate | 6–8 | 300–1000 |
| Fiber | 5–7.5 | 150–180 |
| Formica | 4.6–4.9 | 450 |
| Glass (window) | 7.6–8 | 200–250 |
| Glass (photographic) | 7.5 | |
| Glass (Pyrex) | 4.2–4.9 | 335 |
| Lucite | 2.5–3 | 480–500 |
| Mica | 2.5–8 | |
| Mica (clear India) | 6.4–7.5 | 600–1500 |
| Mycalex | 7.4 | 250 |
| Paper | 2.0–2.6 | 1250 |
| Polyethylene | 2.3–2.4 | 1000 |
| Polystyrene | 2.4–2.9 | 500–2500 |
| Porcelain | 6.2–7.5 | 40–100 |
| Rubber (hard) | 2–3.5 | 450 |
| Steatite (low-loss) | 4.4 | 150–315 |
| Wood (dry oak) | 2.5–6.8 | |

* In volts per mil (0.001 inch).

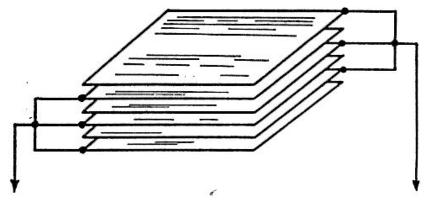


*Fig. 2-9* — **A multiple-plate condenser. Alternate plates are connected together.**

greater the capacitance. The capacitance also depends upon the kind of insulating material between the plates; it is smallest with air insulation, but substitution of other insulating materials for air may increase the capacitance of a condenser many times. The ratio of the capacitance of a condenser with some material other than air between the plates, to the capacitance of the same condenser with air insulation, is called the **specific inductive capacity** or **dielectric constant** of that particular insulating material. The material itself is called a **dielectric.** The dielectric constants of a number of materials commonly used as dielectrics in condensers are given in Table 2-III. If a sheet of photographic glass is substituted for air between the plates of a condenser, for example, the capacitance of the condenser will be increased 7.5 times.

### *Units*

The fundamental unit of capacitance is the **farad,** but this unit is much too large for practical work. Capacitance is usually measured in **microfarads** (abbreviated **$\mu$fd.**) or **micromicrofarads ($\mu\mu$fd.).** The microfarad is one-millionth of a farad, and the micromicrofarad is one-millionth of a microfarad. Condensers nearly always have more than two plates, the alternate plates being connected together to form two sets as shown in Fig. 2-9. This makes it possible to attain a fairly large capacitance in a small space as compared to a two-plate condenser, since several plates of smaller individual area can be stacked to form the equivalent of a single large plate of the same total area. Also, all plates, except the two on the ends, are exposed to plates of the other group on *both sides,* and so are twice as effective in increasing the capacitance.

The formula for calculating the capacitance of a condenser is:

$$C = 0.224 \frac{KA}{d} (n - 1)$$

where $C$ = Capacitance in $\mu\mu$fd.
$K$ = Dielectric constant of material between plates
$A$ = Area of one side of *one* plate in square inches
$d$ = Separation of plate surfaces in inches
$n$ = Number of plates

If the plates in one group do not have the same area as the plates in the other, use the area of the *smaller* plates.

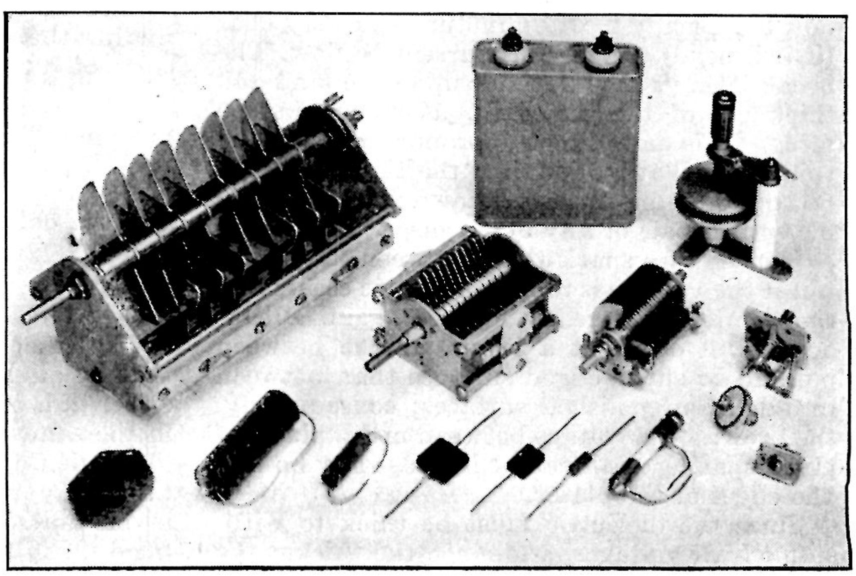
Example: A "variable" condenser has 7 semicircular plates on its rotor, the diameter of the semicircle being 2 inches. The stator has 6 rectangular plates, with a semicircular cut-out to clear the rotor shaft, but otherwise large enough to face the entire area of a rotor plate. The diameter of the cut-out is ½ inch. The distance between the adjacent surfaces of rotor and stator plates is ⅛ inch. The dielectric is air. What is the capacitance of the condenser with the plates fully meshed?

In this case, the "effective" area is the area of the rotor plate minus the area of the cut-out in the stator plate. The area of either semicircle is $\pi r^2/2$, where $r$ is the radius. The area of the rotor plate is $\pi/2$, or 1.57 square inches (the radius is 1 inch). The area of the cut-out is $\pi(\frac{1}{4})^2/2 = \pi/32 = 0.10$ square inch, approximately. The "effective" area is therefore $1.57 - 0.10 = 1.47$ square inches. The capacitance is therefore

$$C = 0.224 \frac{KA}{d} (n - 1) = 0.224 \frac{1 \times 1.47}{0.125} (13 - 1)$$
$$= 0.224 \times 11.76 \times 12 = 31.6 \ \mu\mu\text{fd.}$$

(The answer is only approximate, because of the difficulty of accurate measurement, plus a "fringing" effect at the edges of the plates that makes the actual capacitance a little higher.)

The usefulness of a condenser in electrical circuits lies in the fact that it can be charged



**Fixed and variable condensers. The bottom row includes, left to right, a high-voltage mica fixed condenser, a tubular electrolytic, tubular paper, two sizes of "postage-stamp" micas, a small ceramic type (temperature compensating), an adjustable condenser with ceramic insulation (for neutralizing in transmitters), a "button" ceramic condenser, and an adjustable "padding" condenser. Four sizes of variable condensers are shown in the second row. The two-plate condenser with the micrometer adjustment is used in transmitters. The condenser enclosed in the metal case is a high-voltage paper type used in power-supply filters.**

with electricity at one time and then discharged at a later time. In other words, it is capable of storing electrical energy that can be released later when it is needed; it is an "electrical reservoir."

### *Condensers in Radio*

The types of condensers used in radio work differ considerably in physical size, construction, and capacitance. Some representative types are shown in the photograph. In "variable" condensers (almost always constructed with air for the dielectric) one set of plates is made movable with respect to the other set so that the capacitance can be varied. "Fixed" condensers — that is, having fixed capacitance — also can be made with metal plates and with air as the dielectric, but usually are constructed from plates of metal foil with a thin solid or liquid dielectric sandwiched in between, so that a relatively large capacitance can be secured in a small unit. The solid dielectrics commonly used are mica and paper. An example of a liquid dielectric is mineral oil, but it is seldom used by itself in present-day condensers. The "electrolytic" condenser uses aluminum-foil plates with a semiliquid conducting chemical compound between them; the actual dielectric is a very thin film of insulating material that "forms" on one set of plates through electrochemical action when a d.c. voltage is applied to the condenser. The capacitance obtained with a given plate area in an electrolytic condenser is very large, compared with condensers having other dielectrics, because the film is so extremely thin — much less than any thickness that is practicable with a solid dielectric.

### *Voltage Breakdown*

When a high voltage is applied to the plates of a condenser, a considerable force is exerted on the electrons and nuclei of the dielectric. Because the dielectric is an insulator the electrons do not become detached from atoms the way they do in conductors. However, if the force is great enough the dielectric will "break down"; usually it will puncture and may char (if it is solid) and permit current to flow. The **breakdown voltage** depends upon the kind and thickness of the dielectric, as shown in the table. It is not directly proportional to the thickness; that is, doubling the thickness does not quite double the breakdown voltage. If the dielectric is air or any other gas, breakdown is evidenced by a spark or arc between the plates, but if the voltage is removed the arc ceases and the condenser is ready for use again. Breakdown will occur at a lower voltage between pointed or sharp-edged surfaces than between rounded and polished surfaces; consequently, the breakdown voltage between metal plates of given spacing in air can be increased by buffing the edges of the plates.

Since the dielectric must be thick to withstand high voltages, and since the thicker the dielectric the smaller the capacitance for a given plate area, a high-voltage condenser must have more plate area than a low-voltage condenser of the same capacitance. High-voltage high-capacitance condensers are physically large. The breakdown voltage of paper-dielectric condensers can be increased by saturating the paper with a special insulating oil and by immersing the condenser in oil. Electrolytic condensers can stand 400 to 500 volts before the dielectric film breaks down.

## CONDENSERS IN SERIES AND PARALLEL

The terms "parallel" and "series" when used with reference to condensers have the same circuit meaning as with resistances. When

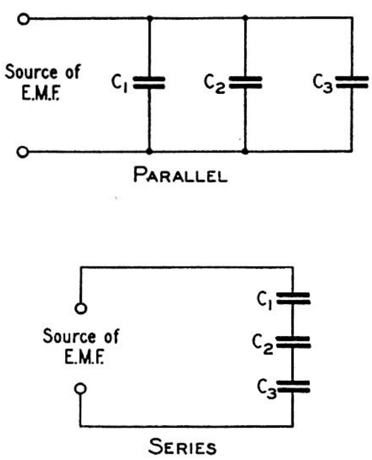



*Fig. 2-10* — Condensers in series and parallel.

a number of condensers are connected in parallel, as in Fig. 2-10, the total capacitance of the group is equal to the sum of the individual capacitances, so

$$C\text{ (total)} = C_1 + C_2 + C_3 + C_4 + \dots\dots\dots$$

However, if two or more condensers are connected in series, as in the second drawing, the total capacitance is less than that of the smallest condenser in the group. The rule for finding the capacitance of a number of series-connected condensers is the same as that for finding the resistance of a number of *parallel*-connected resistors. That is,

$$C\text{ (total)} = \cfrac{1}{\cfrac{1}{C_1} + \cfrac{1}{C_2} + \cfrac{1}{C_3} + \cfrac{1}{C_4} + \dots\dots}$$

and, for only two condensers in series,

$$C\text{ (total)} = \frac{C_1 C_2}{C_1 + C_2}$$

The same units must be used throughout; that is, all capacitances must be expressed in either $\mu$fd. or $\mu\mu$fd.; you cannot use both units in the same equation.

Condensers are connected in parallel to obtain a larger total capacitance than is available in one unit. The largest voltage that can be applied safely to a group of condensers in parallel is the voltage that can be applied safely to the condenser having the *lowest* voltage rating.

When condensers are connected in series, the applied voltage is divided up among the various condensers; the situation is much the same as when resistors are in series and there is a voltage drop across each. However, the voltage that appears across each condenser of a group connected in series is in *inverse* proportion to its capacitance, as compared with the capacitance of the whole group.

Example: Three condensers having capacitances of 1, 2 and 4 μfd., respectively, are connected in series as shown in Fig. 2-11. The total capacitance is

$$C = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}} = \frac{1}{\frac{1}{1} + \frac{1}{2} + \frac{1}{4}} = \frac{1}{\frac{7}{4}} = \frac{4}{7}$$

$$= 0.571 \ \mu\text{fd.}$$

The voltage across each condenser is proportional to the *total* capacitance divided by the capacitance of the condenser in question, so the voltage across $C_1$ is

$$E_1 = \frac{0.571}{1} \times 2000 = 1142 \text{ volts}$$

Similarly, the voltages across $C_2$ and $C_3$ are

$$E_2 = \frac{0.571}{2} \times 2000 = 571 \text{ volts}$$

$$E_3 = \frac{0.571}{4} \times 2000 = 286 \text{ volts}$$

totaling approximately 2000 volts, the applied voltage.

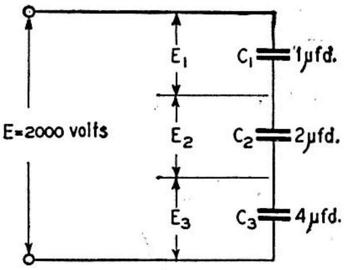



*Fig. 2-11* — An example of condensers connected in series. The solution to this arrangement is worked out in the text.

Condensers are frequently connected in series to enable the group to withstand a larger voltage (at the expense of decreased total capacitance) than any individual condenser is rated to stand. One very common application of this arrangement is in the filter circuits of high-voltage power supplies. However, as shown by the previous example, the applied voltage does not divide equally among the condensers (except when all the capacitances are the same) so care must be taken to see that the voltage rating of no condenser in the group is exceeded. It does no good, for example, to connect a condenser in series with another if the capacitance of the second is many times as great as the first; nearly all of the voltage still will appear across the condenser having the smaller capacitance.

## Inductance

It is possible to show that the flow of current through a conductor is accompanied by magnetic effects; a compass needle brought near the conductor, for example, will be deflected from its normal north-south position. The stronger the current, the more pronounced is the magnetic effect. The current, in other words, sets up a magnetic field.

If a wire conductor is formed into a coil, the same current will set up a stronger magnetic field than it will if the wire is straight. Also, if the wire is wound around an iron or steel "core" the field will be still stronger. The relationship between the strength of the field and the intensity of the current causing it is expressed by the **inductance** of the conductor or coil. If the same current flows through two coils, for example, and it is found that the magnetic field set up by one coil is twice as strong as that set up by the other, the first coil has twice as much inductance as the second. Inductance is a property of the conductor or coil and is determined by its shape and dimensions. The unit of inductance (corresponding to the ohm for resistance and the farad for capacitance) is the **henry.**

If the current through a conductor or coil is made to vary in intensity, it is found that an e.m.f. will appear across the terminals of the conductor or coil. This e.m.f. is entirely separate from the e.m.f. that is causing the current to flow. The strength of this "induced" e.m.f. becomes greater, the greater the intensity of the magnetic field and the more rapidly the current (and hence the field) is made to vary. Since the intensity of the magnetic field depends upon the inductance, the induced voltage (for a given current intensity and rate of variation) is proportional to the inductance of the conductor or coil.

The fact that an e.m.f. is "induced" accounts for the name "inductance" — or "self-inductance" as it is sometimes called. The induced e.m.f. tends to send a current through the circuit in the *opposite* direction to the current that flows because of the external e.m.f. so long as the latter current is *increasing.* However, if the current caused by the applied e.m.f. *decreases,* the induced e.m.f. tends to send current through the circuit in the *same* direction as the current from the applied e.m.f. The effect of inductance, therefore, is to oppose any *change* in the current flowing in the circuit, regardless of the nature of the change. It accomplishes this by storing energy in its magnetic field when the current in the circuit is being increased, and by releasing the stored energy when the current is being decreased. The effect is the same as the mechanical inertia that prevents an automobile from instantly coming up to speed when the accelerator pedal is pressed, and that prevents it from coming to an instant stop when the brakes are applied.

The values of inductance used in radio equipment vary over a wide range. Inductance of several henrys is required in power-supply circuits (see chapter on Power Supply) and to obtain such values of inductance it is necessary to use coils of many turns wound on iron cores. In radio-frequency circuits, the inductance values used will be measured in **millihenrys** (a millihenry is one one-thousandth of a henry) at low frequencies, and in **microhenrys** (one one-millionth of a henry) at medium frequencies and higher. Although coils for radio frequencies may be wound on special iron cores (ordinary iron is not suitable) most r.f. coils made and used by amateurs are the "air-core" type; that is, wound on an insulating form consisting of nonmagnetic material.

### *Inductance Formula*

The inductance of air-core coils may be calculated from the formula

$$L\ (\mu\text{h.}) = \frac{0.2\ a^2n^2}{3a + 9b + 10c}$$

where $L$ = Inductance in microhenrys
$a$ = Average diameter of coil in inches
$b$ = Length of winding in inches
$c$ = Radial depth of winding in inches
$n$ = Number of turns

The notation is explained in Fig. 2-12. The quantity $c$ may be neglected if the coil only has one layer of wire.

Example: Assume a coil having 35 turns of No. 30 d.s.c. wire on a form 1.5 inches in diameter. Consulting the wire table (Chapter 24), 35 turns of No. 30 d.s.c. will occupy 0.5 inch. Therefore, $a = 1.5$, $b = 0.5$, $n = 35$, and

$$L = \frac{0.2 \times (1.5)^2 \times (35)^2}{(3 \times 1.5) + (9 \times 0.5)} = 61.25\ \mu\text{h.}$$

To calculate the number of turns of a single-layer coil for a required value of inductance:

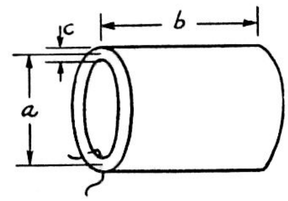



*Fig. 2-12* — Coil dimensions used in the inductance formula.

$$N = \sqrt{\frac{3a + 9b}{0.2a^2} \times L}$$

Example: Suppose an inductance of 10 microhenrys is required. The form on which the coil is to be wound has a diameter of one inch and is long enough to accommodate a coil length of 1¼ inches. Then $a = 1$, $b = 1.25$, and $L = 10$. Substituting,

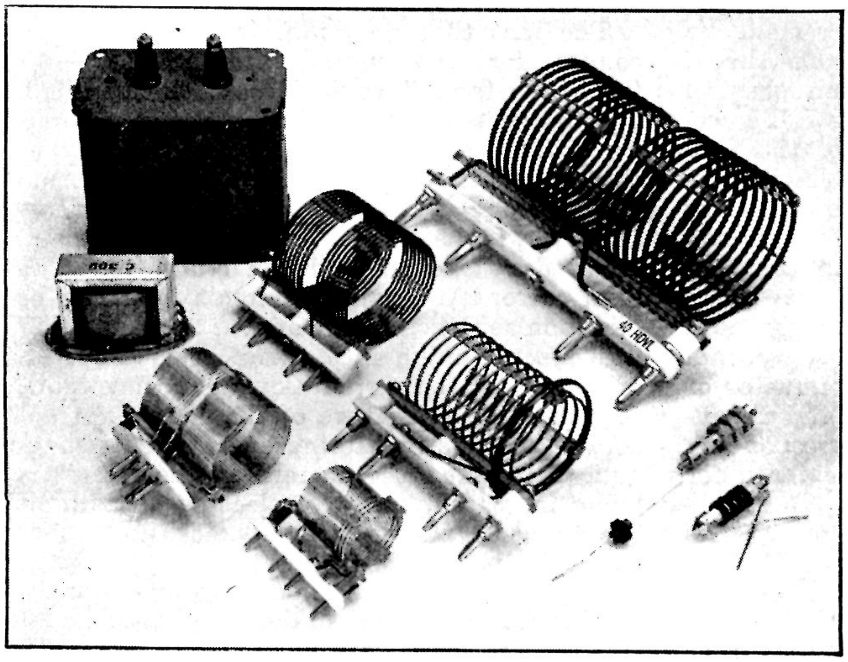
$$N = \sqrt{\frac{(3 \times 1) + (9 \times 1.25)}{0.2 \times 1^2} \times 10}$$

$$= \sqrt{\frac{14.25}{0.2} \times 10} = \sqrt{712.5}$$

$$= 26.6 \text{ turns.}$$

A 27-turn coil would be close enough to the required value of inductance, in practical work. Since the coil will be 1.25 inches long, the number of turns per inch will be 27/1.25 = 21.6. Consulting the wire table, we find that No. 18 enameled wire (or any smaller size) can be used. We obtain the proper inductance by winding the required number of turns on the form and then adjusting the spacing between the turns to make a uniformly-spaced coil 1.25 inches long.

Every conductor has inductance, even though the conductor is not formed into a coil. The inductance of a short length of straight wire is small — but it may not be negligible, because if the current through it changes its intensity rapidly enough the induced voltage may be appreciable. This will be the case in even a few inches of wire when an alternating current having a frequency of the order of 100 Mc. is flowing. However, at much lower frequencies the inductance of the same wire could be left out of any calculations because the induced voltage would be negligibly small.



Inductance coils for power and radio frequencies. The two iron-core coils at the upper left are "chokes" for power-supply filters. The three "pie"-wound coils at the lower right are used as chokes in radio-frequency circuits. The other coils are for r.f. tuned circuits ranging in power from 25 watts to a kilowatt.

## IRON-CORE COILS

We mentioned earlier that the inductance of a coil wound on an iron core is much greater than the inductance of the same coil wound on a nonmagnetic core. As a crude analogy, iron has a much lower "resistance" to the magnetic force than nonferrous materials, just as metals have much lower resistance to the flow of electric current than nonmetallic substances.

### *Permeability*

For example, suppose that the coil in Fig. 2-13 is wound on an iron core having a cross-sectional area of 2 square inches. When a certain current is sent through the coil it is found that there are 80,000 lines of force in the core. Since the area is 2 square inches, the flux density is 40,000 lines per square inch. Now suppose that the iron core is removed and the same current is maintained in the coil, and that the flux density without the iron core is found to be 50 lines per square inch. The ratio of the flux density with the given core material to the flux density (with the same coil and same current) with an air core is called the **permeability** of the material. In this case the permeability of the iron is 40,000/50 = 800. The inductance of the coil is increased 800 times by inserting the iron core, therefore.

The permeability of a magnetic material is not constant, unfortunately, but varies with the flux density. At low flux densities (or with an air core) increasing the current through the coil will cause a proportionate increase in flux. For example, if there are 2000 lines per square inch at a given current, doubling the current will increase the flux density to 4000 lines per square inch. But this cannot be carried on indefinitely; at some value of flux density, depending upon the kind of iron, it will be found that doubling the current only increases the flux density by, say, 10 per cent. At very high flux densities, increasing the current may cause no appreciable change in the flux at all. When this is so, the iron is said to be **saturated.** "Saturation" causes a rapid decrease in permeability, because it decreases the ratio of flux lines to those obtainable with the same current and an air core. Obviously, the inductance of an iron-core coil is highly dependent upon the current flowing in the coil. In an air-core coil, the inductance is independent of current because air does not "saturate."

In amateur work, iron-core coils such as the one sketched in Fig. 2-13 are used chiefly in power-supply equipment. They usually have direct current flowing through the winding, and the variation in inductance with current is usually undesirable. It may be overcome by keeping the flux density below the saturation point of the iron. This is done by cutting the core so that there is a small "air gap," as indicated by the dashed lines. The magnetic "resistance" introduced by such a gap is so large — even though the gap is only a small fraction of an inch — compared with that of the iron that the gap, rather than the iron, controls the flux density. This naturally reduces the inductance compared to what it would be without the air gap — but only for *small* currents. It actually results in a *higher* inductance when the current is large; furthermore, the inductance is practically constant regardless of the value of the current. Further information on the construction of such inductance coils will be found in the chapter on Power Supply.

### *Eddy Currents and Hysteresis*

When alternating current flows through a coil wound on an iron core the magnetic flux in the core goes through variations in intensity and direction that correspond to the variations in the alternating current. Variations in a magnetic field cause an e.m.f. to be induced, as previously explained, and since iron is a conductor a current will flow in the core. Such currents (called **eddy currents**) represent a waste of power because they flow through the resistance of the iron and thus cause heating. Eddy-current losses can be reduced by **laminating** the core; that is, by cutting it into thin strips. These strips or **laminations** must be insulated from each other by painting them with some insulating material such as varnish or shellac.

There is also another type of energy loss in an iron core: the iron tends to resist any change in its magnetic state, so a rapidly-changing current such as a.c. is forced continually to supply energy to the iron to overcome this "inertia." Losses of this sort are called **hysteresis** losses.

Eddy-current and hysteresis losses in iron increase rapidly as the frequency of the alternating current is increased. For this reason, we can use ordinary iron cores only at power and audio frequencies — up to, say, 15,000 cycles. Even so, a very good grade of iron or steel is necessary if the core is to perform well at the higher audio frequencies. Iron cores of this type are completely useless at radio frequencies.

For radio-frequency work, the losses in iron cores can be reduced to a satisfactory figure by grinding the iron into a powder and then mixing it with a "binder" of insulating material in such a way that the individual iron particles are insulated from each other. By this means cores can be made that will function satisfac-

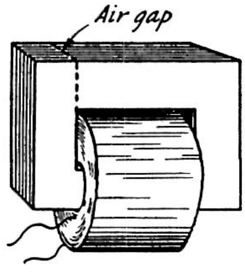



*Fig. 2-13* — **Typical construction of an iron-core coil. The small air gap prevents magnetic saturation of the iron and increases the inductance at high currents.**

torily even through the v.h.f. range — that is, at frequencies up to perhaps 100 Mc. Because a large part of the magnetic path is through a nonmagnetic material, the permeability of the iron is low compared to the values obtained at power-supply frequencies. The core is usually in the form of a "slug" or cylinder which fits inside the insulating form on which the coil is wound. Despite the fact that, with this construction, the major portion of the magnetic path for the flux is in the air surrounding the coil, the slug is quite effective in increasing the coil inductance. By pushing the slug in and out of the coil the inductance can be varied over a considerable range.

## INDUCTANCES IN SERIES AND PARALLEL

When two or more inductance coils (or **inductors,** as they are frequently called) are connected in series (Fig. 2-14, left) the total inductance is equal to the sum of the individual inductances, *provided the coils are sufficiently separated so that no coil is in the magnetic field of another.* That is,

$$L_{\text{total}} = L_1 + L_2 + L_3 + L_4 + \ldots\ldots.$$

If inductances are connected in parallel (Fig. 2-14, right), the total inductance is

$$L_{\text{total}} = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} + \frac{1}{L_4} + \ldots\ldots\ldots}$$

and for two inductances in parallel,

$$L = \frac{L_1 L_2}{L_1 + L_2}$$

Thus the rules for combining inductances in series and parallel are the same as for resistances, *if* the coils are far enough apart so that each is unaffected by another's magnetic field. When this is not so the formulas given above cannot be used.

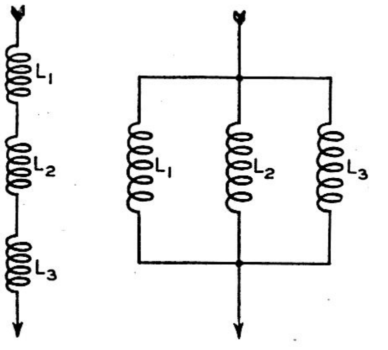



*Fig. 2-14* — Inductances in series and parallel.

In calculating the total inductance of a combination of iron-core coils to be used in a d.c. circuit, it must be remembered that the inductance of each coil may change with the amount of current that flows through it. With air-core coils there is no such change.

Although there is frequent occasion to combine resistances or capacitances in series or parallel in amateur work, there is relatively little necessity for such combinations of inductances — or rather, the cases that do arise in practice seldom require calculations.

## MUTUAL INDUCTANCE

If two coils are arranged with their axes on the same line, as shown in Fig. 2-15, a current sent through Coil 1 will cause a magnetic field which "cuts" Coil 2. Consequently, an e.m.f. will be induced in Coil 2 whenever the field strength is changing. This induced e.m.f. is similar to the e.m.f. of self-induction, but since it appears in the *second* coil because of current flowing in the *first,* it is a "mutual" effect and

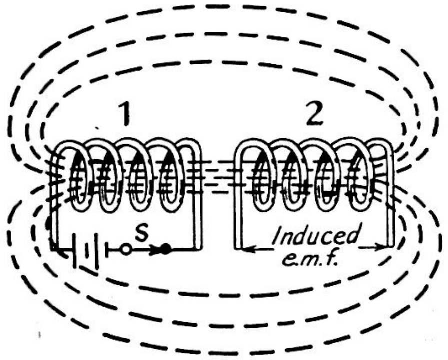



*Fig. 2-15* — **Mutual inductance. When the switch, *S*, is closed current flows through coil No. 1, setting up a magnetic field that induces an e.m.f. in the turns of coil No. 2.**

results from the **mutual inductance** between the two coils.

Mutual inductance may be large or small, depending upon the self-inductances of the coils and the proportion of the flux set up by one coil that cuts the turns of the other coil. If all the flux set up by one coil cuts all the turns of the other coil the mutual inductance has its maximum possible value. If only a small part of the flux set up by one coil cuts the turns of the other the mutual inductance is relatively small. Two coils having mutual inductance are said to be **coupled.**

The ratio of actual mutual inductance to the maximum possible value that could be obtained with two given coils is called the **coefficient of coupling** between the coils. Coils that have nearly the maximum possible mutual inductance are said to be **closely,** or **tightly,** coupled, but if the mutual inductance is relatively small the coils are said to be **loosely** coupled. The degree of coupling depends upon the physical spacing between the coils and how they are placed with respect to each other. Maximum coupling exists when they have a common axis, as shown in Fig. 2–15, and are as close together as possible. The coupling is least when the coils are far apart or are placed so their axes are at right angles.

The maximum possible coefficient of coupling is 1. This value is closely approached only when the two coils are wound on a closed iron core. The coefficient with air-core coils may run as high as 0.6 or 0.7 if one coil is wound over the other, but will be much less if the two coils are separated.

If two coils having mutual inductance are connected to the same source of current, the magnetic field of one coil can either aid or oppose the field of the other. In the former case the mutual inductance is said to be "positive"; in the latter case, "negative." Positive mutual inductance means that the total inductance is *greater* than the sum of the two individual inductances. Negative mutual inductance means that the total inductance is *less* than the sum of the two individual inductances. The mutual inductance may be made either positive or negative simply by reversing the connections to *one* of the coils.

## Time Constant

Both inductance and capacitance possess the property of storing energy — inductance stores magnetic energy and capacitance stores electrical energy. In the case of inductance, electrical energy is converted into magnetic energy when the current through the inductance is increasing, and the magnetic energy is converted back into electrical energy (and thereby restored to the circuit) when the current is decreasing. It is this alternate storing and releasing of energy that makes inductance oppose a change in the current through it. The self-induced e.m.f. is the means by which energy is put into and taken out of the magnetic field.

In the case of capacitance, energy is stored in the condenser (actually in the electric field between the plates) whenever the voltage applied to the condenser is increasing, and restored to the circuit when the applied voltage is decreasing. That is, current flows *into* the condenser in the first case, and *out of* the condenser in the second.

### *Capacitance and Resistance*

In Fig. 2-16A a battery having an e.m.f., $E$, a switch, $S$, a resistor, $R$, and condenser, $C$, are connected in series. Suppose for the moment that $R$ has zero resistance — in other words, is short-circuited — and also that there is no other resistance in the circuit. If $S$ is now closed, condenser $C$ will charge *instantly* to the battery voltage; that is, the electrons that constitute the charge redistribute themselves in a time interval so small that it can be considered to be zero. As soon as the condenser is fully charged the current flow stops completely. But since the condenser became fully charged in zero time, the current during the instantaneous charge must have been very large; mathematically, it would be *infinitely* large if the time actually was zero — this regardless of the actual number of electrons that moved. At the instant of closing the switch, therefore, the condenser can be considered to have a "resistance" of zero, a resistance that becomes an open circuit the instant the charge is complete.

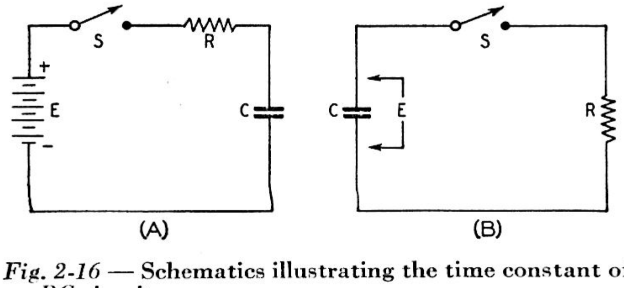



*Fig. 2-16* — Schematics illustrating the time constant of an $RC$ circuit.

If a finite value of resistance, $R$, is put into the circuit the condenser no longer can be charged instantaneously. If the condenser is initially uncharged, it will have zero "resistance" at the instant $S$ is closed, but now the amount of current that can flow is limited by $R$. The infinitely-large current required to charge the condenser in zero time cannot flow through $R$, because even with $C$ considered as a short-circuit the current in the circuit as a whole will be determined by Ohm's Law. If the battery e.m.f. is 100 volts, for example, and $R$ is 10 ohms, the maximum current that can flow with $C$ short-circuited is 10 amperes. Even this much current can flow *only* at the very instant the switch is closed. As soon as *any* current flows, condenser $C$ begins to acquire a charge, which means that the voltage across the condenser plates rises. Since the upper plate (in Fig. 2-16A) will be positive and the lower negative, the voltage on the condenser tends to send a current through the circuit in the opposite direction to the current from the battery. The voltage on the condenser, in other words, opposes the battery voltage. Immediately after the switch is closed, therefore, the current drops below its initial Ohm's Law value, and as the condenser continues to acquire charge and its potential rises, the current becomes smaller and smaller.

The length of time required to complete the charging process depends upon the capacitance of the condenser and the resistance in the circuit. More time is taken if either of these quantities is made larger. Theoretically, the charging process is never really finished, but practically the current eventually drops to a value that is smaller than anything that can be measured. The **time constant** of such a circuit is the length of time, in seconds, required for the voltage across the condenser to reach 63 per cent of the applied e.m.f. (this figure is chosen for mathematical reasons). The voltage

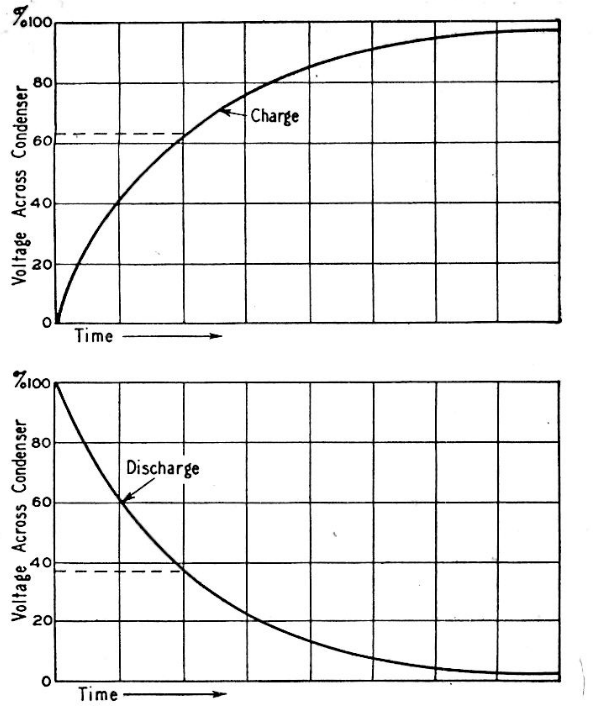



*Fig. 2-17* — **How the voltage across a condenser rises, with time, when a condenser is charged through a resistor. The lower curve shows the way in which the voltage decreases across the condenser terminals on discharging through the same resistor.**

across the condenser rises logarithmically, as shown by Fig. 2-17.

The formula for time constant is

$$T = CR$$

where $T$ = Time constant in seconds
$C$ = Capacitance in farads
$R$ = Resistance in ohms

If $C$ is in microfarads and $R$ in megohms, the time constant also is in seconds. The latter units usually are more convenient.

Example: The time constant of a 2-$\mu$fd. condenser and a 250,000-ohm resistor is

$$T = CR = 2 \times 0.25 = 0.5 \text{ second}$$

If the applied e.m.f. is 1000 volts, the voltage across the condenser plates will be 630 volts at the end of ½ second.

If a charged condenser is *discharged* through a resistor, as indicated in Fig. 2-16B, the same time constant applies. If there were no resistance, the condenser would discharge *instantly* when $S$ was closed, and for *instantaneous* discharge the current would have to be infinitely large. However, if $R$ is present the current cannot exceed the value given by Ohm's Law, where $E$ is the voltage to which the condenser is charged and $R$ is the resistance. Since $R$ limits the current flow, the condenser voltage cannot instantly go to zero, but it will decrease just as rapidly as the condenser can rid itself of its charge through $R$. When the condenser is discharging through a resistance, the time constant (calculated in the same way as above) is the time (in seconds) that it takes for the condenser to *lose* 63 per cent of its voltage; that is, for the voltage to drop to 37 per cent of its initial value.

**Example: If the condenser of the example above is charged to 1000 volts, it will discharge to 370 volts in ½ second through the 250,000-ohm resistor.**

### *Inductance and Resistance*

A comparable situation exists when resistance and inductance are in series. In Fig. 2-18, first consider $L$ to have no resistance (which would be impossible, since the conductor of which it is composed always has resistance) and also assume that $R$ is zero. Then closing $S$ would tend to send a current through the circuit. However, the instantaneous transition from no current to a finite value, however small, represents a very rapid *change* in current, and a back e.m.f. is developed by the self-inductance of $L$ that is practically equal and opposite to the applied e.m.f. The result is that the initial current is very small. However, the back e.m.f. depends upon the *change* in current and would cease to offer opposition if the current did not *continue* to increase. With no resistance in the circuit (which would lead to an infinitely-large current, by Ohm's Law) the current would increase forever, always increasing just fast enough to keep the e.m.f. of self-induction equal to the applied e.m.f. Since such a circuit never would "settle down," the time constant of an inductive circuit without resistance is infinitely long.

When resistance is in series, Ohm's Law sets a limit to the value that the current can reach. In such a circuit the current is small at first, just as in our hypothetical case without resistance. But as the current increases the voltage drop across $R$ becomes larger. The back e.m.f. generated in $L$ has only to equal the *difference* between $E$ and the drop across $R$, because that difference is the voltage actually applied to $L$. This difference becomes smaller as the current approaches the final Ohm's Law value. Theoretically, the back e.m.f. never quite disappears (that is, the current never quite reaches the Ohm's Law value) but practically it becomes unmeasurable after a time. The difference between the actual current and the Ohm's Law value also becomes undetectable. The time required for this to occur is greater the larger the value of $L$, and is shorter the larger $R$ is made. The time constant of an inductive circuit is the time in seconds required for the current to reach 63 per cent of its final value. The formula is,

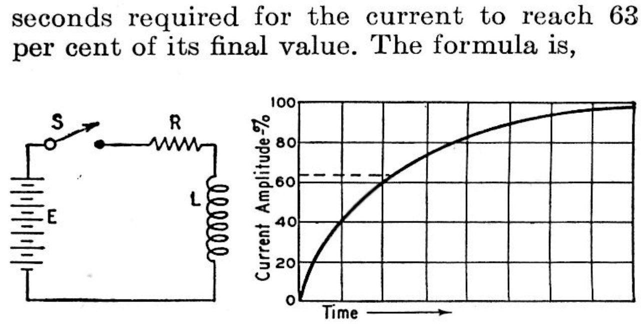



*Fig. 2-18* — **Time constant of an $LR$ circuit.**

$$T = \frac{L}{R}$$

where $T$ = Time constant in seconds
$L$ = Inductance in henrys
$R$ = Resistance in ohms

The resistance of the wire in a coil acts as though it were in series with the inductance.

Example: A coil having an inductance of 20 henrys and a resistance of 100 ohms has a time constant of

$$T = \frac{L}{R} = \frac{20}{100} = 0.2 \text{ second}$$

if there is no other resistance in the circuit. If a d.c. e.m.f. of 10 volts is applied to such a coil, the final current, by Ohm's Law, is

$$I = \frac{E}{R} = \frac{10}{100} = 0.1 \text{ amp. or } 100 \text{ ma.}$$

The current would rise from zero to 63 milliamperes in 0.2 second after closing the switch.

An inductor cannot be discharged in the same way as a condenser, because the magnetic field disappears as soon as current flow ceases. Opening $S$ does not leave the inductor "charged." The energy stored in the magnetic field instantly returns to the circuit when $S$ is opened. The rapid disappearance of the field causes a very large voltage to be induced in the coil — ordinarily many times larger than the voltage applied, because the induced voltage is proportional to the *speed* with which the field changes. The common result of opening the switch in a circuit such as the one shown is that a spark or arc forms at the switch contacts at the instant of opening. If the inductance is large and the current in the circuit is high, a great deal of energy is released in a very short period of time. It is not at all unusual for the switch contacts to burn or melt under such circumstances.

"Filter" circuits used in power-supply equipment represent an excellent example of the application of the $CR$ or $L/R$ time constant to practical work, although calculations of the type illustrated above are seldom necessary with such circuits. An understanding of the principles also is necessary in numerous special devices that are coming into widespread use in amateur stations, such as electronic keys, shaping of keying characteristics by vacuum tubes, and timing devices and control circuits. The time constants of circuits are also important in such applications as automatic gain control and noise limiters.

# Alternating Currents

## PHASE

You cannot really understand alternating currents until you have a clear picture of **phase.** Essentially it means "time," or the *time interval* between the instant when one thing occurs and the instant when a second related thing takes place. As a homely example, when a baseball pitcher throws the ball to the catcher there is a definite interval, represented by the time of flight of the ball, between the act of throwing and the act of catching. The throwing and catching are therefore "out of phase" because they do not occur at exactly the same time.

Time differences are measured in seconds, minutes, hours, and so on. In the baseball example the ball might be in the air two seconds, in which case it could be said that the throwing and catching were out of phase by two seconds. However, simply saying that two events are out of phase does not tell us which one occurred first. To give this information, the later event is said to **lag** the first in phase, while the one that occurs first is said to **lead.** Thus, throwing the ball "leads" the catch by two seconds, or the catch "lags" the throw by two seconds.

In a.c. circuits the current amplitude changes continuously, so the concept of phase or time obviously has utility whenever it becomes necessary to specify the value of the current at a particular instant. Phase can be measured in the ordinary time units, such as the second, but there is a more convenient method: since each a.c. cycle occupies exactly the same amount of time as every other cycle of the same frequency, we can use the cycle itself as the time unit. When this is done it does not matter whether one cycle lasts for a sixtieth of a second or for a millionth of a second so long as *all the cycles are the same.* In other words, using the cycle as the time unit makes the specification or measurement of phase independent of the frequency of the current, so long as only one frequency is under consideration at a time. If there are two or more frequencies, the measurement of phase has to be modified just as the measurements of two lengths must be reconciled if one is given in feet and the other in meters.

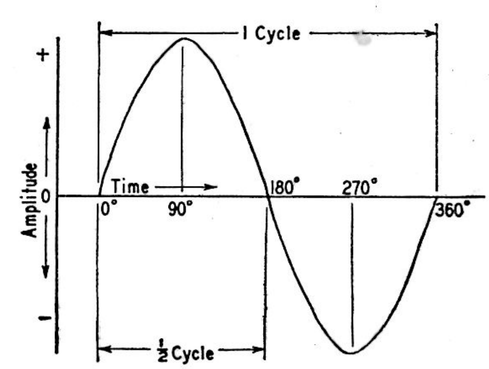



*Fig. 2-19* — An a.c. cycle is divided off into 360 degrees that are used as a measure of time or phase.

The time interval or "phase difference" under consideration usually will be less than one cycle. Phase difference could be measured in decimal parts of a cycle, but for many reasons it is more convenient to divide the cycle into 360 parts or **degrees.** A phase degree is therefore 1/360 of a cycle. (The reason for this choice of unit is this: In a sine-wave alternating current, the value of the current at any instant is proportional to the sine of the angle that corresponds to the number of degrees — that is, length of time — from the time the cycle began. There is of course no actual "angle" associated with an alternating current.) Fig. 2-19 should help make this method of measurement clear.

### *Measuring Phase*

In a steady alternating current each cycle is exactly like the preceding one. To compare the phase of two currents of the same frequency, we measure between corresponding parts of cycles of the two currents. This is shown in Fig. 2-20. The current labeled *A* leads the one marked *B* by 45 degrees, since *A*'s cycles begin 45 degrees sooner in time. (It is equally correct to say that *B lags* *A* by 45 degrees.) The amplitudes of the individual currents do not affect their relative phases — current *B* is shown as having smaller amplitude than *A*. Regardless of the amplitudes, the lagging current always would begin its cycle (the start of the cycle is considered to be the point at which it is passing through zero and starting to increase in the positive direction) the same number of degrees after the current that leads begins *its* cycle.

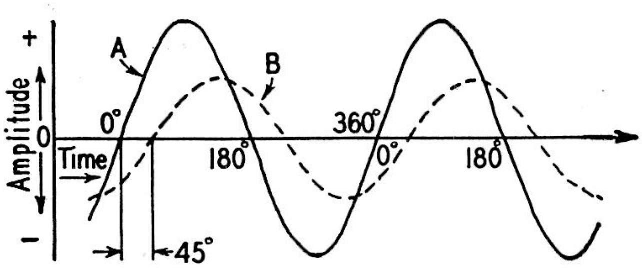



*Fig. 2-20* — **When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.**

Two important special cases are shown in Fig. 2-21. In the upper drawing *B* lags 90 degrees behind *A;* that is, its cycle begins just one-quarter cycle later than that of *A*. When one wave is passing through zero, the other is just at its maximum point. Note that (using *A* as a reference) in the first quarter cycle *A* is positive and *B* is negative; in the second quarter cycle both *A* and *B* are positive, but one is decreasing while the other is increasing; in the third quarter cycle *A* is negative while *B* is positive; and in the last quarter cycle both are negative.

In the lower drawing *A* and *B* are 180 degrees out of phase. In this case it does not matter which one we consider to lead or lag. *B* is always positive while *A* is negative, and vice versa. The two waves are thus *completely* out of phase.

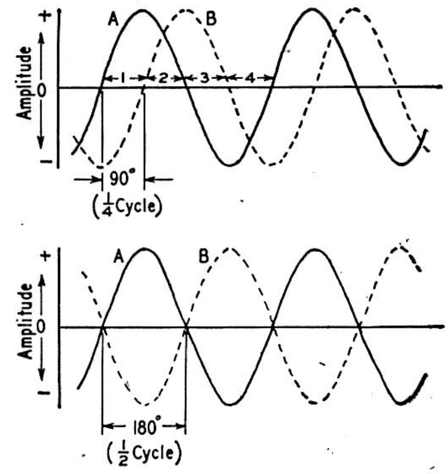



*Fig. 2-21* — **Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.**

The waves shown in Figs. 2-20 and 2-21 could represent current, voltage, or both. *A* and *B* might be two currents in separate circuits, or *A* might represent voltage while *B* represented current in the same circuit. If *A* and *B* represent two currents in the *same* circuit (or two voltages in the same circuit) the *actual* current (or voltage) would take a *single* value at any instant. This value would equal the sum of the two at that instant. (We must take into account the fact that the sum of positive and negative values is actually equal to the *difference* between them.) The resultant current (or voltage) also is a sine wave, because adding any number of sine waves of the same frequency always results in a sine wave also of the same frequency.

## ● REACTANCE

The discussion of capacitance and inductance earlier in this chapter was confined to cases where only d.c. voltages were applied. To understand what happens in a condenser or inductance when an *a.c.* voltage is applied, it is necessary to become acquainted with a fundamental *definition* of electric current (as contrasted to the physical *description* of current given earlier). By definition, the amplitude of an electric current is the *rate* at which electric charge is moved past a point in a circuit. If a large quantity of charge moves past the observing point in a given time, the current is large; if the quantity is small in the same amount of time, the current is small.

### *Alternating Current in Condensers*

The quantity of charge that can be placed on a condenser of given capacitance is proportional to the voltage applied to the condenser. As we explained earlier, the condenser becomes charged *instantly* if there is no resistance in the circuit. Suppose a sine-wave a.c. voltage is applied to a condenser in a circuit containing no resistance, as indicated in Fig. 2-22. For convenience, the first half-cycle of the applied voltage is divided into eight equal time intervals. In the period *OA*, the voltage increases from zero to 38 volts; at the end of this period the condenser is charged to that voltage. In the next interval the voltage increases to 71 volts; that is, 33 volts additional. In this second interval a *smaller* quantity of charge has been added than in the first interval, because the voltage rise during the second interval was smaller. Consequently the average current during the second interval is smaller than during the first. In the third interval, *BC*, the voltage rises from 71 to 92 volts, an increase of 21 volts. This is less than the voltage increase during the second interval, so the quantity of electricity added to the charge during the third interval is less than the quantity added during the second. In other words, the average current during the third interval is still smaller. In the fourth interval, *CD*, the voltage increases only 8 volts; the charge added is smaller than in any preceding interval and therefore the current also is smaller. By dividing the first quarter cycle into a very large number of intervals it could be shown that the current charging the condenser has the shape of a sine wave, just as the applied voltage does. But the current is largest at the beginning of the cycle and becomes zero at the maximum value of the voltage (the condenser cannot be charged to a higher voltage than the maximum applied, so no further current can flow) so there is a phase difference of 90 degrees between the voltage and current. During the first quarter cycle of the applied voltage the current is flowing in the normal way through the circuit, since the condenser is being charged. Hence the current is positive during this first quarter cycle, as indicated by the dashed line in Fig. 2-22.

In the second quarter cycle — that is, in the time from *D* to *H*, the voltage applied to the condenser decreases. During this time the condenser *loses* the charge it acquired during the first quarter cycle. Applying the same reasoning, it is plain that the current is small from *D* to *E* and continues to increase during each succeeding interval. However, the current is flowing *against* the applied voltage because the condenser is *discharging into the circuit.* Hence the current is *negative* during this quarter cycle.

The third and fourth quarter cycles repeat the events of the first and second, respectively, with this difference — the polarity of the applied voltage has reversed, and the current changes to correspond. In other words, *an alternating current flows through a condenser when an a.c. voltage is applied to it.* As shown by Fig. 2-22, the current starts its cycle 90 degrees before the voltage, so the current in a condenser *leads the applied voltage by 90 degrees.*

### *Capacitive Reactance*

Remembering the definition of current as given at the beginning of this section, as well as the mechanism of current flow described above, it should be plain that the more rapid the voltage rise the larger the current, because a rapid change in voltage means a rapid transfer of charge into or out of the condenser. The rapidity with which the voltage changes depends upon two things: (1) the amplitude of the voltage (the greater the maximum value, the faster the voltage must rise from zero to reach that maximum in the time of one-quarter cycle if the frequency is fixed); (2) the frequency (the higher the frequency, the more rapidly the voltage goes through its changes in a given time if the maximum amplitude is fixed). Also, the amplitude of the current depends upon the capacitance of the condenser, because the larger the capacitance the greater the amount of charge transferred during a given change in voltage.

The fact that the current flowing through a condenser is directly proportional to the applied a.c. voltage is extremely important. It is exactly what Ohm's Law says about the flow of direct current in a resistive circuit, and so leads us to the conclusion that Ohm's Law may be applied to an alternating-current circuit containing a condenser. Of course, a condenser does not offer "resistance" to the flow of alternating current, because the condenser does not consume power as a resistor does. It merely stores energy in one part of the cycle and returns it to the circuit in the next part. Furthermore, the larger the capacitance the larger the current; this is just the opposite of what we expect with resistance. And finally, the "opposition" offered by a condenser to alternating current depends on the *frequency* of that current. But with a given capacitance and a given frequency, the condenser follows Ohm's Law on a.c.

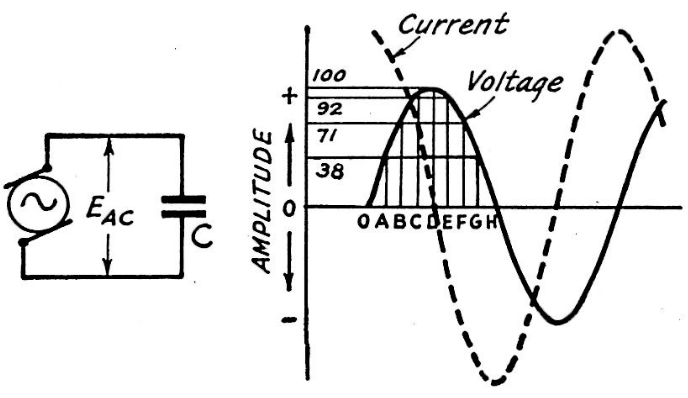



*Fig. 2-22* — Voltage and current phase relationships when an alternating voltage is applied to a condenser.

Since the opposition effect of a condenser is not resistance, it is called by another name, **reactance.** But because reactance holds back current flow in a similar fashion to resistance, the unit of reactance also is the ohm. The reactance of a condenser is

$$X_C = \frac{1}{2\pi f C}$$

where $X_C$ = Condenser reactance in ohms
$f$ = Frequency in cycles per second
$C$ = Capacitance in farads
$\pi$ = 3.14

The fundamental units (cycles per second, farads) are too large for practical use in radio circuits. However, if the capacitance is in microfarads and the frequency is in megacycles, the reactance will come out in ohms in the formula.

Example: The reactance of a condenser of 470 $\mu\mu$fd. (0.00047 $\mu$fd.) at a frequency of 7150 kc. (7.15 Mc.) is

$$X = \frac{1}{2\pi f C} = \frac{1}{6.28 \times 7.15 \times 0.00047} = 47.4 \text{ ohms}$$

### *Inductive Reactance*

In the case of an alternating voltage applied to a circuit containing only inductance, with no resistance, it must be remembered that in such a resistanceless circuit the current always changes just rapidly enough to induce a back e.m.f. that equals and opposes the applied voltage. In Fig. 2-23, the cycle is again divided off into equal intervals. Assuming that the current has a maximum value of 1 ampere, the instantaneous current at the end of each interval will be as shown. The value of the *induced* voltage is proportional to the *rate at which the current changes.* It is therefore greatest in the intervals $OA$ and $GH$ and least in the intervals $CD$ and $DE$. The induced voltage actually is a sine wave (if the current is a sine wave) as shown by the dashed curve. The *applied* voltage, because it is always equal to and opposed by the induced voltage, is equal to and 180 degrees out of phase with the induced voltage, as shown by the second dashed curve. The result, therefore, is that the current flowing in an inductance is 90 degrees out of phase with the applied voltage, and lags behind the applied voltage. This is just the opposite of the condenser case.

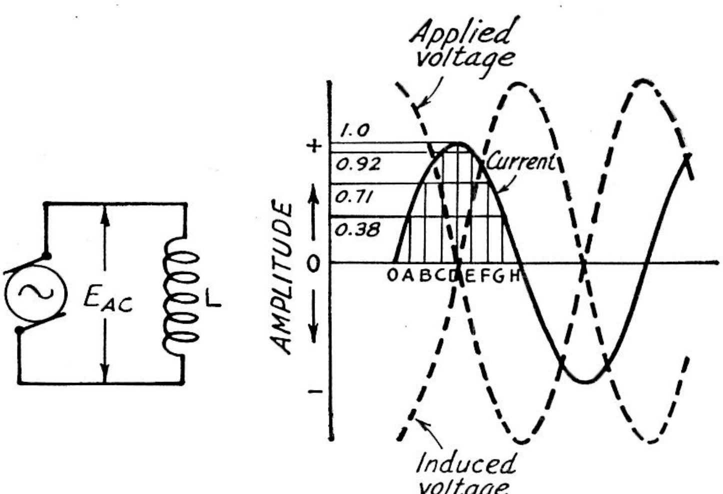



*Fig. 2-23* — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

Just enough current will flow in an inductance to induce an e.m.f. that just equals the applied e.m.f. Since the value of the induced e.m.f. is proportional to the rate at which the current changes, and this rate of change is in turn proportional to the frequency of the current, it should be clear that a small current changing rapidly (that is, at a high frequency) can generate a large back e.m.f. in a given inductance just as well as a large current changing slowly (low frequency). Consequently, the current that flows through a given inductance will decrease as the frequency is raised, if the applied e.m.f. is held constant. However, with both frequency and inductance fixed, the current will be larger when the applied voltage is increased, because the necessary rate of change in the current to induce the required back e.m.f. can only be obtained by having a greater total current flow under such circumstances. Again, when the applied voltage and frequency are fixed, the value of current required is less, as the inductance is made larger, because the induced e.m.f. also is proportional to inductance.

Just as in the capacitance case, the key point here is that — with the frequency and inductance fixed — an increase in the applied a.c. voltage causes a proportionate increase in the current. This is Ohm's Law again — and, again, the opposition effect is similar to, but not identical to, resistance. It is called **inductive reactance** and, like capacitive reactance, is measured in ohms. There is no energy loss in inductive reactance; the energy is stored in the magnetic field in one quarter cycle and then returned to the circuit in the next.

The formula for inductive reactance is

$$X_L = 2\pi f L$$

where $X_L$ = Inductive reactance in ohms
$f$ = Frequency in cycles per second
$L$ = Inductance in henrys
$\pi$ = 3.14

Example: The reactance of a coil having an inductance of 8 henrys, at a frequency of 120 cycles, is

$$X_L = 2\pi f L = 6.28 \times 120 \times 8 = 6029 \text{ ohms}$$

In radio-frequency circuits the inductance values usually are small and the frequencies are large. If the inductance is expressed in millihenrys and the frequency in kilocycles, the conversion factors for the two units cancel, and the formula for reactance may be used without first converting to fundamental units. Similarly, no conversion is necessary if the inductance is in microhenrys and the frequency is in megacycles.

Example: The reactance of a 15-microhenry coil at a frequency of 14 Mc. is

$$X_L = 2\pi fL = 6.28 \times 14 \times 15 = 1319 \text{ ohms}$$

### *Ohm's Law for Reactance*

Ohm's Law for an a.c. circuit containing *only* reactance is

$$I = \frac{E}{X}$$

$$E = IX$$

$$X = \frac{E}{I}$$

where $E$ = E.m.f. in volts
$I$ = Current in amperes
$X$ = Reactance in ohms

The reactance may be either inductive or capacitive.

Example: If a current of 2 amperes is flowing through the condenser of the previous example (reactance = 47.4 ohms) at 7150 kc., the voltage drop across the condenser is

$$E = IX = 2 \times 47.4 = 94.8 \text{ volts}$$

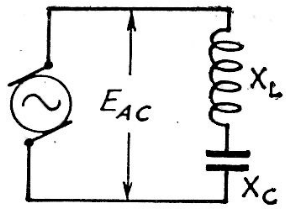
If 400 volts at 120 cycles is applied to the 8-henry inductance of the previous example, the current through the coil will be

$$I = \frac{E}{X} = \frac{400}{6029} = 0.0663 \text{ amp. (66.3 ma.)}$$

These examples show that there is nothing complicated about using Ohm's Law for a reactive a.c. circuit. The question naturally arises, though, as to what to do when the circuit consists of an inductance in series with a capacitance. In such a case the same current flows through both reactances. However, the voltage across the coil *leads* the current by 90 degrees, and the voltage across the condenser *lags* behind the current by 90 degrees. The coil and condenser voltages therefore are 180 degrees out of phase.

A simple circuit of this type is shown in Fig. 2-24. The same figure also shows the current (heavy line) and the voltage drops across the inductance ($E_L$) and capacitance ($E_C$). It is assumed that $X_L$ is larger than $X_C$ and so has a larger voltage drop. Since the two voltages are completely out of phase the *total* voltage ($E_{AC}$) is equal to the *difference* between them. This is shown in the drawing as $E_L - E_C$. Notice that, because $E_L$ is larger than $E_C$, the resultant voltage is exactly in phase with $E_L$. In other words, the circuit as a whole simply acts *as though it were an inductance* — an inductance of smaller value than the actual inductance present, since the effect of the actual inductive reactance is reduced by the capacitive reactance in series with it. If $X_C$ is larger than $X_L$, the arrangement will behave like a capacitance — again of smaller reactance than the actual capacitive reactance present in the circuit.

The "equivalent" or total reactance of any circuit containing inductive and capacitive reactances in series is equal to $X_L - X_C$. If



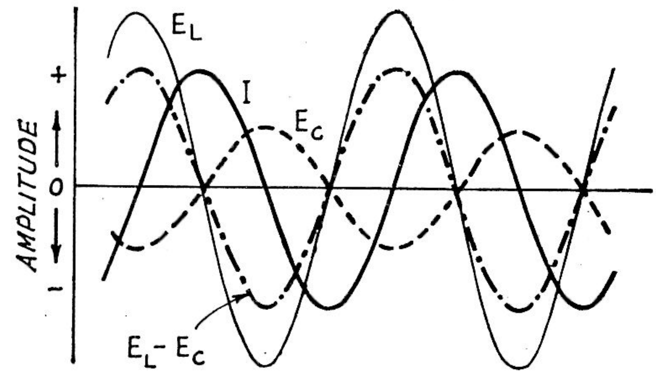




*Fig. 2-24* — **Current and voltages in a circuit having inductive and capacitive reactances in series.**

there are several coils and condensers in series, we simply add up all the inductive reactances, then add up all the capacitive reactances, and then subtract the latter from the former. It is customary to call inductive reactance "positive" and capacitive reactance "negative." If the equivalent or net reactance is positive, the voltage leads the current by 90 degrees; if the net reactance is negative, the voltage lags the current by 90 degrees.

### *Reactive Power*

A curious feature of the drawing in Fig. 2-24 is that the voltage drop across the coil is larger than the voltage applied to the circuit. At first glance this might seem to be an impossible condition. But it is not; the reason is that neither the coil nor condenser *consumes* power. Actually, when energy is being stored in the coil's magnetic field, energy is being returned to the circuit from the condenser's electric field, and vice versa. This stored energy is responsible for the fact that the voltages across reactances in series can be larger than the voltage applied to them.

It will be recalled that in a resistance the flow of current causes heating and a power loss equal to $I^2R$. The power in a reactance is equal to $I^2X$, but is not a "loss"; it is simply power that is transferred back and forth between the field and the circuit but not used up in heating anything. In the quarter cycle when the current and voltage in a reactance both have the same polarity, energy is stored in the field; in the quarter cycle when the current and voltage have *opposite* polarity the energy is returned to the circuit. To distinguish this "nondissipated" power from the power which is actually consumed, the unit of reactive power is called the **volt-ampere** instead of the watt. Reactive power is sometimes called "wattless" power.

## IMPEDANCE

Although resistance, inductive reactance and capacitive reactance all are measured in ohms, the fact that they all are measured by the same unit does not indicate that they can be combined indiscriminately. Reactance does not absorb energy; resistance does. Voltage and current are in phase in resistance, but differ in phase by a quarter cycle in reactance. Furthermore, in inductive reactance the voltage leads the current, while in capacitive reactance the current leads the voltage. All these things must be taken into account when reactance and resistance are combined together in a circuit.

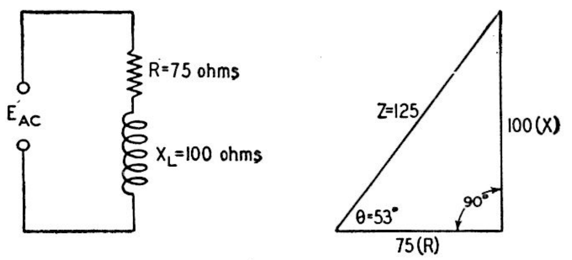



*Fig. 2-25* — **Resistance and inductive reactance connected in series.**

In the simple circuit shown in Fig. 2-25, for example, it is not possible simply to add the resistance and reactance together to obtain a quantity that will indicate the opposition offered by the combination to the flow of current. Inasmuch as both resistance and reactance are present, the total effect can obviously be neither wholly one nor the other. In circuits containing *both* reactance and resistance the opposition effect is called **impedance.** The unit of impedance is also the ohm.

If the inductance in Fig. 2-25 were short-circuited, only the resistance would remain and the circuit would simply have a resistance of 75 ohms. In such a case the current and voltage would be in phase. On the other hand, if the resistance were short-circuited the circuit simply would have a reactance of 100 ohms, and the current would lag behind the voltage by one-quarter cycle or 90 degrees. When both are in the circuit, it would be expected that the impedance would be greater than either the resistance or reactance. It might also be expected that the current would be neither in phase with the voltage nor lagging 90 degrees behind it, but would be somewhere between the complete in-phase and the 90-degree phase conditions. Both things are true. The larger the reactance compared with the resistance, the more nearly the phase angle approaches 90 degrees; the larger the resistance compared to the reactance, the more nearly the current approaches the condition of being in phase with the voltage.

It can be shown that resistance and reactance can be combined in the same way that a right-angled triangle is constructed, if the resistance is laid off to proper scale as the base of the triangle and the reactance is laid off as the altitude to the same scale. This is also indicated in Fig. 2-25. When this is done the hypotenuse of the triangle represents the impedance of the circuit, to the same scale, and the angle between $Z$ and $R$ (usually called $\theta$ and so indicated in the drawing) is equal to the phase angle between the applied e.m.f. and the current. It is unnecessary, of course, actually to draw such a triangle when impedance is to be calculated; by geometry,

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

In the case shown in the drawing,

$$Z = \sqrt{(75)^2 + (100)^2} = \sqrt{15{,}625} = 125 \text{ ohms.}$$

The phase angle can be found from simple trigonometry. Its tangent is equal to $X/R$; in this case $X/R = 100/75 = 1.33$. From trigonometric tables it can be determined that the angle having a tangent equal to 1.33 is approximately 53 degrees. Fortunately, in ordinary amateur work it is seldom necessary to give much consideration to the phase angle because in most practical cases the angle will either be nearly zero (current and voltage in phase) or close to 90 degrees (current and voltage approximately a quarter cycle out of phase).

A circuit containing resistance and capacitance in series (Fig. 2-26) can be treated in the same way. That is, the impedance is

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

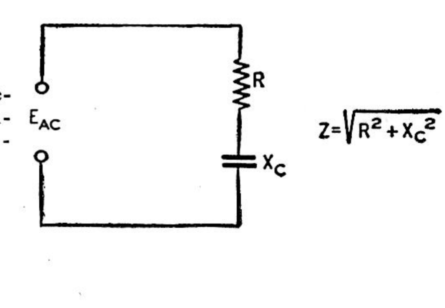
and the phase angle again is the angle whose tangent is equal to $X/R$. It must be remembered, however, that in this case the current *leads* the applied e.m.f., while in the resistance-inductance case it *lags* behind the voltage.

In neither case is the impedance of the circuit equal to the simple arithmetical sum of the resistance and reactance. With $R = 75$ ohms and $X_L = 100$ ohms, simple addition would give 175 ohms while the actual impedance is 125 ohms. However, if either $X$ or $R$ is very small compared to the other (say, 1/10 or less) the impedance is very nearly equal to the larger of the two quantities. For example, if $R = 1$ ohm and $X = 10$ ohms,

$$Z = \sqrt{R^2 + X^2} = \sqrt{(1)^2 + (10)^2}$$
$$= \sqrt{101} = 10.05 \text{ ohms.}$$

Hence if either $X$ or $R$ is at least 10 times as large as the other, the error in assuming that the impedance is equal to the larger of the two will not exceed ½ of 1 per cent, which is

*Fig. 2-26* — **Resistance and capacitive reactance in series.**



usually negligible. This fact is frequently useful.

In working with impedance, remember that one of its components is reactance and that the reactance of a given coil or condenser changes with the applied frequency. Therefore, impedance also changes with frequency. The change in impedance as the frequency is changed may be very slow if the resistance is considerably larger than the reactance. However, if the impedance is mostly reactance a change in frequency will cause the impedance to change practically as rapidly as the reactance itself changes.

### *Ohm's Law for Impedance*

Since impedance is made up of resistance and reactance, Ohm's Law can be applied to circuits containing impedance just as readily as to circuits having resistance or reactance only. The formulas are

$$I = \frac{E}{Z}$$

$$E = IZ$$

$$Z = \frac{E}{I}$$

where $E$ = E.m.f. in volts
$I$ = Current in amperes
$Z$ = Impedance in ohms

Example: Assume that the e.m.f. applied to the circuit of Fig. 2-25 is 250 volts. Then

$$I = \frac{E}{Z} = \frac{250}{125} = 2 \text{ amperes.}$$

The same current is flowing in both $R$ and $X_L$, and Ohm's Law as applied to either of these quantities says that the voltage drop across $R$ should equal $IR$ and the voltage drop across $X_L$ should equal $IX_L$. Substituting,

$$E_R = IR = 2 \times 75 = 150 \text{ volts}$$
$$E_{X_L} = IX_L = 2 \times 100 = 200 \text{ volts}$$

The arithmetical sum of these voltages is greater than the applied voltage. However, the *actual*

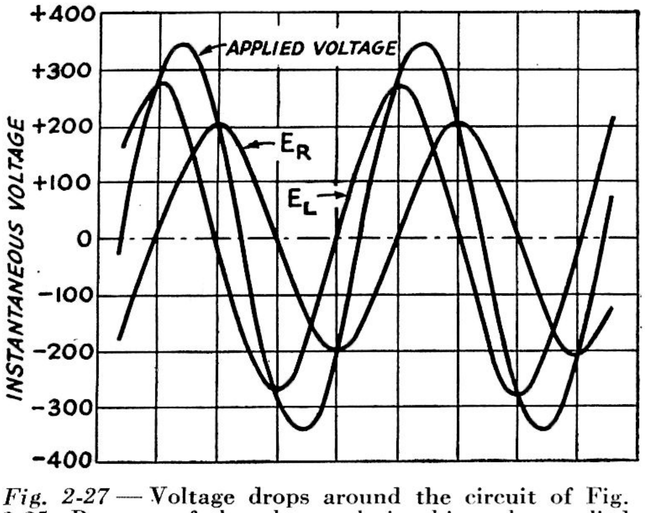



*Fig. 2-27* — Voltage drops around the circuit of Fig. 2-25. Because of the phase relationships, the applied voltage is less than the arithmetical sum of the drops across the resistor and inductor.

sum of the two when the phase relationship is taken into account is equal to 250 volts r.m.s., as shown by Fig. 2-27, where the instantaneous values are added throughout the cycle. Whenever resistance and reactance are in series, the individual voltage drops always add up, arithmetically, to more than the applied voltage. There is nothing fictitious about these voltage drops; they can be measured readily by suitable instruments. It is simply an illustration of the importance of phase in a.c. circuits.

A more complex series circuit, containing resistance, inductive reactance and capacitive reactance, is shown in Fig. 2-28. In this case it is necessary to take into account the fact that the phase angles between current and voltage differ in all three elements. Since it is a series circuit, the current is the same throughout. Considering first just the inductance and

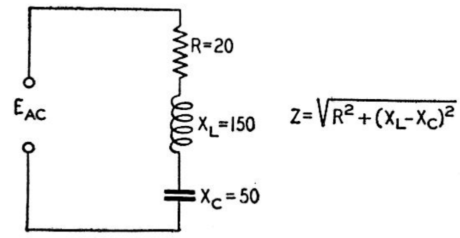



*Fig. 2-28* — Resistance, inductive reactance, and capacitive reactance in series.

capacitance and neglecting the resistance, the phase relationships are the same as in Fig. 2-24. The net reactance in Fig. 2-28 is

$$X_L - X_C = 150 - 50 = 100 \text{ ohms (inductive)}$$

Since the series reactances can be lumped into one equivalent reactance, it is easy to find the impedance of the circuit by the rules previously given. The impedance of a circuit containing resistance, inductance and capacitance in series is

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

Example: In the circuit of Fig. 2-28, the impedance is

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$
$$= \sqrt{(20)^2 + (150 - 50)^2} = \sqrt{(20)^2 + (100)^2}$$
$$= \sqrt{10{,}400} = 102 \text{ ohms}$$

The phase angle can be found from $X/R$, where $X = X_L - X_C$.

### *Parallel Circuits*

Suppose that a resistor, condenser and coil are connected in parallel as shown in Fig. 2-29 and an a.c. voltage is applied to the combination. In any one branch, the current will be unchanged if one or both of the other two branches is disconnected, so long as the applied voltage remains unchanged. For example, $I_L$, the current through the inductance, will not change if both $R$ and $C$ are removed (although the total current, $I$, *will* change). Thus the current in each branch can be calculated quite simply by the Ohm's Law

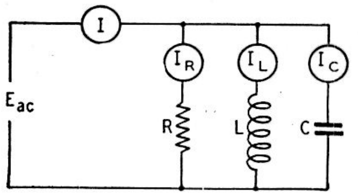



*Fig. 2-29* — **Resistance, inductance and capacitance in parallel. Instruments connected as shown will read the total current, *I*, and the individual currents in the three branches of the circuit.**

formulas given in the preceding sections, if the voltage and reactance or resistance are known. The total current, $I$, is the sum of the currents through all three branches — not the arithmetical sum, but the sum when phase is taken into account.

The currents through the various branches will be as shown in Fig. 2-30, assuming for purposes of illustration that $X_L$ is smaller than $X_C$ and that $X_C$ is smaller than $R$, thus making $I_L$ larger than $I_C$, and $I_C$ larger than $I_R$. The current through $C$ leads the voltage by 90 degrees and the current through $L$ lags the voltage by 90 degrees, so these two currents are 180 degrees out of phase. As shown at E, the total reactive current is the difference between $I_C$ and $I_L$. This resultant current lags the voltage by 90 degrees, because $I_L$ is larger than $I_C$. When the reactive current is added to $I_R$, the total current, $I$, is as shown at F. It can be seen that $I$ lags the applied voltage by an angle smaller than 90 degrees and that the total current, while less than the simple sum (neglecting phase) of the three branch currents, is larger than the current through $R$ alone.

The impedance looking into the parallel circuit from the source of voltage is equal to the applied voltage divided by the total or "line" current, $I$. In the case illustrated, $I$ is greater than $I_R$, so the impedance of the circuit is less than the resistance of $R$. How much less depends upon the net reactive current flowing through $L$ and $C$ in parallel. If $X_L$ and $X_C$ are very nearly equal the net reactive current will be quite small because it is equal to the *difference* between two nearly equal currents. In such a case the impedance of the circuit will be almost the same as the resistance of $R$ alone. On the other hand, if $X_L$ and $X_C$ are quite different the net reactive current can be relatively large and the total current also will be appreciably larger than $I_R$. In such a case the circuit impedance will be lower than the resistance of $R$ alone.

The calculation of the impedance of parallel circuits is somewhat complicated. Fortunately, calculations are not necessary in most amateur work except in a special — and simple — case treated in a later section of this chapter.

### *Power Factor*

In the circuit of Fig. 2-25 an applied e.m.f. of 250 volts results in a current of 2 amperes. If the circuit were purely resistive (containing no reactance) this would mean a power dissipation of 250 × 2 = 500 watts. However, the circuit actually consists of resistance and reactance, and only the resistance consumes power. The power in the resistance is

$$P = I^2R = (2)^2 \times 75 = 300 \text{ watts}$$

This is the actual power consumed by the circuit as compared to the apparent power input of 500 watts. The ratio of the power consumed to the apparent power is called the **power factor** of the circuit, and in the case used as an example would be 300/500 = 0.6. Power factor is frequently expressed as a percentage; in this case, the power factor would be 60 per cent.

"Real" or dissipated power is measured in watts; apparent power, to distinguish it from real power, is measured in volt-amperes (just like the "wattless" power in a reactance). It is simply the product of volts and amperes and has no direct relationship to the power actually used up or dissipated unless the power factor of the circuit is known. The power factor of a purely resistive circuit is 100 per cent or 1, while the power factor of a pure reactance is zero. In this illustration, the reactive power is

$$VA \text{ (volt-amperes)} = I^2X = (2)^2 \times 100 = 400 \text{ volt-amperes.}$$

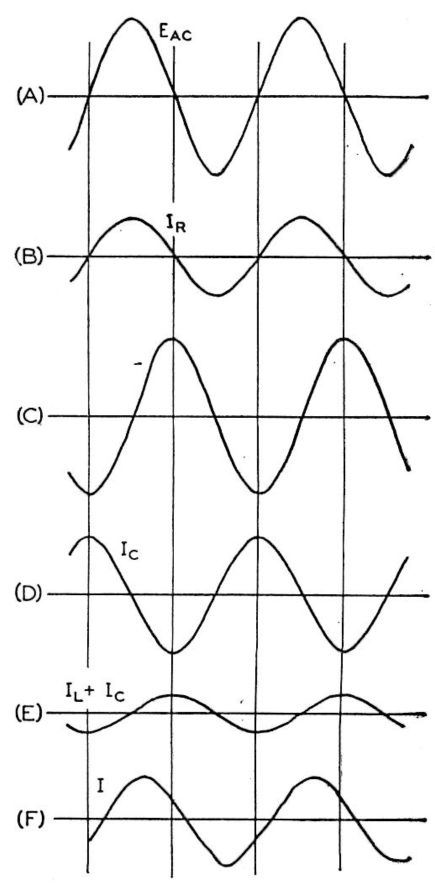



*Fig. 2-30* — **Phase relationships between branch currents and applied voltage for the circuit of Fig. 2-29. The total current through *L* and *C* in parallel ($I_L + I_C$) and the total current in the entire circuit (*I*) also are shown.**

### *Complex Waves*

It was pointed out early in this chapter that a complex wave (a "nonsinusoidal" wave) can be resolved into a fundamental frequency and a series of harmonic frequencies. When such a complex voltage wave is applied to a circuit containing reactance, the current through the circuit will not have the same waveshape as the applied voltage. This is because the reactance of a coil and condenser depend upon the applied frequency. For the second-harmonic component of a complex wave, the reactance of the coil is twice and the reactance of the condenser one-half their values at the fundamental frequency; for the third harmonic the coil reactance is three times and the condenser reactance one-third, and so on.

Just what happens to the current waveshape depends upon the values of resistance and reactance involved and how the circuit is arranged. In a simple circuit with resistance and inductive reactance in series, the amplitudes of the harmonics will be reduced because the inductive reactance increases in proportion to frequency. When a condenser and resistance are in series, on the other hand, the harmonics are likely to be accentuated because the condenser reactance becomes lower as the frequency is raised. When both inductive and capacitive reactance are present the shape of the current wave can be altered in a variety of ways, depending upon the circuit and the "constants," or values of $L$, $C$ and $R$, selected.

This property of nonuniform behavior with respect to fundamental and harmonics is an extremely useful one. It is the basis of "filtering," or the suppression of undesired frequencies in favor of a single desired frequency or group of such frequencies.

## Transformers

It has been shown in the preceding sections that, when an alternating voltage is applied to an inductance, an e.m.f. is induced by the varying magnetic field accompanying the flow of alternating current. If a second coil is brought into the same field, a similar e.m.f. likewise will be induced in this coil. This induced e.m.f. may be used to force a current through a wire, resistance or other electrical device connected to the terminals of the second coil.

Two coils operating in this way are said to be **coupled,** and the pair of coils constitutes a **transformer.** The coil connected to the source of energy is called the **primary** coil, and the other is called the **secondary** coil.

### *Types of Transformers*

The usefulness of the transformer lies in the fact that electrical energy can be transferred from one circuit to another without direct connection, and in the process can be readily changed from one voltage level to another. Thus, if a device to be operated requires, for example, 115 volts and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. The transformer, of course, can be used only on a.c., since no voltage will be induced in the secondary if the magnetic field is not changing. If d.c. is applied to the primary of a transformer, a voltage will be induced in the secondary only at the instant of closing or opening the primary circuit, since it is only at these times that the field is changing.

As shown in Fig. 2-31, the primary and secondary coils of a transformer may be wound on a core of magnetic material. This increases the inductance of the coils so that a relatively small number of turns may be used to induce a given value of voltage with a small current A closed core (one having a continuous magnetic path) such as that shown in Fig. 2-31 also tends to insure that practically all of the field set up by the current in the primary coil will cut the turns of the secondary coil. However, the core introduces a power loss because of hysteresis and eddy currents so this type of construction is practicable only at power and audio frequencies. The discussion in this section is confined to transformers operating at such frequencies.

### *Voltage and Turns Ratio*

For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns on the coil. If the two coils of a transformer are in the same field (which is the case when both are wound on the same closed core) it follows that the induced voltages will be proportional to the number of turns on each coil. In the case of the primary, or coil connected to the source of power, the induced voltage is practi-

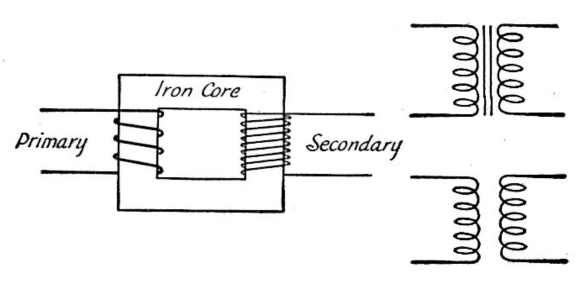



*Fig. 2-31* — **The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.**

cally equal to, and opposes, the applied voltage. Hence, for all practical purposes,

$$E_s = \frac{n_s}{n_p} E_p$$

where $E_s$ = Secondary voltage
$E_p$ = Primary voltage
$n_s$ = Number of turns on secondary
$n_p$ = Number of turns on primary

The ratio $n_s/n_p$ is called the **turns ratio** of the transformer.

Example: A transformer has a primary of 400 turns and a secondary of 2800 turns, and 115 volts is applied to the primary. The secondary voltage will be

$$E_s = \frac{n_s}{n_p} E_p = \frac{2800}{400} \times 115 = 7 \times 115 = 805 \text{ volts}$$

Also, if 805 volts is applied to the 2800-turn winding (which then becomes the primary) the output voltage from the 400-turn winding will be 115 volts.

Either winding of a transformer can be used as the primary, *providing* the winding has enough turns to induce a voltage equal to the applied voltage without requiring an excessive current flow.

### *Effect of Secondary Current*

The current that flows in the primary when no current is taken from the secondary is called the **magnetizing current** of the transformer. In any properly-designed transformer the primary inductance will be so large that the magnetizing current will be quite small. The power consumed by the transformer when the secondary is "open" — that is, not delivering power — is only the amount necessary to supply the losses in the iron core and in the resistance of the wire of which the primary is wound.

When current is drawn from the secondary winding, the secondary current sets up a magnetic field of its own in the core. The field from the secondary current always reduces the strength of the original field. But if the induced voltage in the primary is to equal the applied voltage, the original field *must* be maintained. Consequently, the primary current must change in such a way that the effect of the field set up by the secondary current is completely canceled. This is accomplished when the primary draws additional current that sets up a field exactly equal to the field set up by the secondary current, but which opposes the secondary field. The additional primary current is thus 180 degrees out of phase with the secondary current.

In practical calculations on transformers it is convenient to neglect the magnetizing current and to assume that the primary current is caused entirely by the secondary load. This is justifiable because the magnetizing current should be very small in comparison with the load current when the latter is near the rated value.

If the magnetic fields set up by the primary and secondary currents are to be equal, the primary current multiplied by the primary turns must equal the secondary current multiplied by the secondary turns. From this it follows that the primary current will be equal to the secondary current multiplied by the turns ratio, secondary to primary, or

$$I_p = \frac{n_s}{n_p} I_s$$

where $I_p$ = Primary current
$I_s$ = Secondary current
$n_p$ = Number of turns on primary
$n_s$ = Number of turns on secondary

Example: Suppose that the secondary of the transformer in the previous example is delivering a current of 0.2 ampere to a load. Then the primary current will be

$$I_p = \frac{n_s}{n_p} I_s = \frac{2800}{400} \times 0.2 = 7 \times 0.2 = 1.4 \text{ amp.}$$

Although the secondary *voltage* is *higher* than the primary voltage, the secondary *current* is *lower* than the primary current, and by the same ratio.

### *Power Relationships; Efficiency*

A transformer cannot create power; it can only transfer and transform it. Hence, the power taken from the secondary cannot exceed that taken by the primary from the source of applied e.m.f. There is always some power loss in the resistance of the coils and in the iron core, so in all practical cases the power taken from the source will exceed that taken from the secondary. Thus,

$$P_o = nP_i$$

where $P_o$ = Power output from secondary
$P_i$ = Power input to primary
$n$ = Efficiency factor

The efficiency, $n$, always is less than 1. It is usually expressed as a percentage; if $n$ is 0.65, for instance, the efficiency is 65 per cent.

Example: A transformer has an efficiency of 85% at its full-load output of 150 watts. The power input to the primary at full secondary load will be

$$P_i = \frac{P_o}{n} = \frac{150}{0.85} = 176.5 \text{ watts}$$

The efficiency of a transformer is usually — by design — highest at the normal power output for which it is rated. The efficiency decreases with either lower or higher outputs. On the other hand, the *losses* in the transformer are relatively small at low output but increase as more power is taken. The amount of power that the transformer can handle is determined

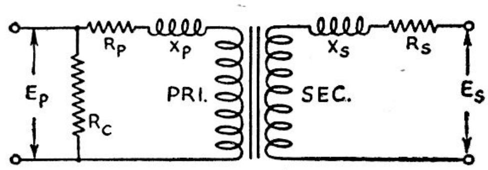



*Fig. 2-32* — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance *Rc* is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected in many approximate calculations.

by its own losses, because these heat the wire and core and raise the operating temperature. There is a limit to the temperature rise that can be tolerated, because too-high temperature either will melt the wire or break down the insulation between turns. A transformer always can be operated at reduced output even though the efficiency is low, because the actual loss also will be low under such conditions.

The full-load efficiency of small power transformers such as are used in radio receivers and transmitters usually lies between about 60 per cent and 90 per cent, depending upon the size and design.

### *Leakage Reactance*

In a practical transformer not all of the magnetic flux is common to both windings, although in well-designed transformers the amount of flux that "cuts" one coil and not the other is only a small percentage of the total flux. This **leakage flux** acts in the same way as flux about any coil that is not coupled to another coil; that is, it causes an e.m.f. of self-induction. Consequently, there are small amounts of **leakage inductance** associated with both windings of the transformer, but not common to them. Leakage inductance acts in exactly the same way as an equivalent amount of ordinary inductance inserted in series with the circuit. It has, therefore, a certain reactance, depending upon the amount of leakage inductance and the frequency. This reactance is called **leakage reactance.**

In the primary, the current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing primary current, hence it increases as more current is drawn from the secondary. The induced voltage consequently decreases, because the applied voltage has been reduced by the voltage drop in the primary leakage reactance. The secondary induced voltage also decreases proportionately.

When current flows in the secondary circuit the secondary leakage reactance causes an additional voltage drop that further reduces the voltage available from the secondary terminals. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistances of the primary and secondary windings of the transformer also cause voltage drops when current is flowing; although these voltage drops are not in phase with those caused by leakage reactance, together they result in a lower secondary voltage under load than is indicated by the turns ratio of the transformer.

At power frequencies (60 cycles) the voltage at the secondary, with a reasonably well-designed transformer, should not drop more than about 10 per cent from open-circuit conditions to full load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies because the leakage reactance increases directly with the frequency.

### *Impedance Ratio*

In an ideal transformer — one without losses or leakage reactance — the following relationship is true:

$$Z_p = Z_s N^2$$

where $Z_p$ = Impedance of primary as viewed from source of power
$Z_s$ = Impedance of load connected to secondary
$N$ = Turns ratio, primary to secondary

That is, a load of any given impedance connected to the *secondary* of the transformer will be changed to a different value "looking into" the *primary* from the source of power. The amount of impedance transformation is proportional to the square of the primary-to-secondary turns ratio.

Example: A transformer has a primary-to-secondary turns ratio of 0.6 (primary has 6/10 as many turns as the secondary) and a load of 3000 ohms is connected to the secondary. The impedance looking into the primary then will be

$$Z_p = Z_s N^2 = 3000 \times (0.6)^2 = 3000 \times 0.36 = 1080 \text{ ohms}$$

By choosing the proper turns ratio, the impedance of a fixed load can be transformed to any desired value, within practical limits. The transformed or "reflected" impedance has the same phase angle as the actual load impedance; if the load is a pure resistance the load presented by the primary to the source of power also will be a pure resistance.

The above relationship is sufficiently accurate in practice to give quite adequate results, even though it is based on an "ideal" transformer. Aside from the normal design requirements of reasonably low internal losses and low leakage reactance, the only other requirement to be met is that the primary have enough inductance to operate with low magnetizing current at the voltage applied to the primary. Despite a common — but mistaken — impression, a transformer operating with

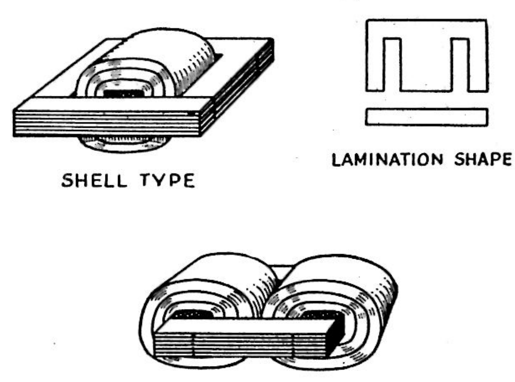



*Fig. 2-33* — **Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.**

a load does not "have" an impedance; the primary impedance — *as it looks to the source of power* — is determined by the load connected to the secondary and by the turns ratio. If the characteristics of the transformer have an appreciable effect on the impedance presented to the power source, the transformer is either poorly designed or is not suited to the voltage applied to it. Most transformers will operate quite well at voltages from slightly above to well below the design figure.

### *Impedance Matching*

Many devices require a specific value of load resistance (or impedance) for optimum operation. The resistance of the actual load that is to dissipate the power may differ widely from this value; so the transformer is frequently called upon to transform the actual load into one of the desired value. This is called **impedance matching.** From the preceding,

$$N = \sqrt{\frac{Z_s}{Z_p}}$$

where $N$ = Required turns ratio, secondary to primary

$Z_s$ = Impedance of load connected to secondary

$Z_p$ = Impedance required

Example: A vacuum-tube a.f. amplifier requires a load of 5000 ohms for optimum performance, and is to be connected to a loudspeaker having an impedance of 10 ohms. The turns ratio, secondary to primary, required in the coupling transformer is

$$N = \sqrt{\frac{Z_s}{Z_p}} = \sqrt{\frac{10}{5000}} = \sqrt{\frac{1}{500}} = \frac{1}{22.4}$$

The primary therefore must have 22.4 times as many turns as the secondary.

Impedance matching means, in general, adjusting the load impedance — by means of a transformer or otherwise — to a desired value. However, there is also another meaning. It is possible to show that any source of power will have its *maximum possible* output when the impedance of the load is equal to the internal impedance of the source. The impedance of the source is said to be "matched" under this condition. However, the efficiency is only 50 per cent in such a case; just as much power is used up in the source as is delivered to the load. Because of the poor efficiency, this type of impedance matching is limited to cases where only a small amount of power is available. Getting the most power output may be more important than efficiency in such a case.

### *Transformer Construction*

Transformers usually are designed so that the magnetic path around the core is as short as possible. A short magnetic path means that the transformer will operate with fewer turns, for a given applied voltage, than if the path were long. It also helps to reduce flux leakage and therefore minimizes leakage reactance. The number of turns required also is affected by the cross-sectional area of the core. Transformer design data will be found in Chapter Seven.

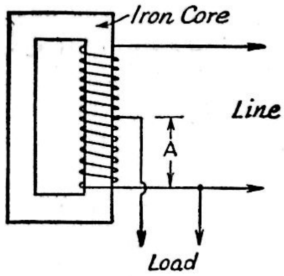



***Fig. 2-34*** **— The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (*A*) flow in opposite directions, so that the resultant current is the difference between them. The voltage across *A* is proportional to the turns ratio.**

Two core shapes are in common use, as shown in Fig. 2-33. In the shell type both windings are placed on the inner leg, while in the core type the primary and secondary windings may be placed on separate legs, if desired. This is sometimes done when it is necessary to minimize capacity effects between the primary and secondary, or when one of the windings must operate at very high voltage.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of laminations, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of eddy currents. The laminations are overlapped at the ends to make the magnetic path as continuous as possible and thus reduce flux leakage.

The number of turns required on the primary for a given applied e.m.f. is determined by the type of core material used, the maximum permissible flux density, and the frequency. As a rough indication, windings of small power transformers frequently have about two turns per volt on a core of 1-square-inch cross section and have a magnetic path 10 or 12 inches in length. A longer path or smaller cross section requires more turns per volt, and vice versa.

In most transformers the coils are wound in layers, with a thin sheet of paper insulation between each layer. Thicker insulation is used between coils and between coils and core.

### *Autotransformers*

The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 2-34; the principles just discussed apply equally well. A one-winding transformer is called an **autotransformer.** The section of the winding common to both the line and load circuits carries less current than the remainder of the coil, because the line and load currents are out of phase as explained previously. Hence the common section of the winding may be wound with comparatively small wire.

This advantage of the autotransformer is of practical value only when the primary (line) and secondary (load) voltages are not very different. On the other hand, it is frequently undesirable to have a direct connection between the primary and secondary circuits. For these reasons the autotransformer is used chiefly for boosting or reducing power-line voltage by relatively small amounts.

# Radio-Frequency Circuits

## ● RESONANCE

Fig. 2-35 shows a resistor, condenser and coil connected in series with a source of alternating current. Assume that the frequency can be varied over a wide range and that, at any frequency, the voltage of the source always has the same value.

At some *low* frequency the condenser reactance will be much larger than the resistance of *R*, and the inductive reactance will be small compared with either the reactance of *C* or the

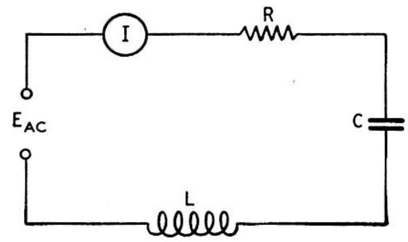



*Fig. 2-35* — **A series circuit containing *L*, *C* and *R* is "resonant" at the applied frequency when the reactance of *C* is equal to the reactance of *L*.**

resistance of *R*. (The resistance, *R*, is assumed to be the same at all frequencies.) On the other hand, at some very *high* frequency the reactance of *C* will be very small and the reactance of *L* will be very large. In the *low*-frequency case the amount of current that can flow will be determined practically entirely by the reactance of *C;* since $X_C$ is large at the low frequency, the current will be small. In the *high*-frequency case the amount of current that can flow will be determined almost wholly by the reactance of *L;* $X_L$ is large at the high frequency so the current is again small.

Now condenser reactance *decreases* as the frequency is raised, but inductive reactance *increases* with frequency. At *some* frequency, therefore, the reactances of *C* and *L* will be equal. At that frequency the voltage drop across the coil equals the voltage drop across the condenser, and since the two drops are 180 degrees out of phase they cancel each other completely. At that frequency the amount of current flow is determined wholly by the resistance, *R*. Also, at that frequency the current has its largest possible value (remember that we assumed the source voltage to be constant regardless of frequency). A series circuit in which the inductive and capacitive reactances are equal is said to be **resonant;** or, to be "in resonance" or "in tune" at the frequency for which the reactances are equal.

Resonance is not peculiar to radio-frequency circuits alone. It can occur at any a.c. frequency, including power-line frequencies. However, resonant circuits are used principally at radio frequencies; in fact, at those frequencies the circuits used almost always are resonant.

### *Resonant Frequency*

The frequency at which a series circuit is resonant is that for which $X_L = X_C$. Substituting the formulas for inductive and capacitive reactance gives

$$f = \frac{1}{2\pi\sqrt{LC}}$$

where $f$ = Frequency in cycles per second
$L$ = Inductance in henrys
$C$ = Capacitance in farads
$\pi$ = 3.14

These units are inconveniently large for radio-frequency circuits. A formula using more appropriate units is

$$f = \frac{10^6}{2\pi\sqrt{LC}}$$

where $f$ = Frequency in kilocycles (kc.)
$L$ = Inductance in microhenrys ($\mu$h.)
$C$ = Capacitance in micromicrofarads ($\mu\mu$fd.)
$\pi$ = 3.14

Example: The resonant frequency of a series circuit containing a 5-$\mu$h. coil and a 35-$\mu\mu$fd. condenser is

$$f = \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28 \times \sqrt{5 \times 35}}$$
$$= \frac{10^6}{6.28 \times 13.2} = \frac{10^6}{83} = 12{,}050 \text{ kc.}$$

The formula for resonant frequency is **not** affected by the resistance in the circuit.

### *Resonance Curves*

If a plot is drawn of the current flowing in the circuit of Fig. 2-35 as the frequency is varied (the applied voltage being constant) it would look like one of the curves in Fig. 2-36. At frequencies very much higher than the resonant frequency the current is limited by the inductive reactance; the condenser and resistor have only a negligible part. At frequencies very much lower than resonance the condenser limits the current, the resistor and inductance playing very little part. Exactly at resonance the current is limited only by the resistance; the smaller the resistance the larger the resonant current. The shape of the **resonance curve** at frequencies *near* resonance is determined by the ratio of reactance to resistance at the particular frequency considered. If the reactance of either the coil or condenser is of the same order of magnitude as the resistance, the current decreases rather slowly as the frequency is moved in either direction away from resonance. Such a curve is said to be **broad.** On the other hand, if the reactance is considerably larger than the resistance the current decreases rapidly as the

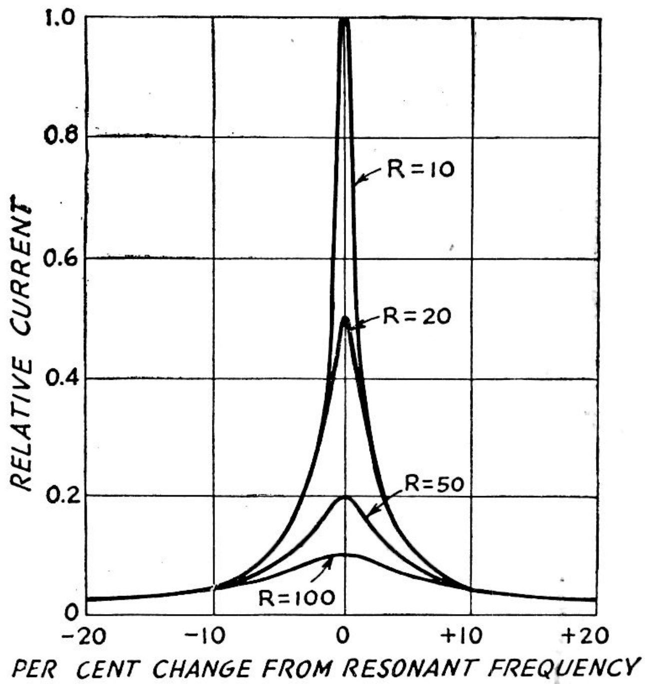



*Fig. 2-36* — Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms (minimum $Q = 10$). Note that at frequencies at least plus or minus ten per cent away from the resonant frequency the current is substantially unaffected by the resistance in the circuit.

frequency moves away from resonance and the circuit is said to be **sharp.** Curves of differing sharpness are shown in Fig. 2-36. A sharp circuit will respond a great deal more readily to the resonant frequency than to frequencies quite close to resonance; a broad circuit will respond almost equally well to a group or band of frequencies centering around the resonant frequency.

Both types of resonance curves are useful. A sharp circuit gives good **selectivity** — the ability to select one desired frequency and discriminate against others. A broad circuit is used when the apparatus must give about the same response over a band of frequencies rather than to a single frequency alone.

## Q

Most diagrams of resonant circuits show only inductance and capacitance; no resistance is indicated. Nevertheless, resistance is always present. At frequencies up to perhaps 30 Mc. this resistance is mostly in the wire of the coil. Above this frequency energy loss in the condenser (principally in the solid dielectric which must be used to form an insulating support for the condenser plates) becomes appreciable. This energy loss is equivalent to resistance. When maximum sharpness or selectivity is needed the object of design is to reduce the inherent resistance to the lowest possible value.

We mentioned above that the sharpness of the resonance curve is determined by the ratio of reactance to resistance. The value of the reactance of *either* the coil or condenser at the resonant frequency, divided by the resistance in the circuit, is called the $Q$ (quality factor) of the circuit, or

$$Q = \frac{X}{R}$$

where $Q$ = Quality factor
$X$ = Reactance of either coil or condenser, in ohms
$R$ = Resistance in ohms

Example: The coil and condenser in a series circuit each have a reactance of 350 ohms at the resonant frequency. The resistance is 5 ohms. Then the $Q$ is

$$Q = \frac{X}{R} = \frac{350}{5} = 70$$

Since the same current flows in $R$ that flows in $X$, the $Q$ of the circuit also is the ratio of the reactive power to the "real" power, or power dissipated in the resistance. The term "volt-ampere-to-watt" ratio or, when the power is large, "kva.-to-kw. ratio," therefore is sometimes used instead of "$Q$." To put it another way, the $Q$ of the circuit is the ratio of the energy stored (in either the magnetic or electric field) to the energy dissipated as heat in the resistance.

The effect of $Q$ on the sharpness of resonance of a circuit is shown by the curves of Fig. 2-37. In these curves the frequency change is shown in percentage above and below the resonant frequency. $Q$s of 10, 20, 50 and 100 are shown; these values cover much of the range commonly used in radio work.

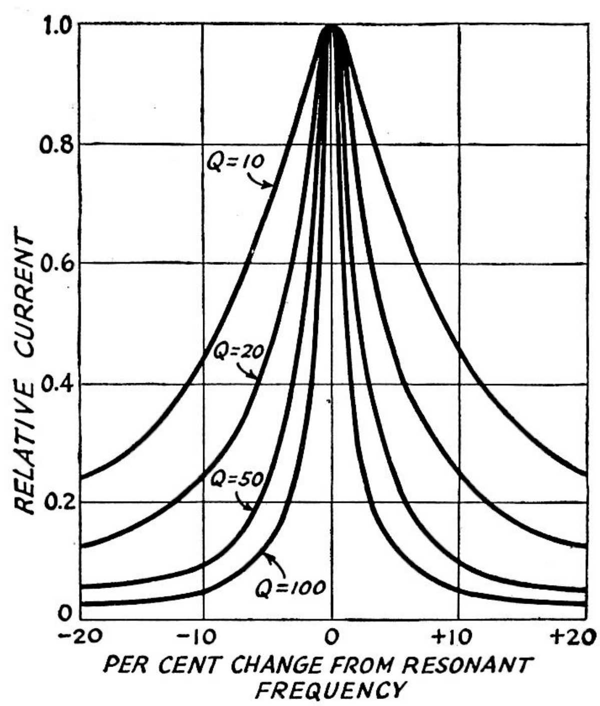



*Fig. 2-37* — Current in series-resonant circuits having different $Q$s. In this graph the current at resonance is assumed to be the same in all cases. The lower the $Q$, the more slowly the current decreases as the applied frequency is moved away from resonance.

### *Voltage Rise*

When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage that appears across either the coil or condenser is considerably higher than the applied voltage. The current in the circuit is limited only by the actual resistance of the coil-condenser combination in the circuit and may have a relatively high value; however, the same current flows through the high reactances of the coil and condenser and causes large voltage drops. (As explained above, the reactances are of opposite types and hence the voltages are opposite in phase, so the net voltage *around the circuit* is only that which is applied.) The ratio of the reactive voltage to the applied voltage is equal to the ratio of reactance to resistance. This ratio is the $Q$ of the circuit. Therefore, the voltage across either the coil or condenser is equal to $Q$ times the voltage inserted in series with the circuit.

Example: The inductive reactance of a circuit is 200 ohms, the capacitive reactance is 200 ohms, the resistance 5 ohms, and the applied voltage is 50. The two reactances cancel and there will be but 5 ohms of pure resistance to limit the current flow. Thus the current will be 50/5, or 10 amperes. The voltage developed across either the coil or the condenser will be equal to its reactance times the current, or $200 \times 10 = 2000$ volts. An alternate method: The $Q$ of the circuit is $X/R = 200/5 = 40$. The reactive voltage is equal to $Q$ times the applied voltage, or $40 \times 50 = 2000$ volts.

### *Parallel Resonance*

When a variable-frequency source of constant voltage is applied to a parallel circuit of the type shown in Fig. 2-38 there is a resonance effect similar to that in a series circuit. However, in this case the current (measured at the point indicated) is *smallest* at the frequency for which the coil and condenser reactances are equal. At that frequency the current through $L$ is exactly canceled by the out-of-phase current through $C$, as explained in an earlier section, so that only the current taken by $R$ flows in the line. At frequencies *below* resonance the current through $L$ is larger than that through $C$, because the reactance of $L$ is smaller and that of $C$ higher at low frequencies; there is only partial cancellation of the two reactive currents and the line current therefore is larger than the current taken by $R$ alone. At frequencies *above* resonance the situation is reversed and more current flows through $C$ than through $L$, so the line current again increases. The current at resonance, being determined wholly by $R$, will be small if $R$ is large and large if $R$ is small.

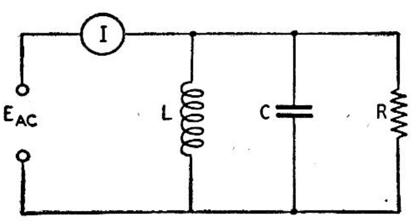



*Fig. 2-38* — Circuit illustrating parallel resonance.

The resistance $R$ shown in Fig. 2-38 seldom is an actual physical resistor. In most cases it will be an "equivalent" resistance that corresponds to the effect of an actual energy loss in the circuit. This energy loss can be inherent in the coil or condenser, or may represent energy transferred to a load by means of the resonant circuit. (For example, the resonant circuit may be used for transferring power from a vacuum-tube amplifier to an antenna system.)

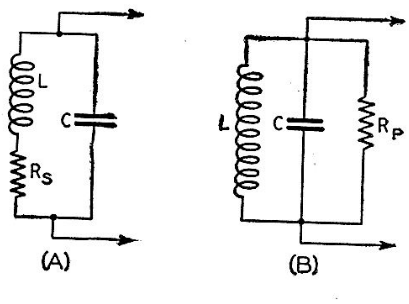



*Fig. 2-39* — Series and parallel equivalents when the two circuits are resonant. The series resistor, $R_s$, in A can be replaced by an equivalent parallel resistor, $R_p$, in B, and vice versa.

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-39 will behave identically, when an external voltage is applied, if (1) $L$ and $C$ are the same in both cases; and (2) $R_p$ multiplied by $R_s$ equals the square of the reactance (at resonance) of either $L$ or $C$. When these conditions are met the two circuits will have the same $Q$s. (These statements are approximate, but are quite accurate if the $Q$ is 10 or more.) Now the circuit at A is a *series* circuit if it is viewed from the "inside" — that is, going around the loop formed by $L$, $C$ and $R$ — so its $Q$ can be found from the ratio of $X$ to $R_s$.

What this means is that a circuit like that of Fig. 2-39A has an equivalent **parallel impedance** (at resonance) equal to $R_p$, the relationship between $R_s$ and $R_p$ being as explained above. Although $R_p$ is not an actual resistor, to the source of voltage the parallel-resonant circuit "looks like" a pure resistance of that value. It is "pure" resistance because the coil and condenser currents are 180 degrees out of phase and are equal; thus there is no reactive current. At the resonant frequency, then, the parallel impedance of a resonant circuit is

$$Z_r = QX$$

where $Z_r$ = Resistive impedance at resonance
$Q$ = Quality factor
$X$ = Reactance (in ohms) of either the coil or condenser

Example: The parallel impedance of a circuit having a $Q$ of 50 and having inductive and capacitive reactances of 300 ohms will be

$$Z_r = QX = 50 \times 300 = 15{,}000 \text{ ohms.}$$

At frequencies off resonance the impedance is no longer purely resistive because the coil and condenser currents are not equal. The off-resonant impedance therefore is complex, and is lower than the resonant impedance for the reasons previously outlined.

The higher the $Q$ of the circuit, the higher the parallel impedance. Curves showing the variation of impedance (with frequency) of a parallel circuit have just the same shape as the curves showing the variation of current with frequency in a series circuit. Fig. 2-40 is a set of such curves.

### *Q of Loaded Circuits*

In many applications of resonant circuits the only power lost is that dissipated in the resistance of the circuit itself. At frequencies below 30 Mc. most of this resistance is in the coil. Within limits, increasing the number of turns on the coil increases the reactance faster than it raises the resistance, so coils for circuits in which the $Q$ must be high are made with relatively large inductance for the frequency under consideration.

However, when the circuit delivers energy to a load (as in the case of the resonant circuits used in transmitters) the energy consumed in the circuit itself is usually negligible compared with that consumed by the load. The equivalent of such a circuit is shown in Fig. 2-41A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the load is at least ten times as great as the power lost in the coil and condenser, the parallel impedance of the resonant circuit itself will be so high compared with the resistance of the load that for all practical purposes the impedance of the combined circuit is equal to the load resistance. Under these conditions the $Q$ of a parallel-

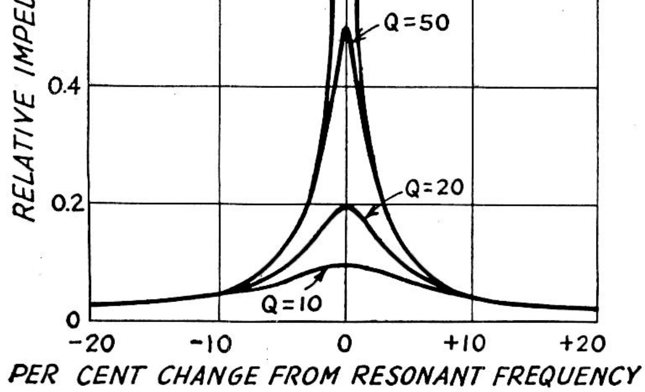



*Fig. 2-40* — **Relative impedance of parallel-resonant circuits with different $Q$s. These curves are similar to those in Fig. 2-37 for current in a series-resonant circuit. The effect of $Q$ on impedance is most marked near the resonant frequency.**

resonant circuit loaded by a resistive impedance is

$$Q = \frac{Z}{X}$$

where $Q$ = Quality factor
$Z$ = Parallel load resistance (ohms)
$X$ = Reactance (ohms) of either the coil or condenser

Example: A resistive load of 3000 ohms is connected across a resonant circuit in which the inductive and capacitive reactances are each 250 ohms. The circuit $Q$ is then

$$Q = \frac{Z}{X} = \frac{3000}{250} = 12$$

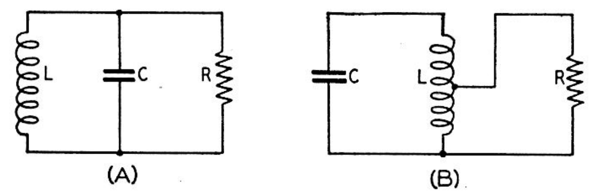



*Fig. 2-41* — **The equivalent circuit of a resonant circuit delivering power to a load. The resistor $R$ represents the load resistance. At B the load is tapped across part of $L$, which by transformer action is equivalent to using a higher load resistance across the whole circuit.**

The effective $Q$ of a circuit loaded by a parallel resistance becomes higher when the reactances of the coil and condenser are decreased. A circuit loaded with a relatively low resistance (a few thousand ohms) must have low-reactance elements (large capacitance and small inductance) to have reasonably high $Q$.

The effect of a given load resistance on the $Q$ of a circuit can be changed by connecting the load across only part of the circuit. A common method is to tap the load across part of the coil, as shown in Fig. 2-41B. The smaller the portion of the coil across which the load is tapped, the less the loading on the circuit; in other words, tapping the load "down" is equivalent to connecting a higher value of load resistance across the whole circuit. This is similar in principle to impedance transformation with an iron-core transformer. In high-frequency resonant circuits the impedance ratio does not vary exactly as the square of the turns ratio, because all the magnetic flux lines do not cut every turn of the coil. A desired reflected impedance usually must be obtained by experimental adjustment.

### *L/C Ratio*

The formula for resonant frequency of a circuit shows that the same frequency always will be obtained so long as the *product* of $L$ and $C$ is constant. Within this limitation, it is evident that $L$ can be large and $C$ small, $L$ small and $C$ large, etc. The relation between the two for a fixed frequency is called the ***L/C* ratio.** A **high-*C*** circuit is one which has more capacity than "normal" for the frequency; a **low-*C*** circuit one which has less than normal capacity. These terms depend to a considerable extent upon the particular application considered, and have no exact numerical meaning.

### *LC Constants*

As pointed out in the preceding paragraph, the product of inductance and capacity is constant for any given frequency. It is frequently convenient to use the numerical value of the **LC constant** when a number of calculations have to be made involving different $L/C$ ratios for the same frequency. The constant for any frequency is given by the following equation:

$$LC = \frac{25{,}330}{f^2}$$

where $L$ = Inductance in microhenrys ($\mu$h.)
$C$ = Capacitance in micromicrofarads ($\mu\mu$fd.)
$f$ = Frequency in megacycles.

Example: Find the inductance required to resonate at 3650 kc. (3.65 Mc.) with capacitances of 25, 50, 100, and 500 $\mu\mu$fd. The $LC$ constant is

$$LC = \frac{25{,}330}{(3.65)^2} = \frac{25{,}330}{13.35} = 1900$$

With 25 $\mu\mu$fd. $L = 1900/C = 1900/25 = 76\ \mu$h.
50 $\mu\mu$fd. $L = 1900/C = 1900/50 = 38\ \mu$h.
100 $\mu\mu$fd. $L = 1900/C = 1900/100 = 19\ \mu$h.
500 $\mu\mu$fd. $L = 1900/C = 1900/500 = 3.8\ \mu$h.

## COUPLED CIRCUITS

### *Energy Transfer and Loading*

Two circuits are **coupled** when energy can be transferred from one to the other. The circuit delivering power is called the **primary** circuit; the one receiving power is called the **secondary** circuit. The power may be practically all dissipated in the secondary circuit itself (this is usually the case in receiver circuits) or the secondary may simply act as a medium through which the power is transferred to a load resistance where it does work. In the latter case, the coupled circuits may act as a radio-frequency impedance-matching device. The matching can be accomplished by adjusting the loading on the secondary and by varying the amount of coupling between the primary and secondary.

A general understanding of coupling methods is essential in amateur work, but there is seldom, if ever, need for *calculation* of the performance of coupled circuits. Very few radio amateurs have the equipment necessary for measuring the quantities that enter into such calculations. In actual practice, the adjustment of a coupled circuit is a cut-and-try process. Satisfactory results readily can be obtained if the principles are understood.

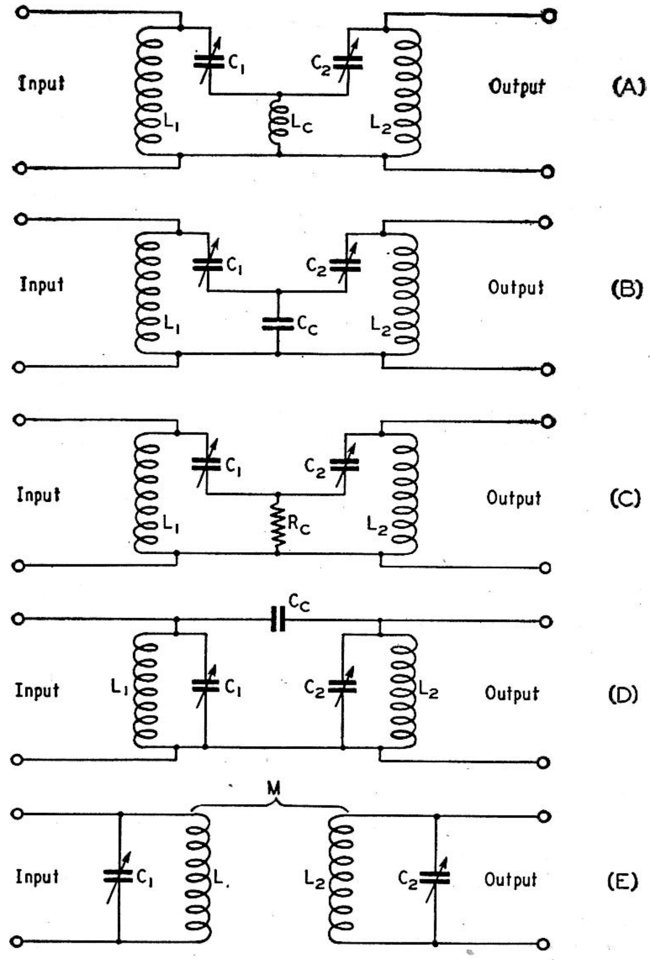



*Fig. 2-42* — **Basic methods of circuit coupling.**

### *Coupling by a Common Circuit Element*

One method of coupling between two resonant circuits is through a circuit element common to both. The three variations of this type of coupling shown at A, B and C of Fig. 2-42, utilize a common inductance, capacitance and resistance, respectively. Current circulating in one $LC$ branch flows through the common element ($L_c$, $C_c$, or $R_c$) and the voltage developed across this element causes current to flow in the other $LC$ branch.

If both circuits are resonant to the same frequency, as is usually the case, the value of impedance — reactance or resistance — required for maximum energy transfer is generally quite small compared to the other reactances in the circuits. The common-circuit-element method of coupling is used only occasionally in amateur apparatus.

### *Capacitive Coupling*

In the circuit at D the coupling increases as the capacitance of $C_c$, the "coupling condenser," is made greater (reactance of $C_c$ is decreased). When two resonant circuits are coupled by this means, the capacitance required for maximum energy transfer is quite small if the $Q$ of the secondary circuit is at all high. For example, if the parallel impedance of the secondary circuit is 100,000 ohms, a reactance of 10,000 ohms or so in the condenser will give ample coupling. The corresponding capacitance required is only a few micromicrofarads at high frequencies.

### *Inductive Coupling*

Fig. 2-42E shows inductive coupling, or coupling by means of the magnetic field. A circuit of this type resembles the iron-core transformer, but because only a small percentage of the magnetic flux lines set up by one coil cut the turns of the other coil, the simple relationships between turns ratio, voltage ratio and impedance ratio in the iron-core transformer do not hold.

Three common types of inductively-coupled circuits are shown in Fig. 2-43. In the first two, only one circuit actually is resonant. The circuit at A is frequently used in receivers for coupling between amplifier tubes when the tuning of the circuit must be varied to respond to signals of different frequencies. Circuit B is used principally in transmitters, for coupling a radio-frequency amplifier to a resistive load. Circuit C is used for fixed-frequency amplification in receivers. The same circuit also is used in transmitters for transferring power to a load that has both reactance and resistance.

In circuits A and B the coupling between the primary and secondary coils usually is "tight" — that is, the coefficient of coupling between the coils is large. With tight coupling either circuit operates much as though the device to which the untuned coil is connected were simply tapped across a corresponding number of turns on the tuned-circuit coil. Any resistance in the circuit to which the untuned coil is connected is coupled into the tuned circuit in proportion to the mutual inductance. This "coupled" resistance increases the effective series resistance of the tuned circuit, thereby lowering its $Q$ and selectivity. If the circuit to which the untuned coil is connected has reactance, a certain amount of reactance will be "coupled in" to the tuned circuit. The coupled reactance makes it necessary to readjust the tuning whenever the coupling is changed, because coupled reactance tunes the circuit just as the actual coil and condenser reactance does.

These circuits may be used for impedance matching by adjusting the mutual inductance between the coils. This can be done by varying the coupling, changing the number of turns in the untuned coil, or both. The parallel impedance of the tuned circuit is affected by the coupled-in resistance in the same way as it would be by a corresponding increase in the actual series resistance. The larger the value of coupled-in resistance the lower the parallel impedance. By proper choice of the number of turns on the untuned coil, and by adjustment of the coupling, the parallel impedance of the tuned circuit may be adjusted to the value required for the proper operation of the device to which it is connected.

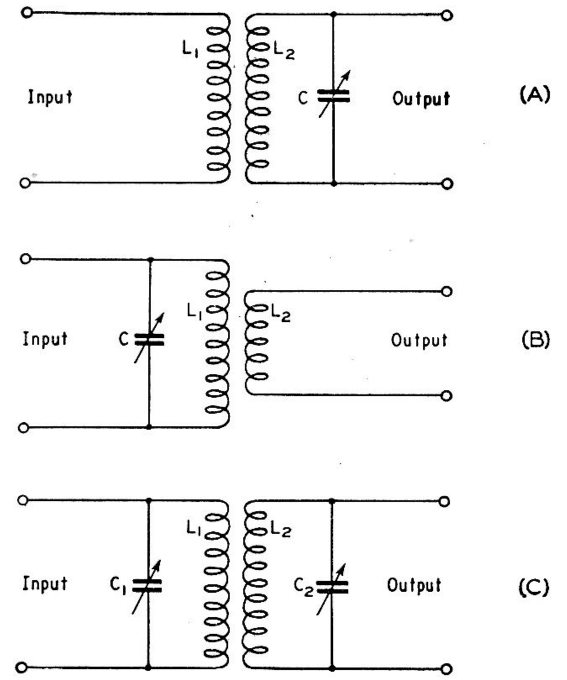


*Fig. 2-43* — **Types of inductively-coupled circuits. In A and B, one circuit is tuned, the other untuned. C shows the method of coupling between two tuned circuits.**

### *Coupled Resonant Circuits*

When the primary and secondary circuits are both tuned, as in Fig. 2-43C, the resonance effects in both circuits make the operation somewhat more complicated than in the simpler circuits just considered. Imagine first that the two circuits are not coupled and that each is independently tuned to the resonant frequency. The impedance of each will be purely resistive. If the two are then coupled, the secondary will couple resistance into the primary, causing its parallel impedance to decrease. As the coupling is made greater (without changing the tuning of either circuit) the coupled resistance becomes larger and the parallel impedance of the primary continues to decrease. Also, as the coupling is made tighter the amount of power transferred from the primary to the secondary will increase — but only up to a certain point. The power transfer becomes maximum at a "critical" value of coupling, but then decreases if the coupling is tightened beyond the critical point. At **critical coupling,** the resistance coupled into the primary circuit is equal to the resistance of the primary itself. This represents the matched-impedance condition and gives maximum power transfer.

Critical coupling is a function of the $Q$s of the two circuits taken independently. A higher coefficient of coupling is required to reach critical coupling when the $Q$s are low; if the $Q$s are high, as in receiving applications, a coupling coefficient of a few per cent may give critical coupling.

With loaded circuits it is not impossible for the $Q$ to reach such low values that critical coupling cannot be obtained even with the highest practicable coefficient of coupling (coils

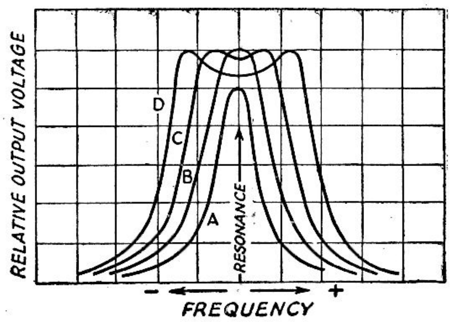



*Fig. 2-44* — **Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output voltage is measured across the secondary.**

as physically close as possible). In such case the only way to secure sufficient coupling is to increase the $Q$ of one or both of the coupled circuits. This can be done either by decreasing the $L/C$ ratio or by tapping the load down on the secondary coil. If the load resistance is known beforehand, the circuits may be designed for a $Q$ in the vicinity of 10 or so with assurance that sufficient coupling will be available; if unknown, the proper $Q$s can be determined by experiment.

### *Selectivity*

In A and B, Fig. 2-43, only one circuit is tuned and the selectivity curve will be that of a single resonant circuit having the appropriate $Q$. As stated, the effective $Q$ depends upon the resistance connected to the untuned coil.

In Fig. 2-43C, the selectivity is the same as that of a single tuned circuit having a $Q$ equal to the *product* of the $Q$s of the individual circuits — *if* the coupling is below critical and both circuits are tuned to resonance. The $Q$s of the individual circuits are affected by the degree of coupling, because each couples resistance into the other; the tighter the coupling, the lower the individual $Q$s and therefore the lower the over-all selectivity.

If both circuits are independently tuned to resonance, the over-all selectivity will vary about as shown in Fig. 2-44 as the coupling is varied. At loose coupling, $A$, the output voltage (across the secondary circuit) is small and the selectivity is high. As the coupling is increased the secondary voltage also increases until critical coupling, $B$, is reached. At this point the output voltage at the resonant frequency is maximum but the selectivity is lower than with looser coupling. At still tighter coupling, $C$, the output voltage at the resonant frequency decreases, but as the frequency is varied either side of resonance it is found that there are two "humps" to the curve, one on either side of resonance. With very tight coupling, $D$, there is a further decrease in the output voltage at resonance and the "humps" are farther away from the resonant frequency. Resonance curves such as those at $C$ and $D$ are called **flat-topped** because the output voltage does not change much over an appreciable band of frequencies.

Note that the off-resonance humps have the same maximum value as the resonant output voltage at critical coupling. These humps are caused by the fact that at frequencies off resonance the secondary circuit is reactive and couples reactance as well as resistance into the primary. The coupled resistance decreases off resonance and the humps represent a new condition of impedance matching — at a frequency to which the primary is detuned by the coupled-in reactance from the secondary.

When the two circuits are tuned to slightly *different* frequencies a double-humped resonance curve results even though the coupling is below critical. This is to be expected, because each circuit will respond best to the frequency to which it is tuned. Tuning of this type is called **stagger tuning,** and often is used when substantially uniform response over a wide band of frequencies is desired.

### *Link Coupling*

A modification of inductive coupling, called **link coupling,** is shown in Fig. 2-45. This gives the effect of inductive coupling between two coils that have no mutual inductance; the link is simply a means for providing the mutual inductance. The total mutual inductance between two coils coupled by a link cannot be made as great as if the coils themselves were coupled. This is because the coefficient of coupling between air-core coils is considerably less than 1, and since there are two coupling points the over-all coupling coefficient is less than for any *pair* of coils. In practice this need not be disadvantageous because the power transfer can be made great enough by making the tuned circuits sufficiently high-$Q$. Link coupling is convenient when ordinary inductive coupling would be impracticable for constructional reasons. It finds wide use in transmitters, for example.

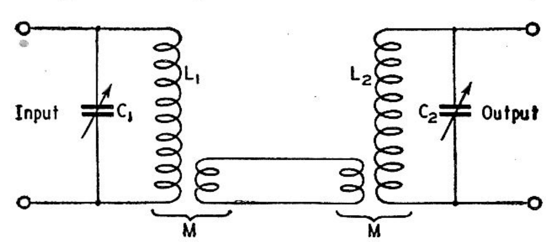



*Fig. 2-45* — **Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.**

The link coils usually have a small number of turns compared with the resonant-circuit coils. The number of turns is not greatly important, because the coefficient of coupling is relatively independent of the number of turns on either coil; it is more important that both link coils should have about the *same* number of turns. The length of the link between the coils is not critical if it is very small compared with the wavelength; if the length becomes an appreciable fraction of a wavelength the link operates more as a transmission line than as a means for providing mutual inductance. In such case it should be treated by the methods described in Chapter Ten.

### *Piezoelectric Crystals*

A number of crystalline substances found in nature have the ability to transform mechanical strain into an electrical charge, and vice versa. This property is known as **piezoelectricity.** A small plate or bar cut in the proper way from a quartz crystal, for example, and placed between two conducting electrodes, will be mechanically strained when the electrodes are connected to a source of voltage. Conversely, if the crystal is squeezed between two electrodes a voltage will develop between the electrodes.

Piezoelectric crystals can be used to transform mechanical energy into electrical energy, and vice versa. They are used, for example, in microphones and phonograph pick-ups, where mechanical vibrations are transformed into alternating voltages of corresponding frequency. They are also used in headsets and loudspeakers, transforming electrical energy into mechanical vibration. Crystal plates for these purposes are cut from large crystals of Rochelle salts.

Crystalline plates also are mechanical vibrators that have natural frequencies of vibration ranging from a few thousand cycles to several megacycles per second. The vibration frequency depends on the kind of crystal, the way the plate is cut from the natural crystal, and on the dimensions of the plate. Such a crystal is, in fact, the mechanical counterpart of an electrical tuned circuit; its resonant frequency is the natural frequency of the mechanical vibration. Because of the piezoelectric effect, the crystal plate can be coupled to an electrical circuit and made to substitute for a coil-and-condenser resonant circuit. The thing that makes crystals valuable as "resonators" is the fact that they have extremely high $Q$, ranging from 5 to 10 times the $Q$s obtainable with $LC$ resonant circuits.

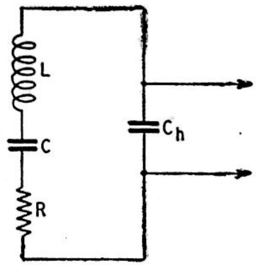



*Fig. 2-46* — Equivalent circuit of a crystal resonator. $L$, $C$ and $R$ are the electrical equivalents of mechanical properties of the crystal; $C_h$ is the capacitance of the electrodes with the crystal plate between them.

Analogies can be drawn between various mechanical properties of the crystal and the electrical characteristics of a tuned circuit. This leads to an "equivalent circuit" for the crystal. The electrical coupling to the crystal is through the electrodes between which it is sandwiched; these electrodes form, with the crystal as the dielectric, a small condenser like any other condenser constructed of two plates with a dielectric between. The crystal itself is an equivalent to a series-resonant circuit, and together with the capacitance of the electrodes forms the equivalent circuit shown in Fig. 2-46. The equivalent inductance of the crystal is extremely large and the series capacitance, $C$, is correspondingly low; this is the reason for the high $Q$ of a crystal. The electrode capacitance, $C_h$, is so very large compared with the series capacitance of the crystal that it has only a very small effect on the resonant frequency. It will be realized, also, that because $C_h$ is so large compared with $C$ the electrical coupling to the crystal is quite loose.

Crystal plates for use as resonators in radio-frequency circuits are almost always cut from quartz crystals, because quartz is by far the most suitable material for this purpose. Quartz crystals are used as resonators in receivers, to give highly-selective reception, and as frequency-controlling elements in transmitters.

# Practical Circuit Details

## ● COMBINED A.C. AND D.C.

Most radio circuits are built around vacuum tubes, and it is the nature of these tubes to require direct current (usually at a fairly high voltage) for their operation. They *convert* the direct current into an alternating current (and sometimes the reverse) at frequencies varying from ones well down in the audio range to well up in the superhigh range. The conversion process almost invariably requires that the direct and alternating currents meet somewhere in the circuit.

In this meeting, the a.c. and d.c. are actually combined into a single current that "pulsates" (at the a.c. frequency) about an *average* value equal to the direct current. This is shown in Fig. 2-47. It is easier, though, to think of them separately and to consider that the alternating current is **superimposed** on the direct current. Thus we look upon the actual current as having two components, one d.c. and the other a.c.

If the alternating current is a sine wave, its positive and negative alternations have the same maximum amplitude. When the wave is superimposed on a direct current the latter is alternately increased and decreased *by the same amount.* There is thus no *average* change in the direct current. If a d.c. instrument is being used to read the current, the reading will be exactly the same whether or not the sine-wave a.c. is superimposed.

However, there is actually more *power* in such a combination current than there is in the direct current alone. This is because power

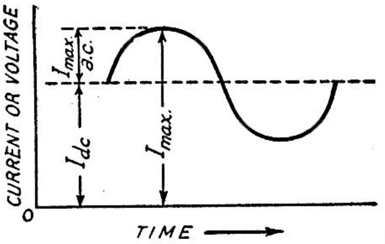



*Fig. 2-47* — Pulsating current, composed of an alternating current or voltage superimposed on a steady direct current or voltage.

varies as the *square* of the instantaneous value of the current, so more power is added to the circuit on the half-cycle of the a.c. wave that *increases* the instantaneous current than is subtracted on the half-cycle that *decreases* it. If the peak value of the alternating current is just equal to the direct current, the average power in the circuit is 1.5 times the power in the direct current alone.

In many circuits, also, we may have two alternating currents of different frequencies; for example, an audio frequency and a radio frequency may be combined in the same circuit. The two in turn may be combined with a direct current. In some cases, too, two r.f. currents of widely-different frequencies may be combined in the same circuit.

### *Series and Parallel Feed*

Fig. 2-48 shows in simplified form how d.c. and a.c. may be combined in a vacuum-tube circuit. (The tube is shown only in bare outline; so far as the d.c. is concerned, it can be looked upon as a resistance of rather high value. On the other hand, the tube may be looked upon as the *generator* of the a.c. The mechanism of tube operation is described in the next chapter.) In this case, we have assumed that the a.c. is at radio frequency, as suggested by the coil-and-condenser tuned circuit. We also assume that r.f. current can easily flow through the d.c. supply; that is, the impedance of the supply at radio frequencies is so small as to be negligible.

In the circuit at the left, the tube, tuned circuit, and d.c. supply all are connected in series. The direct current flows through the r.f. tuned circuit to get to the tube; the r.f. current generated by the tube flows through the d.c. supply to get to the tuned circuit. This is **series feed.** It works because the impedance of the d.c. supply at radio frequencies is so low that it does not affect the flow of *r.f.* current, and because the d.c. resistance of the coil is so low that it does not affect the flow of *direct* current.

In the circuit at the right the direct current does not flow through the r.f. tuned circuit, but instead goes to the tube through a second coil, *RFC* **(radio-frequency choke).** Direct current cannot flow through *L* because a **blocking condenser,** *C*, is placed in the circuit to prevent it. (Without *C*, the d.c. supply would be short-circuited by the low resistance of *L*.) On the other hand, the r.f. current generated by the tube can easily flow through *C* to the tuned circuit because the capacitance of *C* is intentionally chosen to have low reactance (compared with the impedance of the tuned circuit) at the radio frequency. The r.f. current cannot flow through the d.c. supply because the inductance of *RFC* is intentionally made so large that it has a very high reactance at the radio frequency. The resistance of *RFC*, however, is too low to have an appreciable effect on the flow of direct current. The two currents are thus in *parallel*, hence the name **parallel feed.**

Both types of feed are in use. They may be used for both a.f. and r.f. circuits. In parallel feed there is no d.c. voltage on the a.c. circuit (the blocking condenser prevents that); this is a desirable feature from the viewpoint of safety to the operator, because the voltages applied to tubes — particularly transmitting tubes — are dangerous to human beings. On the other hand, it is somewhat difficult to make an r.f. choke work well over a wide range of frequencies. Series feed is usually preferred, therefore, because it is relatively easy to keep the impedance between the a.c. circuit and the tube low.

### *By-Passing*

In the series-feed circuit just discussed, it was assumed that the d.c. supply had very low impedance at radio frequencies. This is not likely to be true in a practical power supply — if for no other reason than that the normal physical separation between the supply and the r.f. circuit would make it necessary to use rather long connecting wires or leads. At radio frequencies, even a few feet of wire can have fairly large reactance — too large to be considered a really "low-impedance" connection.

To get around this, an actual circuit would be provided with a **by-pass condenser,** as shown in Fig. 2-49. Condenser *C* is chosen to have low reactance at the operating frequency, and is installed right in the circuit where it can be wired to the other parts with quite short connecting wires. (The condenser will be an open circuit for the d.c. voltage across which it is connected, of course.) Since condenser *C* offers a low-impedance path, the r.f. current will tend to flow through it rather than through the d.c. supply; thus the current is confined to a known path rather than one of dubious impedance through the power supply.

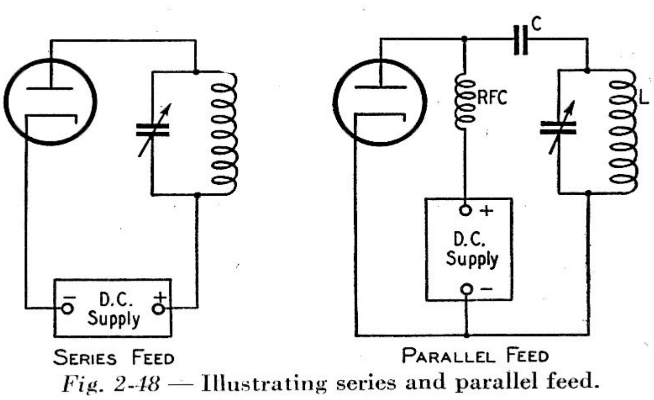



*Fig. 2-48* — Illustrating series and parallel feed.

To be effective, a by-pass should have very low impedance compared to the impedance of the circuit element around which it is supposed to shunt the current. The reactance of the condenser should not be more than one-tenth of the impedance of the by-passed part of the circuit. Very often the latter impedance is not known, in which case it is desirable to use the largest capacitance in the by-pass that circumstances permit. To make doubly sure that r.f. current will not flow through a non-r.f. circuit such as a power supply, an r.f. choke may be connected in the lead to the latter, as shown in Fig. 2-49. The choke, having high reactance, will prevent the r.f. from going where it is not wanted and thereby ensure that it goes where it *is* wanted — i.e., through the by-pass condenser.

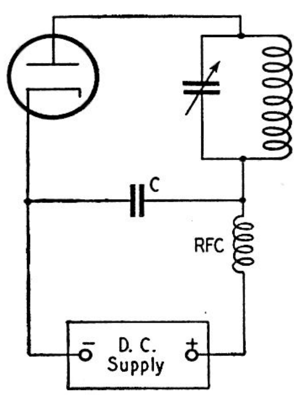



*Fig. 2-49* — **Typical use of a by-pass condenser in a series-feed circuit.**

The use of a by-pass condenser is not confined only to circuits where r.f. is to be kept out of a d.c. source. The same type of by-passing is used when audio frequencies are present in addition to r.f. Because the reactance of a condenser changes with frequency, it is readily possible to choose a capacitance that will represent a very low reactance at radio frequencies but that will have such high reactance at audio frequencies that it is practically an open circuit. A capacitance of 0.001 $\mu$fd. is practically a short-circuit for r.f., for example, but is almost an open circuit at audio frequencies. (The actual value of capacitance that is usable will be modified by the impedances concerned.)

By-pass condensers also are used in audio-frequency circuits, to carry the audio frequencies around a d.c. supply. In this case a capacitance of several microfarads is needed if the reactance is to be low enough at the lower audio frequencies.

### *Distributed Capacitance and Inductance*

In the discussions earlier in this chapter it was assumed that a condenser has only capacitance and that a coil has only inductance. Unfortunately, this is not strictly true. There is always a certain amount of inductance in a conductor of any length, and since a condenser is made up of conductors it is bound to have a little inductance in addition to its intended capacitance. Also, there is always capacitance between two conductors or between parts of the same conductor, and so we find that there is appreciable capacitance between the turns of an inductance coil.

This distributed inductance in a condenser and the distributed capacitance in a coil have important practical effects. Actually, every condenser is a tuned circuit, resonant at the frequency where its capacitance and distributed inductance have the same reactance. The same thing is true of a coil and its distributed capacitance. At frequencies well below these "natural" resonances, the condenser will act like a normal capacitance and the coil will act like a normal inductance. Near the natural resonant points, the coil and condenser act like self-tuned circuits. Above resonance, the condenser acts like an inductance and the coil acts like a condenser. If we want our circuit components to behave properly, they must always be used at frequencies well on the low side of their natural resonances.

Because of these effects, there is a limit to the amount of capacitance that can be used at a given frequency. There is a similar limit to the inductance that can be used. At audio frequencies, capacitances measured in microfarads and inductances measured in henrys are practicable. At low and medium radio frequencies, inductances of a few millihenrys and capacitances of a few thousand micromicrofarads are the largest practicable. At high radio frequencies, usable inductance values drop to a few microhenrys and capacitances to a few hundred micromicrofarads.

Distributed capacitance and inductance are important not only in r.f. tuned circuits, but in by-passing and choking as well. It will be appreciated that a by-pass condenser that actually acts like an inductance, or an r.f. choke that acts like a condenser, cannot work as it is intended they should. That is why you will find, in the circuits described later in this *Handbook*, by-pass condenser capacitances and r.f.-choke inductances that may look rather small — considering that, theoretically, a larger condenser or larger coil should be even more effective at its job.

### *Grounds*

Throughout this book you will find frequent references to **ground** and **ground potential.** When a connection is said to be "grounded" it does not mean that it actually goes to earth (although in many cases such earth connections *are* used). What it means, more often, is that an actual earth connection *could* be made to that point in the circuit without disturbing the operation of the circuit in any way. The term also is used to indicate a "common" point in the circuit where power supplies and metallic supports (such as a metal chassis) are electrically tied together. It is customary, for example, to "ground" the negative terminal of a d.c. power supply, and to "ground" the filament or heater power supplies for vacuum tubes. Since the cathode of a vacuum tube is a junction point for grid and plate voltage supplies, it is a natural point to "ground." Also, since the various circuits connected to the tube elements have at least one point connected to cathode, these points also are "returned to ground."

"Ground" is therefore a common reference point in the circuit. In circuit diagrams, it is customary (for the sake of making the diagrams easier to read) to show such common connections by the ground symbol rather than by showing a large number of wires all connected together.

"Ground potential" means that there is no "difference of potential" — that is, no voltage — between the circuit point and the earth. A direct earth connection at such a point would cause no disturbance to the operation of the circuit.

### *Single-Ended and Balanced Circuits*

With reference to ground, a circuit may be either single-ended (unbalanced) or balanced. In a single-ended circuit, one *side* of the circuit is connected to ground. In a balanced circuit, the *electrical midpoint* of the circuit is connected to ground, so that the circuit has two ends each at the same voltage "above" ground. A balanced circuit also is called a "symmetrical" circuit.

Typical single-ended and balanced circuits are shown in Fig. 2-50. R.f. circuits are shown in the upper line, while iron-core transformers (such as are used in power-supply and audio circuits) are shown in the lower line. The r.f. circuits may be balanced either by connecting the center of the coil to ground or by using a "balanced" or "split-stator" condenser — that is, one having two identical sets of stator and rotor plates with the rotor plates on the same shaft — and connecting the condenser rotor to ground. In the iron-core transformer, one or both windings may be tapped at the center of the winding to provide the ground connection.

In the single-ended circuit, only one side of the circuit is "hot" — that is, has a voltage that differs from ground potential. In the balanced circuit, both ends are "hot" and the grounded center point is "cold" — that is, at ground potential. The applications of both types of circuits are discussed in later chapters.

### *Nonlinear Circuits; Beats*

The circuits that have been discussed in this chapter are, essentially, ones obeying Ohm's Law. That is, an increase or decrease of the applied voltage causes an exactly proportional increase or decrease in current. (This neglects relatively minor effects such as the temperature rise and consequent change in resistance of conductors with increasing current, etc.) However, many devices (such as vacuum tubes under some conditions of operation) do not obey any such straightforward rules. There may be no current flow at all with an applied voltage of one polarity, but the current may be large if the polarity of the voltage is reversed. Also, the current may increase with increasing voltage up to a certain point and then stay at a fixed value no matter how much more the voltage is raised. Such devices, and the circuits in which they are used, are called **nonlinear.**

One important result of nonlinearity is the behavior of the circuit when two or more alternating currents of different frequencies are flowing in it. In a normal circuit, the two frequencies will have no particular effect on each other. However, if two (or more) alternating currents of different frequencies are present in a nonlinear circuit, additional currents having frequencies equal to the sum, and difference, of the original frequencies will be set up. These sum and difference frequencies are called the **beat** frequencies. For example, if frequencies of 2000 and 3000 kc. are present in a normal circuit only those two frequencies exist, but if they are passed through a nonlinear circuit there will be present in the output not only the two original frequencies of 2000 and 3000 kc. but also currents of 1000 (3000 − 2000) and 5000 (3000 + 2000) kc. Suitable circuits can be used to select the desired beat frequency.

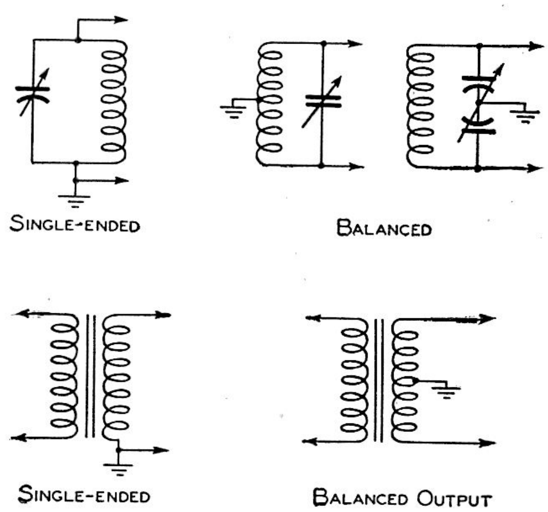



*Fig. 2-50* — **Single-ended and balanced circuits.**

Beat frequencies are generated, and used to advantage, in very many radio circuits. For example, all of our modern reception methods are based on the use of beat frequencies.

### *Shielding*

Two circuits that are physically near each other usually will be coupled to each other in some degree even though no coupling is intended. The metallic parts of the two circuits form a small capacitance through which energy can be transferred by means of the electric field. Also, the magnetic field about the coil or wiring of one circuit can couple that circuit to a second through the latter's coil and wiring. In many cases these unwanted couplings must be prevented if the circuits are to work properly.

Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers, called **shields.** The electric field from the circuit components does not penetrate the shield, because the lines of force are short-circuited by the metal. A metallic plate, called a **baffle shield,** inserted between two components also may suffice to prevent electrostatic coupling between them. Very little of the field tends to bend around such a shield if it is large enough to make the components invisible to each other.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. In this case the magnetic field induces a current in the shield; this current in turn sets up its own magnetic field opposing the original field. The amount of current induced is proportional to the frequency and also to the conductivity of the shield; therefore the shielding effect increases with frequency and with the conductivity and thickness of the shielding material.

A closed shield is required for good magnetic shielding; in some cases separate shields, one about each coil, may be required. The baffle shield is rather ineffective for magnetic shielding, although it will give partial shielding if placed at right angles to the axes of, as well as between, the two coils to be shielded from each other.

Shielding a coil reduces its inductance, because part of its field is canceled. Also, there is always a small amount of resistance in the shield, and there is therefore an energy loss. This loss raises the effective resistance of the coil. The decrease in inductance and increase in resistance lower the $Q$ of the coil. The reduction in inductance and $Q$ will be small if the shield is sufficiently far away from the coil; the spacing between the sides of the coil and the shield should be at least half the coil diameter, and the spacing at the ends of the coil should at least equal the coil diameter. The higher the conductivity of the shield material, the less the effect on the inductance and $Q$. Copper is the best material, but aluminum is quite satisfactory.

At low (audio) frequencies this type of magnetic shielding does not work, because the current induced in the shield is too small. For good shielding at audio frequencies it is necessary to enclose the coil in a container of high-permeability iron or steel. This provides a much better path for the magnetic flux than air — so much so that most of the stray flux stays in the iron in preference to spreading out in the space around the coil. In this case the shield can be quite close to the coil without harming its performance.

# Vacuum-Tube Principles

Present-day methods of radio communication rely heavily on the vacuum tube. The tube is used to generate radio-frequency power, to amplify it in transmitters, to amplify and detect weak radio signals picked up from distant stations, to magnify the human voice, to change alternating current into direct current for power supplies — in fact, to do innumerable things that, without it, could not be done. An understanding of vacuum-tube principles is just as necessary to the radio amateur as an understanding of the circuit principles discussed in Chapter Two.

In this chapter we shall confine ourselves to the *fundamentals* of vacuum-tube operation. The special circuits and special types of tubes that find application in amateur radio will be taken up in later chapters.

The operation of vacuum tubes can be predicted mathematically, just as the operation of circuits can be predicted from mathematical formulas. It happens, though, that the amateur rarely has need to perform any calculations in connection with vacuum tubes, other than simple ones having to do with the power supplies for the tube elements. These are straightforward applications of Ohm's Law. Tube manufacturers invariably supply sets of data that give optimum operating conditions for their tubes, and thus save any need for calculation. What you need, to get the most out of your tubes, is mostly a picture of how they work.

## Diodes and Rectification

### ● CURRENT IN A VACUUM

The outstanding difference between the vacuum tube and most other electrical devices is that the electric current does not flow through a conductor but through empty space — a vacuum. This is only possible when "free" electrons — that is, electrons that are not attached to atoms — are somehow introduced into the vacuum. It will be recalled from Chapter Two that electrons are particles of negative electricity. Free electrons in an evacuated space therefore can be attracted to a positively-charged object within the same space, or can be repelled by a negatively-charged object. The movement of the electrons under the attraction or repulsion of such charged objects constitutes the current in the vacuum.

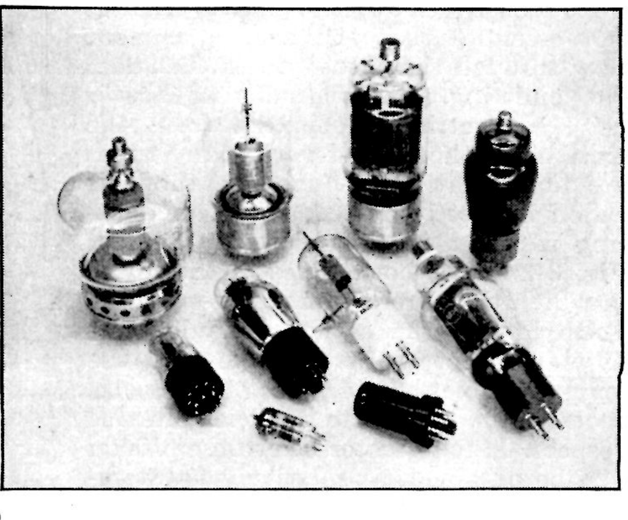
The most practical way to introduce a sufficiently-large number of electrons into the evacuated space is by **thermionic emission.**

#### *Thermionic Emission*

If a thin wire or filament is heated to incandescence in a vacuum, electrons near the surface are given enough energy of motion to fly off into the surrounding space. The higher the temperature, the greater the number of electrons **emitted.** A more general name for the filament is **cathode.**

If the cathode is the only thing in the vacuum, most of the emitted electrons stay in its immediate vicinity, forming a "cloud" about the cathode. The reason for this is that the electrons in the space, being negative electricity, form a negative charge **(space charge)** in the region of the cathode. The negatively-charged space repels those electrons nearest the cathode, tending to make them fall back on it.

Now suppose a second conductor is introduced into the vacuum, but not connected to anything else inside the tube. If this second conductor is given a positive charge with respect to the cathode, electrons in the space will be attracted to the positively-charged conductor. The conductor can be given the requisite charge by connecting a source of e.m.f. between it and the cathode, as indicated in Fig. 3-1. The electrons emitted by the cathode and attracted to the positively-



charged conductor then constitute an electric current, with the circuit completed through the source of e.m.f. In Fig. 3-1 this e.m.f. is supplied by a battery (**"B" battery**); a second battery (**"A" battery**) is also indicated for heating the cathode or filament to the proper operating temperature.

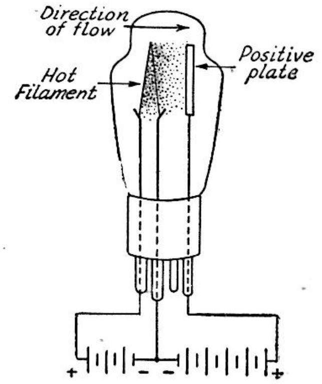



*Fig. 3-1* — **Conduction by thermionic emission in a vacuum tube. One battery is used to heat the filament to a temperature that will cause it to emit electrons. The other battery makes the plate positive with respect to the filament, thereby causing the emitted electrons to be attracted to the plate. Electrons captured by the plate flow back through the battery to the filament.**

The positively-charged conductor is usually a metal plate or cylinder (surrounding the cathode) and is called an **anode** or **plate.** Like the other working parts of a tube, it is a **tube element** or **electrode.** The tube shown in Fig. 3-1 is a **two-element** or **two-electrode** tube, one element being the cathode or filament and the other the anode or plate.

Since electrons are *negative* electricity, they will be attracted to the plate *only* when the plate is positive with respect to the cathode. If the plate is given a negative charge, the electrons will be repelled back to the cathode and no current will flow in the vacuum. The vacuum tube therefore can conduct *only in one direction.*

### *Cathodes*

Before electron emission can occur, the cathode must be heated to a high temperature. The only satisfactory way to heat it is by electricity. However, it is not essential that the heating current flow through the actual metal that does the emitting. The filament or heater can be electrically separate from the emitting cathode, and very many tubes are built that way. Such a cathode is called **indirectly heated,** while an emitting filament is called **directly heated.** Fig. 3-2 shows both types in the forms in which they are commonly used.

Obviously, the cathode should emit as many electrons as possible with the least possible heating power. A plain metal cathode is quite inefficient in this respect. Much greater electron emission can be obtained, at relatively low temperatures, by using special cathode materials. One of these is **thoriated tungsten,** or tungsten in which thorium is dissolved. Still greater efficiency is achieved in the **oxide-coated** cathode, a cathode in which rare-earth oxides form a coating over a metal base.

Although the oxide-coated cathode has much the highest efficiency, it can be used successfully only in tubes that operate at rather low plate voltages. Its use is therefore confined to receiving-type tubes and to the smaller varieties of transmitting tubes. The thoriated filament, on the other hand, will operate well in high-voltage tubes and is therefore found in most of the transmitting types used by amateurs.

### *Plate Current*

The number of electrons attracted to the plate depends upon the strength of the positive charge on the plate — that is, on the amount of voltage between the cathode and plate. The electron current — called the **plate current** — increases as the plate voltage is increased (although the relationship is not the simple proportionality of Ohm's Law). Actually, this statement is true only up to a certain point; if the plate voltage is made high enough, *all* the electrons emitted by the cathode would be attracted to the plate. Obviously, when this occurs, a further increase in plate voltage cannot cause an increase in plate current.

Fig. 3-3 shows a typical plot of plate current with increasing plate voltage for a two-element tube or **diode.** A curve of this type can be obtained with the circuit shown, if the plate voltage can be increased in small steps and a current reading taken (by means of the current-indicating instrument — a "milliammeter") at each voltage. The plate current is zero with no plate voltage and the curve rises almost in a straight line until a "saturation point" is reached. This is where the positive

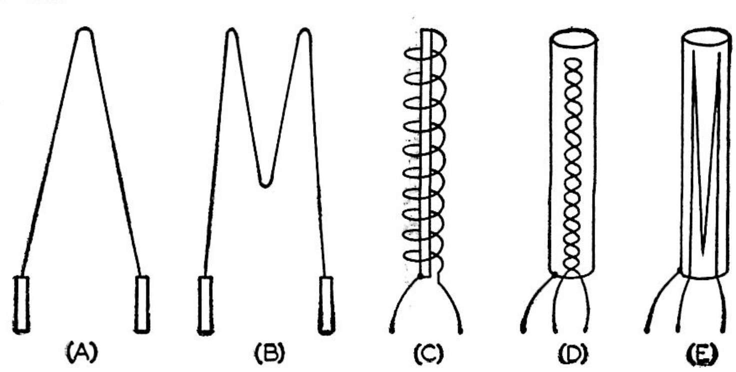



*Fig. 3-2* — **Types of cathode construction. Directly-heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly-heated cathodes at D and E show two types of heater construction, one a twisted loop and the other bunched heater wires. Both types tend to cancel the magnetic fields set up by the current through the heater.**

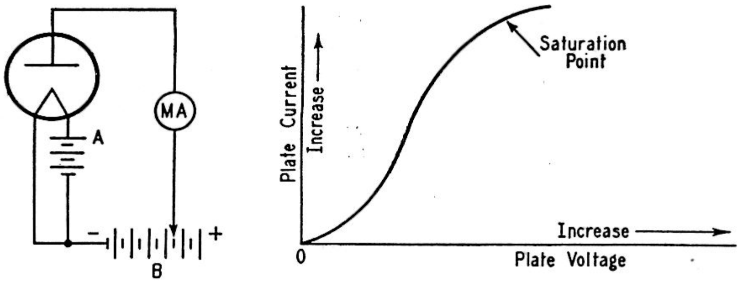
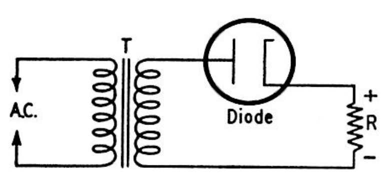



*Fig. 3-3* — The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

charge on the plate has completely overcome the space charge and practically all the electrons are going to the plate. At any higher voltages the plate current stays at the same value.

The curve of Fig. 3-3 does not show actual values of plate voltage and plate current, since these will vary with the type of tube. The *shape* of the curve, however, is typical of all diodes.

The plate voltage multiplied by the plate current is the *power input* to the tube. In a circuit like that of Fig. 3-3 this power is all used in heating the plate. If the power input is large, the plate temperature may rise to a very high value (the plate may become red or even white hot). The heat developed in the plate is radiated to the bulb of the tube, and in turn radiated by the bulb to the surrounding air.



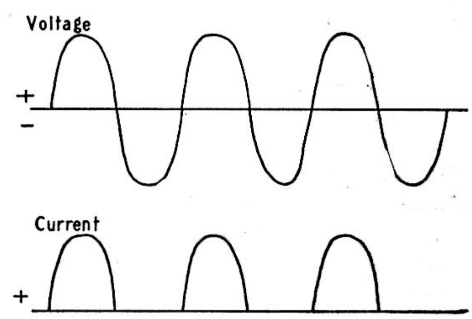




*Fig. 3-4* — Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.

## RECTIFICATION

Since current can flow through a tube in only one direction, a diode can be used to change alternating current into direct current. It does this by permitting current to flow when the plate is positive with respect to the cathode, but by shutting off current flow when the plate is negative.

Fig. 3-4 shows a representative circuit. Alternating voltage from the secondary of the transformer, *T*, is applied to the diode tube in series with a **load resistor,** *R*. The voltage varies as is usual with a.c., but current flows through the tube and *R only* when the plate is positive with respect to the cathode — that is, during the half-cycle when the upper end of the transformer winding is positive. During the negative half-cycle there is simply a gap in the current flow. This **rectified** alternating current therefore is an *intermittent* direct current. (The "humps" in the output current may be smoothed out by a "filter." A filter uses inductance and capacitance to store up energy during the time that current flows through the diode, energy that is then released to the circuit during the period when the diode is nonconducting. Filters of this type are discussed in later chapters.)

The load resistor, *R*, represents the actual circuit in which the rectified alternating current does work. All tubes work into a load of one type or another; in this respect a tube is much like a generator or transformer. A circuit that did not provide a load for the tube would be like a short-circuit across a transformer; no useful purpose would be accomplished and the only result would be the generation of heat in the transformer. So it is with vacuum tubes; they must *deliver* power to a load in order to serve a useful purpose. Also, to be *efficient* most of the power must do useful work in the load and not be used in heating the plate of the tube. This means that most of the voltage should appear as a drop across the load rather than as a drop between the plate and cathode of the diode. That is, the "resistance" of the tube should be small compared to the resistance of the load.

Notice that, with the diode connected as shown in Fig. 3-4, the polarity of the voltage drop across the load is such that the end of the load nearest the cathode is positive. If the connections to the diode elements are reversed, the direction of rectified current flow also will be reversed through the load.

# Vacuum-Tube Amplifiers

## TRIODES

### *Grid Control*

It was shown in Fig. 3-3 that, within the normal operating range of a tube, the plate current will increase when the plate voltage is increased. The reason why all the electrons are not drawn to the plate when a *small* positive voltage is placed on it is that the space charge (which is negative) counteracts the effect of the positive charge on the plate. The higher the positive plate voltage, the more

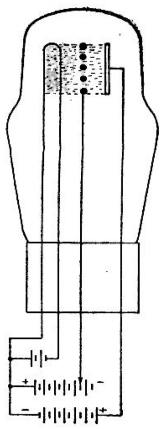


*Fig. 3-5* — **Construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.**

effectively the space charge is overcome.

If a third element — called the **control grid,** or simply **grid** — is inserted between the cathode and plate as in Fig. 3-5, it can be used to control the effect of the space charge. If the grid is given a positive voltage with respect to the cathode, the positive charge will tend to neutralize the negative space charge. The result is that, at any selected plate voltage, more electrons will flow to the plate than if the grid were not present. On the other hand, if the grid is made negative with respect to the cathode the negative charge on the grid will *add* to the space charge. This will *reduce* the number of electrons that can reach the plate at any selected plate voltage.

The grid is inserted in the tube to control the space charge and not to attract electrons to itself, so it is made in the form of a wire mesh or spiral. Electrons then can go through the open spaces in the grid and to the plate.

### *Characteristic Curves*

For any particular tube, the effect of the grid voltage on the plate current can be shown by a set of **characteristic curves.** A typical set of curves is shown in Fig. 3-6, together with the circuit that is used for getting them. With several fixed values of plate voltage (in these curves, the plate voltage is increased in 50-volt steps, starting at 100 volts) the grid voltage is varied in small steps and a plate-current reading taken at each value of grid voltage. The curves show the result. In Fig. 3-6, the grid voltage is varied between zero and 25 volts negative with respect to the cathode. It can be seen that, for each value of plate voltage, there is a value of negative grid voltage that will reduce the plate current to zero; that is, there is a value of negative grid voltage that will **cut off** the plate current.

The curves could be extended by making the grid voltage positive as well as negative. The practical effect would be to lengthen each of the curves upward along the same line. However, in some types of operation the grid is **always kept negative** with respect to the cathode, and the particular tube used as an illustration happens to be one that normally would be used that way. Whenever the grid is negative, it repels electrons and therefore none of them reaches it; in other words, no current flows in the grid circuit. When the grid is positive, it attracts electrons and a current **(grid current)** flows, just as current flows to the positive plate. Whenever there is grid current there is an accompanying power loss in the grid circuit, but so long as the grid is negative there is no current and therefore no power is used.

It is obvious that the grid can act as a valve to control the flow of plate current. Actually, the grid has a much greater effect on plate current flow than does the plate voltage. A *small* change in grid voltage is just as effective in bringing about a given change in plate current as is a *large* change in plate voltage.

The fact that a small voltage acting on the grid is equivalent to a large voltage acting on the plate indicates the possibility of **amplification** with the triode tube; that is, the generation of a large voltage by a small one, or the generation of a relatively large amount of power from a small amount. The many uses of the electronic tube nearly all are based upon this amplifying feature. The amplified power or voltage output from the tube is not obtained from the tube itself, but from the source of e.m.f. connected between its plate and cathode. The tube simply *controls* the power from this source, changing it to the desired form.

To utilize the controlled power, a load must be connected in the plate or "output" circuit, just as in the diode case. The load may be either a resistance or an impedance. The term "impedance" is frequently used even when the load is purely resistive.

### *Tube Characteristics*

The physical construction of a triode determines the relative effectiveness of the grid and plate in controlling the plate current. If a very small change in the grid voltage has just as much effect on the plate current as a very large change in plate voltage, the tube is said

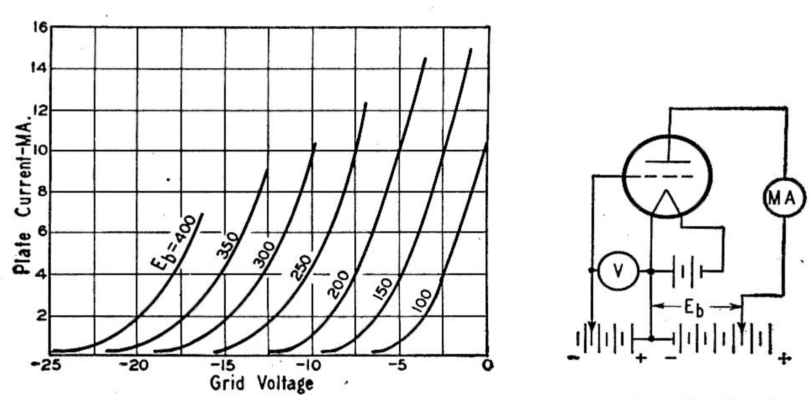



*Fig 3-6* — **Grid-voltage-*vs.*-plate-current curves at various fixed values of plate voltage ($E_b$) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.**

to have a high **amplification factor.** Amplification factor is commonly designated by the Greek letter $\mu$. An amplification factor of 20, for example, means this: if the grid voltage is changed by 1 volt, the effect on the plate current will be the same as when the plate voltage is changed by 20 volts. The amplification factors of triode tubes range from 3 to something of the order of 100. A **high-$\mu$** tube is one with an amplification factor of perhaps 30 or more; **medium-$\mu$** tubes have amplification factors in the approximate range 8 to 30, and low-$\mu$ tubes in the range below 7 or 8.

It would be natural to think that a tube that has a large $\mu$ would be the best amplifier, but such is not necessarily the case. If the $\mu$ is high it is difficult for the plate to attract large numbers of electrons. Quite a large change in the plate voltage must be made to effect a given change in plate current. This means that the resistance of the plate-cathode path — that is, the **plate resistance** — of the tube is high. Since this resistance acts in series with the load, the amount of current that can be made to flow through the load is relatively small. On the other hand, the plate resistance of a low-$\mu$ tube is relatively low. Whether or not a high-$\mu$ tube is better than one with a low $\mu$ depends on the operation we want the tube to perform.

The best all-around indication of the effectiveness of the tube as an amplifier is its **transconductance** — also called **mutual conductance.** This characteristic takes account of both amplification factor and plate resistance, and therefore is a sort of figure of merit for the tube. Actually, transconductance is the change in plate *current* divided by the change in grid *voltage* that causes the plate-current change (the plate voltage being fixed at a desired value). Since current divided by voltage is equal to conductance, transconductance is measured in the unit of conductance, the mho. Practical values of transconductance are very small, so the micromho (one-millionth of a mho) is the commonly-used unit. Different types of tubes have transconductances ranging from a few hundred to several thousand. The higher the transconductance the greater the possible amplification.

## AMPLIFICATION

To understand amplification, it is first necessary to become acquainted with a type of graph called the **dynamic characteristic.** Such a graph, together with the circuit used for obtaining it, is shown in Fig. 3-7. The curves are taken with the plate-supply voltage fixed at the desired operating value. The difference between this circuit and the one shown in Fig. 3-6 is that there is a load resistance connected in series with the plate of the tube in Fig. 3-7, while there is none in Fig. 3-6. Fig. 3-7 thus shows how the plate current will vary, with different grid voltages, *when the plate current is made to flow through a load and thus do useful work.*

The several curves in Fig. 3-7 are for various values of load resistance. The effect of the amount of load resistance is worth noting. When the resistance is small (as in the case of the 5000-ohm load) the plate current changes rather rapidly with a given change in grid voltage. If the load resistance is high (as in the 100,000-ohm curve), the change in plate current for the same grid-voltage change is relatively small, so the curve tends to be straighter.

Going now to Fig. 3-8, we have the same type of curve, but with the circuit arranged so that a source of alternating voltage **(signal)** is inserted between the grid and the grid battery **("C" battery).** The voltage of the grid battery is fixed at −5 volts, and from the curve

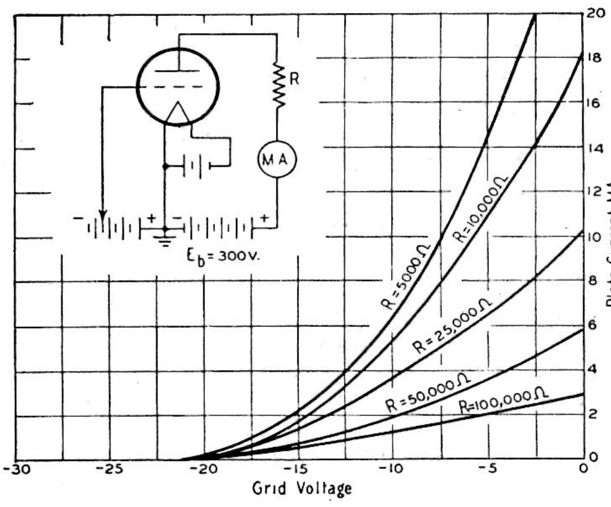



*Fig. 3-7* — **Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.**

it is seen that the plate current at this grid voltage is 2 milliamperes. This current flows when the load resistance is 50,000 ohms, as indicated in the circuit diagram. If there is no a.c. signal in the grid circuit, the voltage drop in the load resistor is $50{,}000 \times 0.002 = 100$ volts, leaving 200 volts between the plate and cathode.

Now when a sine-wave signal having a peak value of 2 volts is applied in series with the bias voltage in the grid circuit, the instantaneous voltage at the grid will swing to −3 volts at the instant the signal reaches its positive peak, and to −7 volts at the instant the signal reaches its negative peak. The maximum plate current will occur at the instant the grid voltage is −3 volts. As shown by the graph, it will have a value of 2.65 milliamperes. The minimum plate current occurs at the instant the grid voltage is −7 volts, and has a value of 1.35 ma. At intermediate values of grid voltage, intermediate plate-current values will occur.

The instantaneous voltage between the plate and cathode of the tube also is shown on the

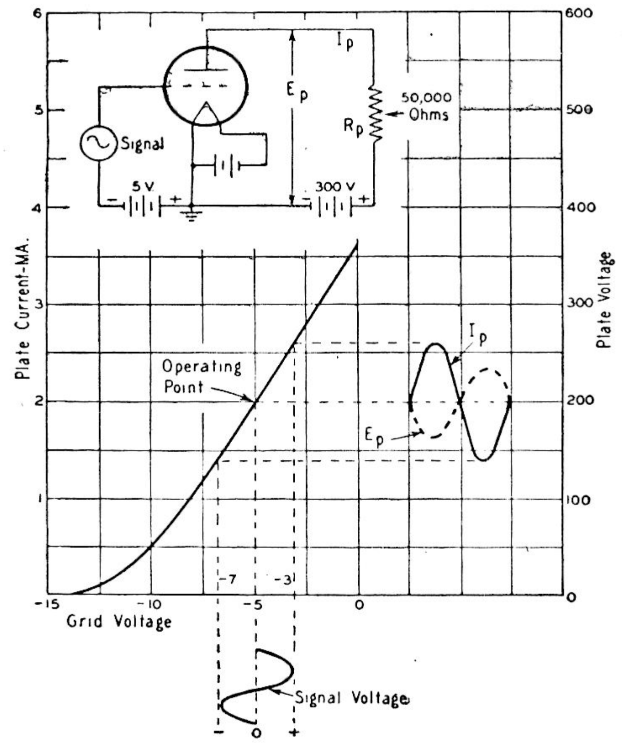



*Fig. 3-8* — Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, $R_p$, as shown by the dashed curve, $E_p$. $I_p$ is the plate current.

graph. When the plate current is maximum, the instantaneous voltage drop in $R_p$ is 50,000 $\times$ 0.00265 = 132.5 volts; when the plate current is minimum the instantaneous voltage drop in $R_p$ is 50,000 $\times$ 0.00135 = 67.5 volts. The actual voltage between plate and cathode is the difference between the plate-supply potential, 300 volts, and the voltage drop in the load resistance. The plate-to-cathode voltage is therefore 167.5 volts at maximum plate current and 232.5 volts at minimum plate current.

This varying plate voltage is an **a.c.** voltage superimposed on the steady plate-cathode potential of 200 volts (as previously determined for no-signal conditions). The peak value of this a.c. output voltage is the difference between either the maximum or minimum plate-cathode voltage and the no-signal value of 200 volts. In the illustration this difference is 232.5 − 200 or 200 − 167.5; that is, 32.5 volts in either case. Since the grid signal voltage has a peak value of 2 volts, the **voltage-amplification ratio** of the amplifier is 32.5/2 or 16.25. That is, approximately 16 times as much voltage is obtained from the plate circuit as is applied to the grid circuit.

One feature of the alternating component of plate voltage is worth special note. As shown by the drawings in Fig. 3-8, the positive swing in the grid signal voltage is accompanied by a *downward* swing in the voltage ($E_p$) between the plate and cathode of the tube. Also, when the alternating grid voltage swings in the *negative* direction, the plate-to-cathode voltage swings to a *higher* value. In other words, the alternating component of the plate voltage swings in the *negative* direction (with reference to the no-signal value of plate-cathode voltage) when the grid swings in the *positive* direction, and vice versa. This means that the alternating component of plate voltage (that is, the amplified signal) is 180 degrees out of phase with the signal voltage on the grid.

### *Bias*

The fixed negative grid voltage (called **grid bias**) in Fig. 3-8 serves a very useful purpose. In the first place, one of the things we want to do in the type of amplification shown in this drawing is to obtain, from the plate circuit, an alternating voltage that has the *same waveshape* as the signal voltage applied to the grid. To do so, we must choose an operating point on the *straight* part of the curve; not only that, the curve must be straight in both directions from the operating point at least far enough to accommodate the maximum value of the signal applied to the grid. If the grid signal swings the plate current back and forth, over a part of the curve that is not straight, as in Fig. 3-9, the shape of the a.c. wave in the plate circuit will not be the same as the shape of the grid-signal wave. In such a case the output waveshape will be **distorted.**

The second reason for using negative grid bias is this: The grid will not attract electrons — that is, there will be no grid current — if the grid is always negative with respect to the cathode. When the grid has a negative bias, any signal whose peak *positive* voltage does not exceed the fixed *negative* voltage on the grid cannot cause grid current to flow. With no current flow there is no power consumption, so the tube will amplify *without taking any power from the signal source.* However, if the positive

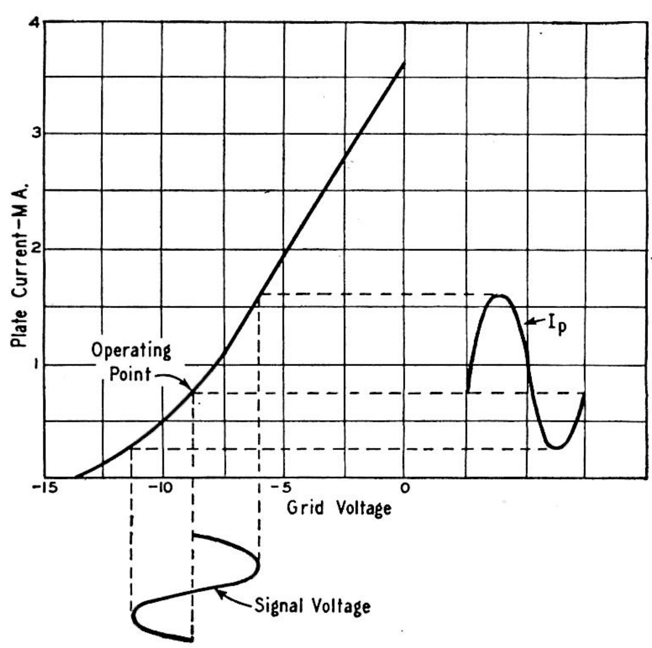



*Fig. 3-9* — Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the upper half-cycle.

peak of the signal does exceed the negative bias, current will flow in the grid circuit during the time the grid is positive. While it is perfectly possible to operate the tube in the "positive-grid region," in many cases we do not want the grid to consume power.

Distortion of the output waveshape that results from working over a part of the curve that is not straight (that is, a **nonlinear** part of the curve) has the effect of transforming a sine-wave grid signal into a more complex waveform. As explained in Chapter Two, a complex wave can be resolved into a fundamental and a series of harmonics. In other words, distortion from nonlinearity causes the generation of harmonic frequencies — frequencies that are not present in the signal applied to the grid. Harmonic distortion is undesirable in most amplifiers, although there are occasions when harmonics are deliberately generated and used. This is particularly so in certain types of r.f. transmitting circuits.

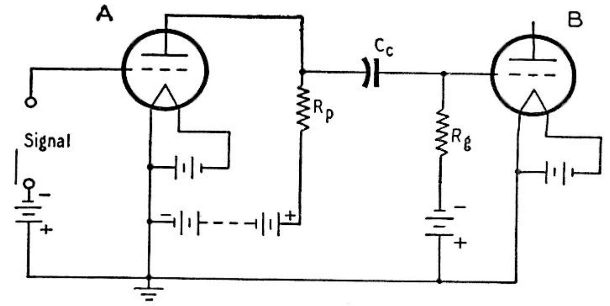
### *Amplifier Output Circuits*

The thing that is wanted from the output circuit of a vacuum-tube amplifier is the *alternating* component of plate current or plate voltage. The d.c. voltage on the plate of the tube is essential, of course, for the tube's operation. However, it almost invariably would cause difficulties if it were applied, along with the a.c. output voltage, to the load. The output circuits of vacuum tubes are therefore arranged so that the a.c. is transferred to the load but the d.c. is not.

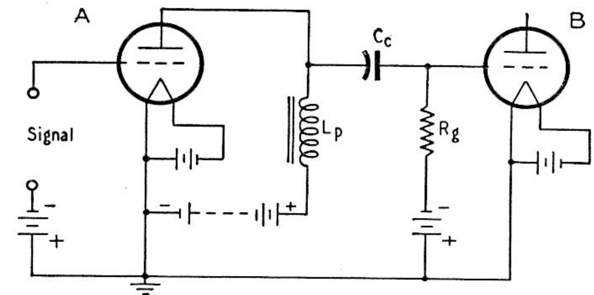
Three types of coupling are in common use at audio frequencies. These are **resistance coupling, impedance coupling, and transformer coupling.** They are shown in Fig. 3-10. In all three cases the output is shown coupled to the grid circuit of a subsequent amplifier tube, but the same types of circuits can be used to couple to other devices than tubes.

In the resistance-coupled circuit, the a.c. voltage developed across the plate resistor $R_p$ (that is, between the plate and cathode of the tube) is applied to a second resistor, $R_g$, through a coupling condenser, $C_c$. The condenser "blocks off" the voltage on the plate of the first tube and prevents it from being applied to the grid of tube *B*. The latter tube should have negative grid bias, of course, and this is supplied by the battery shown. No current flows in the grid circuit of tube *B* and there is therefore no d.c. voltage drop in $R_g$; in other words, the full voltage of the bias battery is applied to the grid of tube *B*.

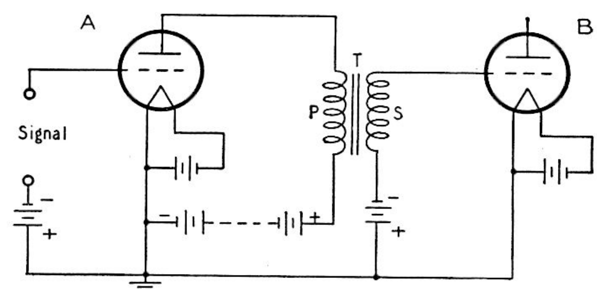
The grid resistor, $R_g$, usually has a rather high value (0.5 to 2 megohms). The reactance of the coupling condenser, $C_c$, must be low enough compared to the resistance of $R_g$ so that the a.c. voltage drop in $C_c$ is negligible at the lowest frequency to be amplified. If $R_g$ is at least 0.5 megohm, a 0.1-$\mu$fd. condenser will be amply large for the usual range of audio frequencies.




RESISTANCE COUPLING




IMPEDANCE COUPLING




TRANSFORMER COUPLING

*Fig. 3-10* — **Three basic forms of coupling between vacuum-tube amplifiers.**

So far as the alternating component of plate voltage is concerned, it will be realized that if the voltage drop in $C_c$ is negligible then $R_p$ and $R_g$ are effectively in parallel (although they are quite separate so far as d.c. is concerned). The resultant parallel resistance of the two is therefore the actual load resistance for the tube. That is why $R_g$ is made as high in resistance as possible; then it will have the least effect on the load represented by $R_p$.

The impedance-coupled circuit differs from that using resistance coupling only in the substitution of a high-inductance coil (usually several hundred henrys) for the plate resistor. The advantage of using an inductance rather than a resistor is that its impedance is high for alternating currents, but its resistance is relatively low for d.c. (A resistor, of course, has the same resistance for d.c. that it does for a.c.). It thus permits us to obtain a high value of load impedance for a.c., but without an excessive d.c. voltage drop that would use up a good deal of the voltage from the plate supply.

The transformer-coupled amplifier uses a transformer with its primary connected in the plate circuit of the tube and its secondary connected to the load (in the circuit shown, a following amplifier). There is no direct connection between the two windings, so the plate voltage on tube *A* is isolated from the grid of tube *B*. The transformer-coupled amplifier has the same advantage as the impedance-coupled circuit with respect to loss of voltage from the plate supply. There is an additional advantage as well: if the secondary has more turns than the primary, the output voltage will be "stepped up" in proportion to the turns ratio.

All three circuits have good points. Resistance coupling is simple, inexpensive, and will give the same amount of amplification — or **voltage gain** — over a wide range of frequencies; it will give substantially the same amplification at any frequency in the audio range, for example. Impedance coupling will give somewhat more gain, with the same tube and same plate-supply voltage, than resistance coupling. However, it is not quite so good over a wide frequency range; it tends to "peak," or give maximum gain, over a comparatively narrow band of frequencies. With a good transformer the gain of a transformer-coupled amplifier can be kept fairly constant over the audio-frequency range. On the other hand, transformer coupling is best suited to triodes having amplification factors of about 10 or less, for the reason that the primary inductance of a practicable transformer cannot be made large enough to work well with a tube having high plate resistance.

### *Voltage and Power Amplifiers*

In the preceding discussion of amplifiers we have talked chiefly about voltage gain. That is only part of the picture. The end result of any amplification is that the amplified signal does some *work*. As a familiar example, an audio-frequency amplifier usually drives a loudspeaker that in turn produces sound waves. The greater the amount of a.f. *power* supplied to the 'speaker, the louder the sound it will produce.

In some amplifiers, therefore, *power* output rather than voltage is the primary consideration. A tube is operated somewhat differently when we want the greatest possible power than it is when we want the largest possible voltage. It is chiefly a matter of the resistance of the load. It was mentioned in Chapter Two that any source of power will deliver the largest possible output when the resistance of the load is equal to the internal resistance of the source. In the case of a vacuum tube, the "source" resistance is the plate resistance of the tube. Therefore if we want the utmost power from the tube the load resistance should be equal to the plate resistance of the tube. Actually, however, this is not the best operating condition because the use of such a relatively low value of load resistance generally results in more distortion than we want.

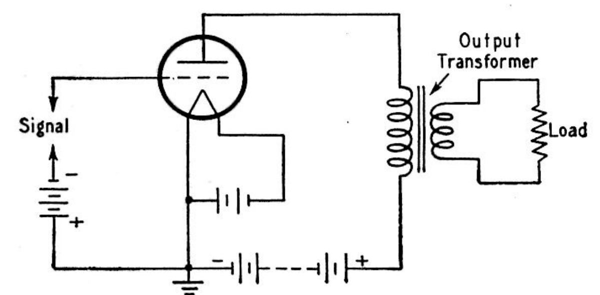



*Fig. 3-11* — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

For this reason the load resistance for a power amplifier usually is two or three times the plate resistance; this represents a good compromise between distortion and power output.

Fig. 3-11 shows an elementary power-amplifier circuit. It is simply a transformer-coupled amplifier with the load connected to the secondary. Although the load is shown as a resistor, it actually would be some device, such as a loudspeaker, that employs the power usefully. The resistance of the actual load is rarely the right value for "matching" the load resistance that the tube wants for optimum power output. Therefore the transformer turns ratio is chosen to reflect the proper value of resistance into the primary. The turns ratio may be either step-up or step-down, depending on whether the actual load resistance is higher or lower than the load the tube wants.

About the only time that voltage rather than power is wanted is when a tube is working into a following amplifier, as in the circuits given in Fig. 3-10. Even then, this is only so when the grid of the second tube always operates in the negative-voltage region. If the signal applied to its grid is large enough to overcome the bias and make grid current flow, power is consumed and special methods are required for transferring maximum power rather than maximum voltage. There are, nevertheless, many useful applications for tube circuits intended to deliver maximum voltage rather than maximum power. The chief difference is that the load resistance is made as high as possible for a voltage amplifier. This means that of the total voltage generated (part of which is lost in the plate resistance of the tube), the major portion will appear across the load resistance if the latter is *large* compared to the tube resistance. The amount of power that can be obtained under these conditions is small, but the voltage output is large.

Resistance- and impedance-coupled amplifiers are primarily voltage amplifiers, because neither type is capable of transferring power efficiently. For efficient power transfer the transformer is a practical necessity.

The best triodes for power amplification are those having low and medium values of amplification factor. High-$\mu$ tubes make the best voltage amplifiers, but have relatively little power output. These statements are true principally for amplifiers operating without grid current, which is the only type we have considered so far. Other types of operation are taken up in later chapters. A tube operated in the fashion we have described is called a **Class A** amplifier.

The **power-amplification ratio** of an amplifier is the ratio of the power output obtained from the plate circuit to the power required from the a.c. signal in the grid circuit. There is no power lost in the grid circuit of a Class A amplifier that is operating without grid current, so such an amplifier has an infinitely large power-amplification ratio. Another term used in connection with power amplifiers is **power sensitivity.** It means the ratio of the power output to the grid signal voltage that causes it. A tube with high power sensitivity is one that will give a large power output with a relatively small signal voltage on its grid.

### *Parallel and Push-Pull*

When it is necessary to obtain more power output than one tube is capable of giving, two or more tubes may be connected in **parallel.** In this case the similar elements in all tubes are connected together. This method is shown in Fig. 3-12 for a transformer-coupled amplifier. The power output is in proportion to the number of tubes used; the grid signal or "exciting" voltage required, however, is the same as for one tube.

If the amplifier operates in such a way as to consume power in the grid circuit, the grid power required also is in proportion to the number of tubes used.

An increase in power output also can be secured by connecting two tubes in **push-pull.** In this case the grids and plates of the two tubes are connected to opposite ends of a balanced circuit as shown in Fig. 3-12. At any

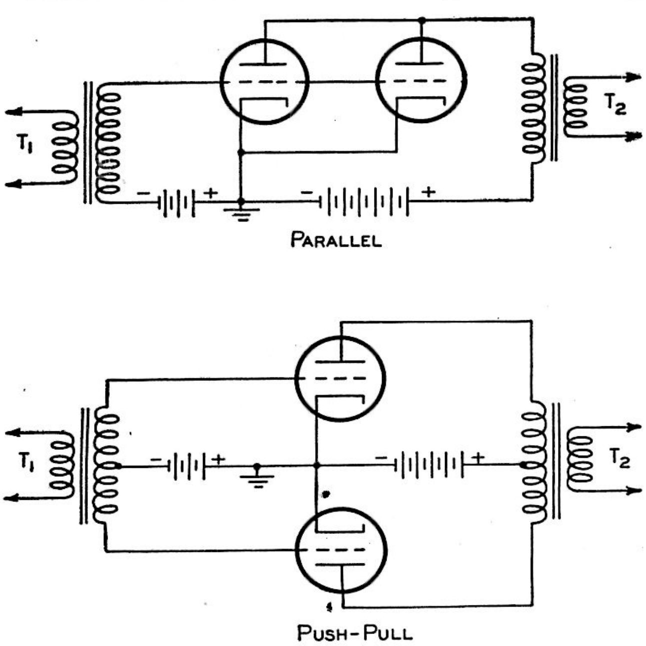



*Fig. 3-12* — Parallel and push-pull a.f. amplifier circuits.

instant the ends of the secondary winding of the input transformer, $T_1$, will be at opposite polarity with respect to the cathode connection, so the grid of one tube is swung positive at the same instant that the grid of the other is swung negative. Hence, in any push-pull-connected amplifier the voltages and currents of one tube are out of phase with those of the other tube.

In push-pull operation the even-harmonic (second, fourth, etc.) distortion is balanced out in the plate circuit. This means that for the same power output the distortion will be less than with parallel operation.

The exciting voltage measured between the two grids must be twice that required for one tube. If the grids consume power, the driving power for the push-pull amplifier is twice that taken by either tube alone.

### *Cascade Amplifiers*

It is of course thoroughly possible to take the output of one amplifier and apply it as a signal on the grid of a second amplifier, then take the second amplifier's output and apply it to a third, and so on. Each amplifier is called **a stage,** and a number of amplifier stages used to successively increase the amplitude of the signal are said to be in **cascade.**

The number of amplifiers that can be connected in cascade is not unlimited. If the over-all amplification becomes too great, there is danger that some of the output voltage will get back into one of the early stages. This "feed-back," discussed in the next section, may make the amplifier unstable and prevent it from functioning as it should.

## FEED-BACK

As we have shown, there is more energy in the plate circuit of an amplifier than there is in the grid circuit. It is easily possible to take a part of the plate-circuit energy and insert it into the grid circuit. When this is done the amplifier is said to have **feed-back.**

There are two types of feed-back. If the voltage that is inserted in the grid circuit is 180 degrees out of phase with the signal voltage acting on the grid, the feed-back is called **negative,** or **degenerative.** On the other hand, if the voltage is fed back *in* phase with the grid signal, the feed-back is called **positive,** or **regenerative.** With negative feed-back the voltage that is fed back *opposes* the signal voltage; this decreases the amplitude of the voltage acting between the grid and cathode. With a smaller signal voltage, of course, the output also is smaller. The effect of negative feed-back, then, is to *reduce* the amount of amplification.

### *Negative Feed-Back*

The circuit shown at A in Fig. 3-13 gives degenerative feed-back. Resistor $R_c$ is in series with the regular plate resistor, $R_p$, and

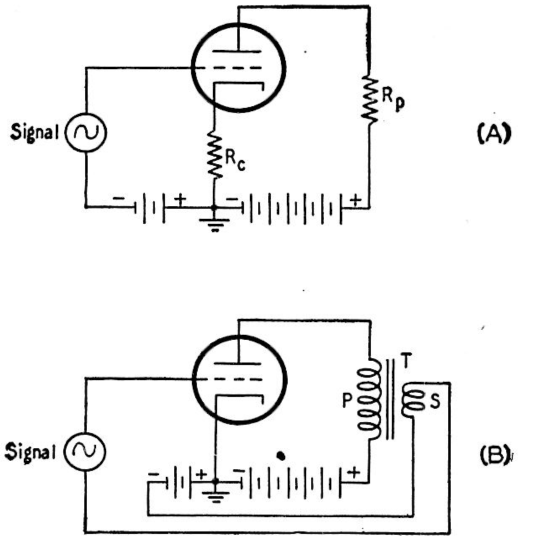



*Fig. 3-13* — Circuits for producing feed-back. In A, part of the a.c. plate voltage appears across the cathode resistor, $R_c$, and is therefore also applied between grid and cathode. The feed-back is negative in this case. In B, the voltage that is generated in the secondary of the transformer is inserted in series in the grid circuit. Feed-back may be either positive or negative, depending upon the transformer connections.

thus is a part of the load for the tube. Therefore, part of the output voltage will appear across $R_c$. However, $R_c$ also is connected in series with the *grid* circuit, and so the output voltage that appears across $R_c$ is in series with the signal voltage. In this circuit, the output voltage across $R_c$ opposes the signal voltage and the actual a.c. voltage between the grid and cathode therefore is equal to the *difference* between the two voltages.

While it would be natural to assume that there could be no point in reducing the amplification by negative feed-back, it does have uses. The greater the amount of negative feed-back (when properly applied) the more independent the amplification becomes of tube characteristics and circuit conditions. This means that the frequency-response characteristic of the amplifier becomes **flat** — that is, amplification tends to be the same at all frequencies within the range for which the amplifier is designed. Also, any distortion generated in the plate circuit of the tube tends to "buck itself out" when some of the output voltage is fed back to the grid. Amplifiers with negative feed-back are therefore comparatively free of harmonic distortion. These advantages, secured at the expense of voltage amplification, are worth while if the amplifier otherwise has enough gain for its intended use.

The circuit shown at B in Fig. 3-13 can be used to give either negative or positive feed-back. In this case the secondary of a transformer is connected back into the grid circuit to insert a desired amount of feed-back voltage. Reversing the terminals of either the primary or secondary of the transformer (but not both windings simultaneously) will reverse the phase of the voltage fed back. Thus either type of feed-back is available.

### *Positive Feed-Back*

Positive feed-back *increases* the amplification because the fed-back voltage adds to the original signal voltage and the resulting larger voltage on the grid causes a larger output voltage. It has the opposite characteristics to negative feed-back; the amplification tends to be greatest at one frequency (depending upon the particular circuit arrangement) and harmonic distortion is increased. If the energy fed back becomes large enough, a self-sustaining **oscillation** will be set up at one frequency; in this case *all* the signal voltage on the grid is supplied from the plate circuit; no external signal is needed. It is not even necessary to have an external signal to *start* the oscillation; any small irregularity in the plate current — and there are always some such irregularities — will be amplified and thus give the oscillation an opportunity to build up. Oscillations obviously would be undesirable in an audio-frequency amplifier, and for that reason (as well as the others mentioned above) positive feed-back is never used in a.f. amplifiers. Positive feed-back finds its use in "oscillators" at both audio and radio frequencies, as described in a subsequent section.

The two circuits shown in Fig. 3-13 are only two of many that can be used to provide feed-back. Despite differences in appearance, such circuits are alike in this fundamental — energy is fed back from the output circuit to the grid circuit in the proper phase to give the type of feed-back that is wanted.

## INTERELECTRODE CAPACITANCES

Each pair of elements in a tube actually forms a small "condenser," with each element acting as a condenser "plate." There are three such capacitances in a triode — that between the grid and cathode, that between the grid and plate, and that between the plate and cathode. The capacitances are very small — only a few micromicrofarads at most — but they frequently have a very pronounced effect on the operation of an amplifier circuit.

### *Input Capacitance*

It was explained previously that the a.c. grid voltage and a.c. plate voltage of an amplifier are 180 degrees out of phase, using the cathode of the tube as a reference point. However, these two voltages are *in* phase if we go around the circuit from plate to grid as shown in Fig. 3-14. This means that their sum is acting between the grid and plate; that is, across the grid-plate capacitance of the tube. When an a.c. voltage is applied to a condenser, a current flows through the condenser. As viewed from the source of the signal on the grid, this current is flowing because of the signal voltage.

The larger the current, the lower the effective reactance in the grid circuit. The larger the grid-plate capacitance the larger the current;

also, the greater the voltage amplification the larger the current, because this puts more voltage across the grid-plate condenser. The result is that the source of signal "sees" a capacitive reactance that is much smaller than the actual reactance of the capacitance between the grid and cathode.

Since a small reactance is equivalent to a large capacitance, the input capacitance of an amplifier may be many times its actual grid-cathode capacitance. In practice, the input capacitance of a triode may be as much as a few hundred micromicrofarads, particularly if the triode has a large amplification factor. Such a capacitance is not negligible, even at audio frequencies, when it is placed in parallel with a resistor of 50,000 ohms or more.

### *Tube Capacitance at R.F.*

At radio frequencies the reactances of the interelectrode capacitances drop to such low values that they must always be taken into account in circuit design. A resistance-coupled amplifier cannot be used at r.f., for example, because the reactances of the interelectrode "condensers" are so low that they, and not the resistors, would be the actual load. Furthermore, they are so low that they practically short-circuit the input and output circuits and thus the tube is unable to amplify. We get around this at radio frequencies by using *tuned* circuits for the grid and plate, and making the tube capacitances part of the tuning capacitances. In this way the circuits can have the high impedances necessary for satisfactory amplification.

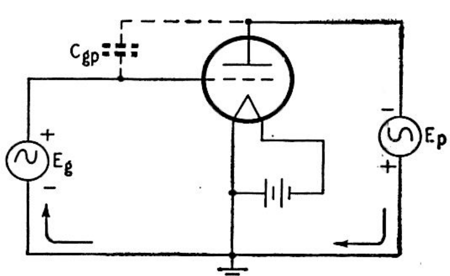



*Fig. 3-14* — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

The grid-plate capacitance is important at radio frequencies because it is, in effect, a coupling condenser between the grid and plate circuits. Since its reactance is relatively low at r.f., it offers a path over which energy can be fed back from the plate to the grid. In practically every case the feed-back is in the right phase and of sufficient amplitude to cause oscillation, so the amplifier becomes useless. Special circuits can be used to prevent feed-back but they are, in general, not too satisfactory when used in radio receivers. (They are, however, widely used in transmitters.) A better solution to this problem is found in the use of the screen-grid tube.

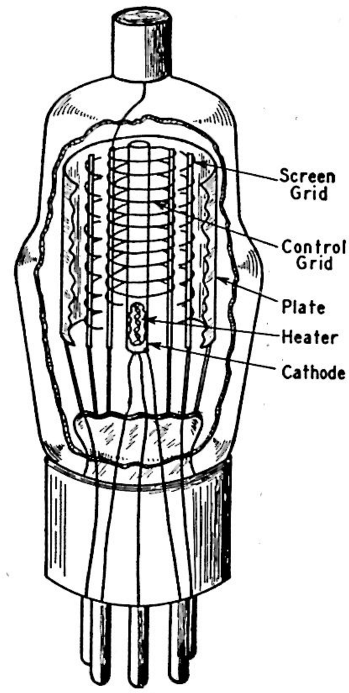



*Fig. 3-15* — Representative arrangement of elements in a screen-grid tube, with front part of plate and screen grid cut away. In this drawing the control-grid connection is made through a cap on the top of the tube, thus eliminating the capacitance that would exist between the plate-and grid-lead wires if both passed through the base. Some modern tubes that have both leads going through the base use special shielding and construction to eliminate interlead capacitance.

## SCREEN-GRID TUBES

The grid-plate capacitance can be eliminated — or at least reduced to a negligible value — by inserting a second grid between the control grid and the plate, as indicated in Fig. 3-15. The second grid, called the **screen grid,** acts as a shield between the control grid and plate. It is made in the form of a grid or coarse screen so that electrons can pass through it; a solid shield would entirely prevent the flow of plate current. The screen grid is usually grounded through a by-pass condenser that has low reactance at the radio frequency being amplified.

Because of the shielding action of the screen grid, the plate voltage cannot control the flow of plate current as it does in a triode. In order to get electrons to the plate, it is necessary to apply a positive voltage (with respect to the cathode) to the screen. The screen then attracts electrons much as does the plate in a triode tube. In traveling toward the screen the electrons acquire such velocity that most of them shoot between the screen wires and go on to the plate. A certain proportion do strike the screen, however, with the result that some current also flows to the screen-grid circuit of the tube.

A tube having a cathode, control grid, screen grid and plate (four elements) is called a **tetrode.**

### *Pentodes*

When an electron traveling at appreciable velocity through a tube strikes the plate it dislodges other electrons which "splash" from the plate into the interelement space. This is called **secondary emission.** In a triode the negative grid repels the secondary electrons back into the plate and they cause no disturbance. In

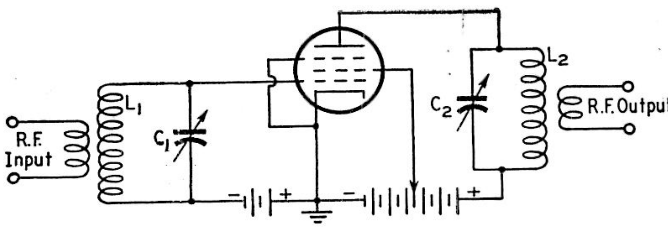



*Fig. 3-16* — **Simplified pentode r.f.-amplifier circuit. $L_1C_1$ and $L_2C_2$ are tuned to the same frequency.**

the screen-grid tube, however, the positively-charged screen *attracts* the secondary electrons, causing a reverse current to flow between screen and plate.

To overcome the effects of secondary emission, a third grid, called the **suppressor grid,** may be inserted between the screen and plate. This grid, which usually is connected directly to the cathode, repels the relatively low-velocity secondary electrons. They are driven back to the plate without appreciably obstructing the regular plate-current flow. A five-element tube of this type is called a **pentode.**

Although the screen grid in either the tetrode or pentode greatly reduces the influence of the plate upon plate-current flow, the control grid still can control the plate current in essentially the same way that it does in a triode. Consequently, the grid-plate transconductance (or mutual conductance) of a tetrode or pentode will be of the same order of value as in a triode of corresponding structure. On the other hand, since the plate voltage has very little effect on the plate-current flow, both the amplification factor and plate resistance of a pentode or tetrode are very high. In small receiving pentodes the amplification factor is of the order of 1000 or higher, while the plate resistance may be from 0.5 to 1 or more megohms. Because of the high plate resistance, the actual voltage amplification possible with a pentode is very much less than the large amplification factor might indicate. A voltage gain in the vicinity of 50 to 200 is typical of a pentode stage.

### *Pentode R.F. Amplifier*

Fig. 3-16 shows a simplified form of r.f. amplifier circuit, using a pentode tube. Radio-frequency energy in the small coil coupled to $L_1$ is built up in voltage in the tuned circuit, $L_1C_1$, when $L_1C_1$ is tuned to resonance with the frequency of the incoming signal. The voltage that appears across $L_1C_1$ is applied to the grid and cathode of the tube and is amplified by the tube. A second resonant circuit, $L_2C_2$, is the load for the plate of the tube, its parallel impedance being high because it is tuned to resonance with the frequency applied to the grid. R.f. output can be taken from the coil coupled to $L_2$. The screen-grid voltage is obtained from a tap on the plate battery; most tubes are designed for operation with the screen voltage considerably lower than the plate voltage. In this circuit the batteries are assumed to have low impedance for the r.f. current; in a practical circuit, by-pass condensers would be used to make sure that the impedances of the return paths actually are low enough to be negligible.

In addition to their applications as radio-frequency amplifiers, pentode or tetrode screen-grid tubes also can be constructed for audio-frequency power amplification. In tubes designed for this purpose the shielding effect of the screen grid is not so important; the chief function of the screen is to serve as an accelerator of the electrons, so that large values of plate current can be drawn at relatively low plate voltages. Such tubes have quite high power sensitivity compared to triodes of the same power output. Harmonic distortion is somewhat greater with pentodes and tetrodes than with triodes, however.

### *Variable-μ Tubes*

The mutual conductance of a vacuum tube decreases with increasing negative grid bias, assuming that the other electrode voltages are held constant. Since the mutual conductance controls the amount of amplification, it is possible to adjust the gain of the amplifier by adjusting the grid bias. This method of gain control is universally used in radio-frequency amplifiers designed for receivers. Some means of controlling the r.f. gain is essential in a receiver having a number of amplifiers, because of the wide range in the strengths of the incoming signals.

The ordinary type of tube has what is known as a **sharp cut-off** characteristic. The mutual conductance decreases at a uniform rate as the

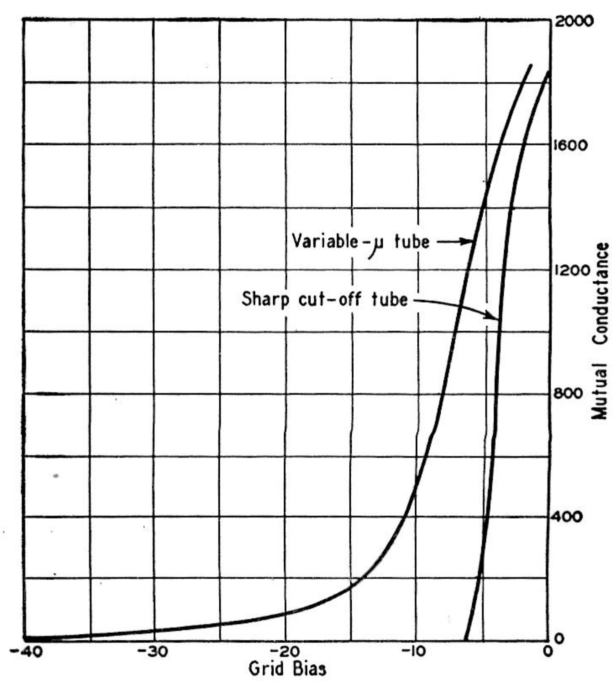


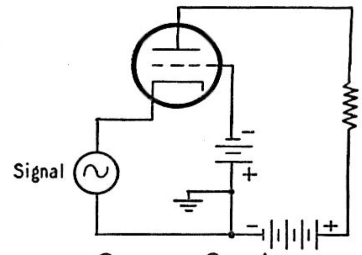

*Fig. 3-17* — **Curves showing the relationship between mutual conductance and negative grid bias for two small receiving pentodes, one a sharp cut-off type and the other a variable-μ type.**

negative bias is increased, as shown in Fig. 3-17. The amount of signal voltage that such a tube can handle without causing distortion is quite limited, and not sufficient to take care of very strong signals. To overcome this, some tubes are made with a **variable-$\mu$** characteristic (that is, the amplification factor changes with the grid bias), resulting in the type of curve shown in Fig. 3-17. It is evident that the variable-$\mu$ tube can handle a much larger signal than the sharp cut-off type before the signal swings either beyond the zero grid-bias point or the plate-current cut-off point.

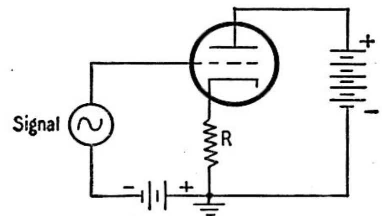
## OTHER TYPES OF AMPLIFIERS

In the amplifier circuits so far discussed, the signal has been applied between the grid and cathode and the amplified output has been taken from the plate-to-cathode circuit. That is, the *cathode* has been the common point, or meeting point, for the input and output circuits. However, since there are three elements (the screen and suppressor in a pentode ordinarily do not enter *directly* into the amplifying action) it is possible to use any one of the three as the common point. This leads to two different kinds of amplifiers, commonly called the **grounded-grid amplifier** (or **grid-separation** circuit) and the **cathode follower.**

These two circuits are shown in simplified form in Fig. 3-18. In both circuits the resistor *R* represents the load into which the amplifier works; the actual load may be resistance-capacitance-coupled, transformer-coupled, may be a tuned circuit if the amplifier operates at radio frequencies, and so on. Also, in both circuits the batteries that supply grid bias and plate power are assumed to have such negligible impedance that they do not enter into the operation of the circuits.



GROUNDED-GRID AMPLIFIER



CATHODE FOLLOWER

*Fig. 3-18* — In the upper circuit, the grid is the junction point between the input and output circuits. In the lower drawing, the plate is the junction. In either case the out put is developed in the load resistor, *R*, and may be coupled to a following amplifier by the usual methods.

### *Grounded-Grid Amplifier*

In the grounded-grid amplifier the input signal is applied between the cathode and grid, and the output is taken between the plate and grid. The grid is thus the common element. The plate current (including the a.c. component) has to flow through the signal source to reach the cathode. Since this source always has appreciable impedance, the alternating plate current causes a voltage drop that acts between the grid and cathode. Because of the phase relationship between the signal and output voltages, the circuit is degenerative. Also, since the source of signal is in series with the load through the plate-to-cathode resistance of the tube, some of the power in the load is supplied by the signal source. The result of these two things is that the signal source is called upon to furnish a considerable amount of power.

The grounded-grid amplifier finds its chief application at v.h.f. and u.h.f., where the more conventional amplifier circuit fails to work properly. With a triode tube designed for this type of operation, an r.f. amplifier can be built that is free from the type of feed-back that causes oscillation. This requires that the grid act as a shield between the cathode and plate, reducing the plate-cathode capacitance to a very low value.

### *Cathode Follower*

The cathode follower uses the plate of the tube as the common element. The input signal is applied between the grid and plate (assuming negligible impedance in the batteries) and the output is taken from between cathode and plate. This circuit, like the grounded-grid amplifier, is degenerative. In fact, *all* of the output voltage is fed back into the input circuit to buck the applied signal. The input signal therefore has to be larger than the output voltage; in other words, the cathode follower not only gives no voltage gain but actually results in a *loss* in voltage. (It can still give just as much *power* gain as ever, though.)

The cathode follower has two advantages: It has a very high input impedance (impedance between grid and ground—in the customary cathode-follower circuit the plate is at ground for signal voltage); and its output impedance is very low. (The large amount of negative feed-back has the effect of greatly reducing the plate resistance of the tube.) These two characteristics are valuable in an amplifier that must work over a very wide range of frequencies. Also, the high input impedance and low output impedance can be used to obtain an impedance step-down over wide ranges of frequencies that could not possibly be covered by a transformer. The cathode follower is useful both at audio and radio frequencies.

## CATHODE CIRCUITS AND GRID BIAS

Most of the equipment used by amateurs is powered by the a.c. line. This includes the filaments or heaters of vacuum tubes. Although supplies for the plate (and sometimes the grid) are usually rectified and filtered to give "pure" d.c. — that is, direct current that is constant and without a superimposed a.c. component — the relatively large currents required by filaments and heaters make a d.c. supply impracticable.

### *Filament Hum*

Alternating current is just as good as direct current from the heating standpoint, but some of the a.c. voltage is likely to get on the grid and cause a low-pitched "a.c. hum" to be superimposed on the output. The voltage can get on the grid either by a direct circuit connection, through the electric field about the heater, or through the magnetic field set up by the current.

Hum troubles are worst with directly-heated cathodes or filaments, because with such cathodes there has to be a direct connection between the source of heating power and the rest of the circuit. The hum can be minimized by either of the connections shown in Fig. 3-19. In both cases the grid- and plate-return circuits are connected to the electrical midpoint **(center-tap)** of the filament supply. Thus, so far as the grid and plate are concerned, the voltage and current on one side of the filament are balanced by an equal and opposite voltage and current on the other side. This balances out the hum. The balance is never quite perfect, however, so filament-type tubes are never completely hum-free. For this reason directly-heated filaments are used chiefly in transmitting power tubes, where the amount of hum introduced is quite small in comparison to the power-output level.

With indirectly-heated cathodes the source of heating power does not introduce hum by a direct connection. The chief problem with such tubes is the magnetic field set up by the heater. Occasionally, also, there is leakage between the heater and cathode; leakage that allows a small a.c. voltage to get to the grid. Both these things are principally a matter of tube design. However, it is found in practice that, if hum appears, grounding one side of the heater supply will help to reduce it. Sometimes better results are obtained if the heater supply is center-tapped and the center-tap grounded, as in Fig. 3-19.

### *Cathode Bias*

In the simplified amplifier circuits discussed in this chapter, grid bias has been supplied by a battery. However, it is seldom obtained that way in an actual piece of equipment that operates from the power line. **Cathode bias** is the type commonly used.

The cathode-bias method uses a resistor connected in series with the cathode, as shown at $R$ in Fig. 3-20. The direction of plate-current flow is such that the end of the resistor nearest the cathode is positive. The voltage drop across $R$ therefore places a *negative* voltage on the grid. This negative bias is obtained from the steady d.c. plate current.

If the alternating component of plate current flows through $R$ when the tube is amplifying, the voltage drop caused by the a.c. will be degenerative (note the similarity between this circuit and that of Fig. 3-13A). To prevent this the resistor is by-passed by a condenser, $C$, that has very low reactance compared to the resistance of $R$. The capacitance required at $C$ depends upon the value of $R$ and the frequency being amplified. Depending on the type of tube and the particular kind of operation, $R$ may be between about 250 and 3000 ohms. For good by-passing at the low audio frequencies, $C$ should be 10 to 50 micro-

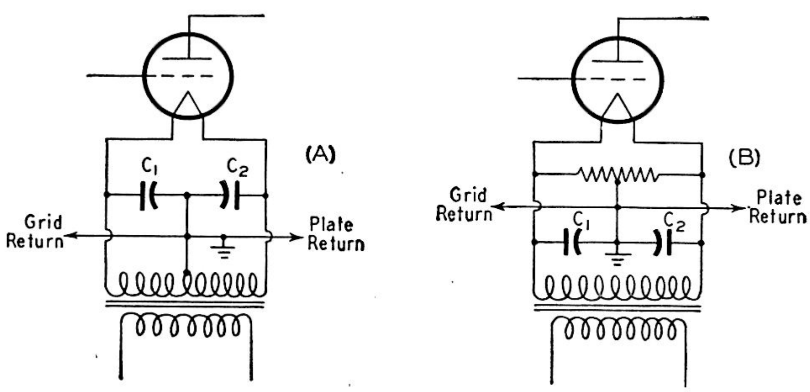



*Fig. 3-19* — **Filament center-tapping methods for use with directly-heated tubes.**

farads (electrolytic condensers are used for this purpose). At radio frequencies, capacitances of about 100 $\mu\mu$fd. to 0.1 $\mu$fd. are used; the small values are sufficient at very high frequencies and the largest at low and medium frequencies. In the range 3 to 30 megacycles a capacitance of 0.01 $\mu$fd. is satisfactory.

The value of cathode resistor can easily be calculated from the known operating conditions of the tube. The proper grid bias and plate current always are specified by the manufacturer. Knowing these, the required resistance can be found by applying Ohm's Law.

Example: It is found from tube tables that the tube to be used should have a negative grid bias of 8 volts and that at this bias the plate current will be 12 milliamperes (0.012 amp.). The required cathode resistance is then

$$R = \frac{E}{I} = \frac{8}{0.012} = 667 \text{ ohms.}$$

The nearest standard value, 680 ohms, would be close enough. The power used in the resistor is

$$P = EI = 8 \times 0.012 = 0.096 \text{ watt.}$$

A ¼-watt or ½-watt resistor would have ample rating.

The current that flows through $R$ is the *total* cathode current. In an ordinary triode amplifier this is the same as the plate current, but in a screen-grid tube the cathode current is the sum of the plate and screen currents.

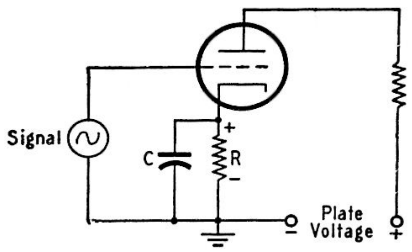



*Fig. 3-20* — Cathode biasing. *R* is the cathode resistor and *C* is the cathode by-pass condenser.

Hence these two currents must be added when calculating the value of cathode resistor required for a screen-grid tube.

Example: A receiving pentode requires 3 volts negative bias. At this bias and the recommended plate and screen voltages, its plate current is 9 ma. and its screen current is 2 ma. The cathode current is therefore 11 ma. (0.011 amp.). The required resistance is

$$R = \frac{E}{I} = \frac{3}{0.011} = 272 \text{ ohms.}$$

A 270-ohm resistor would be satisfactory. The power in the resistor is

$$P = EI = 3 \times 0.011 = 0.033 \text{ watt.}$$

The cathode resistor method of biasing is convenient because it avoids the use of batteries or other source of fixed voltage. However, that is not its only advantage: it is also self-regulating, because if the tube characteristics vary slightly from the published values (as they do in practice) the bias will increase if the plate current is slightly high, or decrease if it is slightly low. This tends to hold the plate current at the proper value. For the same reason, the value of the cathode resistance is not highly critical. Cathode bias also avoids any tendency toward unwanted feed-back that might occur when a single fixed-bias source is used to furnish bias for several amplifiers. Even a very small a.c. voltage drop in the impedance of a bias source can cause oscillation (if the feed-back is positive) or loss of gain (if the feed-back is negative) when the voltage is applied to the first stage of amplification in an amplifier having several stages, simply because the gain in a multistage amplifier is likely to be very large.

The calculation of the bias resistor in a resistance-coupled amplifier is not as easy as the examples above. This is because the actual voltages that should be used on the plate and grid are not ordinarily known. The difficulty is that the voltage drop in the plate resistor causes the actual voltage at the plate of the tube to be considerably less than the plate-supply voltage, and the lower plate voltage requires a different value of bias than that given in the published operating conditions for the tube. The proper voltages can be found by a cut-and-try process from the tube characteristic curves. However, representative data for the tubes commonly used as resistance-coupled amplifiers are given in Chapter Nine, including cathode-resistor values.

### *Screen Supply*

In practical circuits using tetrodes and pentodes the voltage for the screen frequently is taken from the plate supply through a resistor. A typical circuit for an r.f. amplifier is shown in Fig. 3-21. Resistor *R* is the **screen dropping resistor,** and *C* is the **screen by-pass condenser.** In flowing through *R*, the screen current causes a voltage drop in *R* that reduces the plate-supply voltage to the proper value for the screen. When the plate-supply voltage and the screen current are known, the value of *R* can be calculated from Ohm's Law.

Example: An r.f. receiving pentode has a rated screen current of 2 milliamperes (0.002 amp.) at normal operating conditions. The rated screen voltage is 100 volts, and the plate supply gives 250 volts. To put 100 volts on the screen, the drop across *R* must be equal to the difference between the plate-supply voltage and the screen voltage; that is, 250 — 100 = 150 volts. Then

$$R = \frac{E}{I} = \frac{150}{0.002} = 75{,}000 \text{ ohms.}$$

The power to be dissipated in the resistor is

$$P = EI = 150 \times 0.002 = 0.3 \text{ watt.}$$

A ½- or 1-watt resistor would be satisfactory.

The reactance of the screen by-pass condenser, *C*, should be low compared with the screen-to-cathode impedance. For radio-frequency applications a capacitance of 0.01 μfd. is amply large.

In some circuits the screen voltage is obtained from a voltage divider connected across the plate supply. The design of voltage dividers is discussed in Chapter Seven.

## SPECIAL TUBE TYPES

### *Beam Tubes*

"Beam tetrodes" are tetrode tubes constructed in such a way that the power sensitivity is very high. Beam tubes are useful as both radio-frequency and audio-frequency power amplifiers, and are available in output ratings from a few watts up to several hundred watts. The grids in a beam tube are so constructed and aligned as to form the electrons traveling to the plate into concentrated beams. This makes it possible to draw large plate currents at relatively low plate voltages, and also

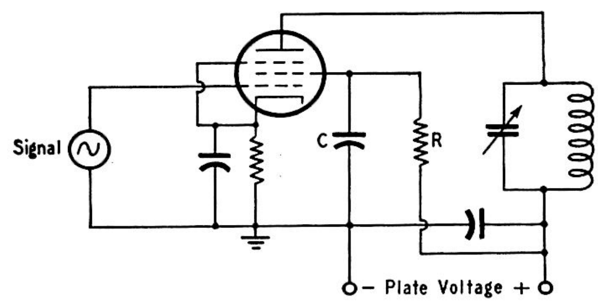



*Fig. 3-21* — Screen-voltage supply for a pentode tube through a dropping resistor, *R*. The screen by-pass condenser, *C*, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

reduces the number of electrons that are captured by the screen. Additional design features overcome the effects of secondary emission, so that a suppressor grid is not needed.

Beam tubes are used with the same circuits, and operated in the same way, as conventional tetrodes and pentodes.

#### *Multipurpose Tubes*

A number of "combination" tubes are available to perform more than one function, particularly in receiver circuits. For the most part these are simply multiunit tubes made up of individual tube-element structures, combined in a single bulb for compactness and economy.

Among the simplest multipurpose types are full-wave rectifiers, combining two diodes in one envelope, and twin triodes, consisting of two triodes in one bulb. More complex types include duplex-diode triodes (two diodes and a triode in one structure), duplex-diode pentodes, converters and mixers (for superheterodyne receivers), combination power tubes and rectifiers, and so on. In many cases the nature of the tubes so combined can be identified by the name given the composite structure.

#### *Mercury-Vapor Rectifiers*

For a given value of plate current, the power lost in a diode rectifier will be reduced if it is possible to decrease the voltage drop from plate to cathode. A small amount of mercury in the tube will vaporize when the cathode is heated and, further, will ionize when plate voltage is applied. The positive ions neutralize the space charge and reduce the plate-cathode voltage drop to a practically constant value of about 15 volts, regardless of the value of plate current.

Since this voltage drop is smaller than can be attained with purely thermionic conduction, there is less power loss in a mercury-vapor rectifier than in a vacuum rectifier. Also, the voltage drop in the tube is constant despite variations in load current. Mercury-vapor tubes are widely used in rectifiers built to deliver large power outputs.

#### *Grid-Control Rectifiers*

If a grid is inserted in a mercury-vapor rectifier it is found that, with sufficient negative grid bias, it is possible to prevent plate current from flowing. However, this is true *only if the bias is present before plate voltage is applied.* If the bias is lowered to the point where plate current can flow, the mercury vapor will ionize and the grid will lose control of plate current, because the space charge disappears when ionization occurs. The grid can assume control again only after the plate voltage is reduced below the ionizing voltage.

The same phenomenon also occurs in triodes filled with other gases that ionize at low pressure. Grid-control rectifiers or **thyratrons** find considerable application in "electronic switching."

## Oscillators

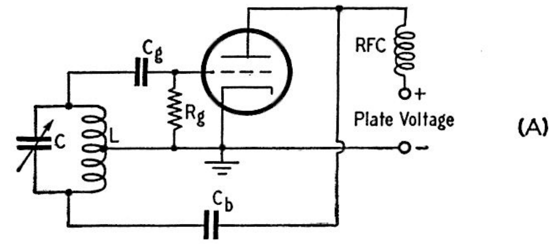
It was mentioned earlier in this chapter that if there is enough positive feed-back in an amplifier circuit, self-sustaining oscillations will be set up. When an amplifier is arranged so that this condition exists it is called an **oscillator.**

Oscillations normally take place at only one frequency, and a desired frequency of oscillation can be obtained by using a resonant circuit tuned to that frequency. The proper phase for positive feed-back can be obtained quite easily from a single tuned circuit. For example, in Fig. 3-22A the circuit $LC$ is tuned to the desired frequency of oscillation. The coil $L$ is tapped and the cathode of the tube is connected to the tap. The grid and plate are connected to opposite ends of the tuned circuit. There will be a voltage drop across the tuned circuit, a voltage drop that increases progressively along the turns of the coil when viewed from one end. At an instant when the upper end of $L$ is positive, for instance, the lower end is negative. However, the tap on the coil is at an intermediate voltage and so is negative with respect to the upper end of $L$, and positive with respect to the lower end. Or, viewed from the tap, the upper end of $L$ is positive and the lower end is negative. Therefore the grid and plate ends of the coil are opposite in polarity, or opposite in phase. This is the right phase relationship for positive feed-back.

The *amount* of feed-back depends on the *position* of the tap. If the tap is too close to either end of the coil the circuit will not oscillate. If the tap is too near the grid end the voltage drop is too small to give enough feed-back, and if it is too near the plate end the impedance between the cathode and plate is too small to permit good amplification. Maximum feed-back usually is obtained when the tap is somewhere near the center of the coil.

It will be observed that the circuit of Fig. 3-22A is parallel fed, $C_b$ being the blocking condenser. The value of $C_b$ is not critical so long as its reactance is low at the operating frequency.

Condenser $C_g$ is the **grid condenser.** It and $R_g$ (the **grid leak**) are used for the purpose of obtaining grid bias for the tube. In this (and practically all) oscillator circuits the tube generates its own bias. When the grid end of the tuned circuit is positive with respect to the cathode, the grid attracts electrons from the cathode. These electrons cannot flow through $L$ back to the cathode because $C_g$ "blocks"



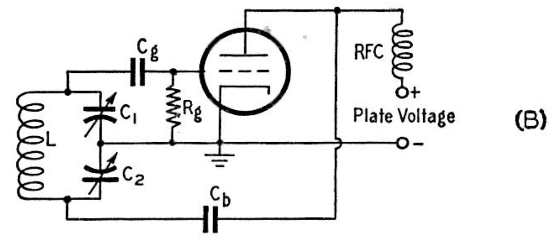




*Fig. 3-22* — **Basic oscillator circuits. Feed-back voltage is obtained by tapping the grid and cathode across a portion of the tuned circuit. In the Hartley circuit the tap is on the coil, but in the Colpitts circuit the voltage is obtained from the drop across a condenser.**

direct current. They therefore have to flow or "leak" through $R_g$ to cathode, and in doing so cause a voltage drop in $R_g$ that places a negative bias on the grid. The amount of bias so developed is equal to the grid current multiplied by the resistance of $R_g$ (Ohm's Law). The value of grid-leak resistance required depends upon the kind of tube used and the purpose for which the oscillator is intended. Values range all the way from a few thousand to several hundred thousand ohms. The capacitance of $C_g$ should be large enough to have low reactance at the operating frequency.

The circuit shown at B in Fig. 3-22 uses the voltage drops across two condensers in series in the tuned circuit to supply the feed-back. Other than this, the operation is the same as just described. The feed-back can be varied by varying the ratio of the reactances of $C_1$ and $C_2$ (that is, by varying the ratio of their capacitances). To maintain the same oscillation frequency the total *capacitance* across $L$ must be constant; this means that every time $C_1$, for example, is adjusted to change the feed-back, $C_2$ must be adjusted in the *opposite* sense to return the total capacitance and thereby the frequency to the original value.

Another type of oscillator, called the **tuned-plate tuned-grid** circuit, is shown in Fig. 3-23. Resonant circuits tuned approximately to the same frequency are connected between grid and cathode and between plate and cathode. The two coils, $L_1$ and $L_2$, are not magnetically-coupled. The feed-back is through the grid-plate capacitance of the tube, and will be in the right phase to be positive when the plate circuit, $C_2L_2$, is tuned to a slightly higher frequency than the grid circuit, $L_1C_1$. The amount of feed-back can be adjusted by varying the tuning of either circuit. The frequency of oscillation is determined by the tuned circuit that has the higher $Q$. The grid leak and grid condenser have the same functions as in the other circuits. In this case it is convenient to use series feed for the plate circuit, so $C_b$ is a by-pass condenser to guide the r.f. current around the plate supply.

Practically all feed-back oscillator circuits (and there is an endless variety of them) are variations of these general types. They differ in details and appearance, and some use two or more tubes to accomplish the purpose. However, the basic feature of all of them is that there is positive feed-back in the proper amplitude to sustain oscillation.

### *Oscillator Operating Characteristics*

As a general rule, oscillators are *power*-generating devices. There are exceptions: in some cases the oscillator is used primarily to generate a *voltage* that is then applied to an amplifier that does not require power in its grid circuit. This type of oscillator is used principally in certain types of measuring equipment; the oscillators used in transmitters and receivers usually are called upon to deliver some power.

When an oscillator is delivering power to a load, the adjustment for proper feed-back will depend on how heavily the oscillator is loaded. If the feed-back is not large enough — that is, if the **grid excitation** is too small — a slight change in load may tend to throw the circuit into and out of oscillation. On the other hand, too much feed-back will make the grid current excessively high, with the result that the power loss in the grid circuit is larger than necessary. The oscillator itself supplies this grid power, so excessive feed-back lowers the over-all efficiency because whatever power is used in the grid circuit is not available as useful output.

One of the most important considerations in oscillator design is frequency stability. Almost invariably we want the generated frequency to be as constant as possible. The principal factors that cause a change in frequency are (1) temperature, (2) plate voltage, (3) loading, (4) mechanical variations of circuit elements. Temperature changes will cause vacuum-tube elements to expand or contract slightly, thus causing variations in the interelectrode capacitances. Since these are unavoidably part of the tuned circuit, the frequency will change correspondingly. Temperature changes in the

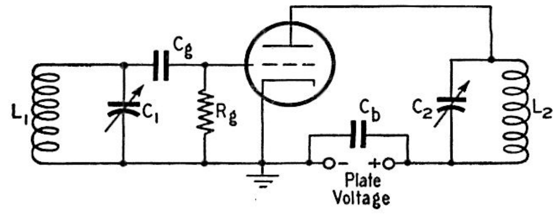



*Fig. 3-23* — **The tuned-plate tuned-grid oscillator.**

coil or condenser will alter their inductance or capacitance slightly, again causing a shift in the resonant frequency. These effects are relatively slow in operation, and the frequency change caused by them is called **drift.**

Load variations act in much the same way as plate-voltage variations. A temperature change in the load may also result in drift.

Plate-voltage variations will cause a corresponding shift in frequency; this type of frequency shift is called **dynamic** instability. Dynamic instability can be reduced by using a tuned circuit of high effective $Q$. Since the tube and load represent a relatively low resistance in parallel with the circuit, this means that a low $L/C$ ratio ("high-$C$") must be used and that the circuit should be lightly loaded. Dynamic stability also can be improved by using a high value of grid leak; this increases the grid bias and raises the effective resistance of the tube as seen by the tank circuit. Using relatively high plate voltage and low plate current also helps.

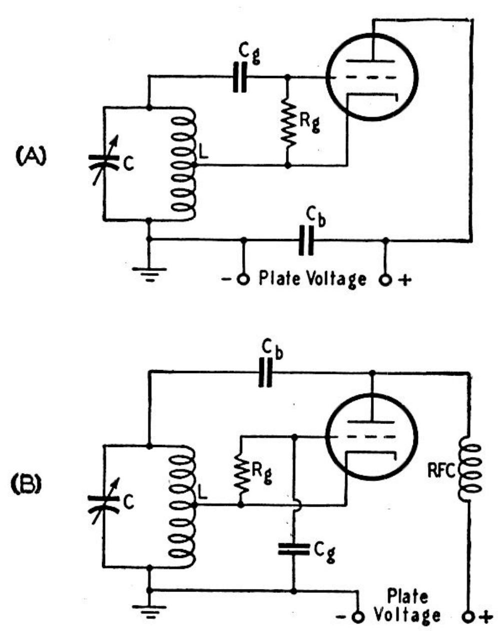



*Fig. 3-24* — **Showing how the r.f. ground on a typical oscillator circuit (Hartley) may be placed on either the plate (A) or grid (B) instead of the more conventional method of grounding the cathode. Provided the proper provisions are made for supplying cathode and plate voltages, the circuit operation is unchanged by shifting the r.f. ground to any desired point.**

Mechanical variations, usually caused by vibration, cause changes in inductance and/or capacitance that in turn cause the frequency to "wobble" in step with the vibration.

Methods of minimizing frequency variations in oscillators are taken up in detail in later chapters.

### *Ground Point*

In the oscillator circuits shown in Figs. 3-22 and 3-23 the cathode is connected to ground. It is not actually essential that the radio-frequency circuit should be grounded at the cathode; in fact, there are many times when an *r.f.* ground on some other point in the circuit is desirable. The r.f. ground can be placed at any point so long as proper provisions are made for feeding the supply voltages to the tube elements.

Fig. 3-24 shows the Hartley circuit with (A) the plate end of the circuit grounded, and (B) the grid end. In A, no r.f. choke is needed in the plate circuit because the plate already is at ground potential and there is no r.f. to choke off. All that is necessary is a by-pass condenser, $C_b$, across the plate supply. Direct current flows to the cathode through the lower part of the tuned-circuit coil, $L$.

The grounded-grid circuit at B is essentially the same as the circuit in Fig. 3-22A except that the ground point and negative plate-voltage connection have been placed at the grid end of the tuned circuit.

One advantage of either type of circuit (the one in Fig. 3-24A is widely used) is that the frame of the tuning condenser can be grounded. With a grounded-cathode oscillator, both ends of the tuned circuit are "hot"; that is, there is an r.f. voltage to ground from both ends of the circuit. When the ordinary type of tuning condenser is used in such a circuit there is a slight change in capacitance when the hand is brought near the tuning shaft for adjustment of capacitance. This "hand capacitance" or "body capacitance" is annoying because the oscillator frequency changes when the hand is brought near the tuning control. It is overcome by grounding (for r.f.) the condenser shaft and by using a condenser that has a frame with metal end plates.

Tubes having indirectly-heated cathodes are more easily adaptable to circuits grounded at other points than the cathode than are tubes having directly-heated filaments. With the latter tubes special precautions have to be taken to prevent the filament from being by-passed to ground by the capacitance of the filament-heating transformer.

## NEGATIVE-RESISTANCE OSCILLATORS

If a tuned circuit could be built without resistance, a small amount of energy introduced into the circuit would start an oscillation that would continue indefinitely. It would do so because, in a circuit having no power losses, the power never diminishes and therefore is always available to keep the oscillation going. Of course, such a circuit cannot be built.

However, it was explained in Chapter Two that a resonant circuit has a definite value of parallel impedance at resonance, and that that impedance is a pure resistance. If we could connect across the circuit a value of "negative" resistance equal to the parallel resistance of the circuit, the negative resistance would cancel the "positive" (real) resistance of the circuit and we would have a circuit that is, in effect, without resistance.

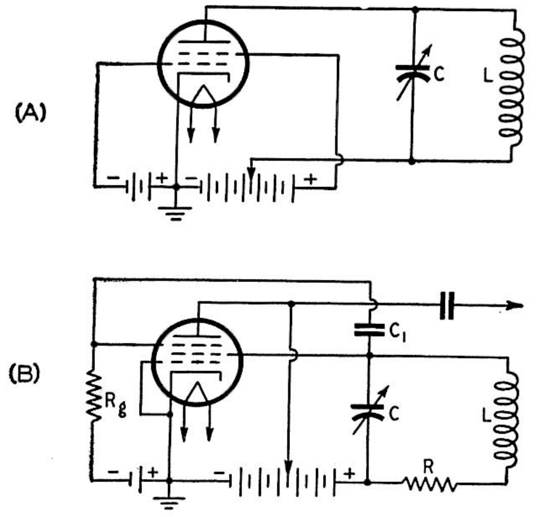



*Fig. 3-25* — **Negative-resistance oscillator circuits. A, dynatron; B, transitron.**

A negative resistance is one having the opposite characteristics to real or positive resistance. In a negative resistance the current *increases* when the voltage is decreased, and vice versa. Also, a negative resistance does not consume power; it generates it. Under certain conditions a vacuum tube can be made to operate like a negative resistance, and thus can be connected to a tuned circuit to set up oscillations. Two circuits for doing this are shown in Fig. 3-25.

The circuit at A is called the **dynatron** oscillator. It functions because of the secondary emission from the plate that occurs in certain types of screen-grid tetrodes. It makes use of the fact that, at certain values of screen voltage, the plate current of a screen-grid tetrode decreases when the plate voltage is increased. This gives a negative plate-resistance characteristic.

In Fig. 3-25B, negative resistance is produced by virtue of the fact that, if the suppressor grid of a pentode is given negative bias, electrons that normally would pass through the suppressor to the plate are turned back to the screen, thus increasing the screen current and reversing normal tube action. The negative resistance produced between the screen and suppressor grids is sufficiently low so that ordinary tuned circuits will oscillate readily up to 15 Mc. or so. This circuit is known as the **transitron.**

For most amateur applications, negative-resistance oscillators do not have enough advantages to bring them into wide use. Feedback oscillators are generally more adaptable to wide frequency ranges, can generate more power, and are more readily adjusted to meet varying conditions. The transitron oscillator is used occasionally in measuring equipment.

# High-Frequency Communication

Much of the appeal of amateur communication on the high frequencies lies in the fact that the results are not always predictable. Transmission conditions on the same frequency vary with the year and even with the time of day. Although these variations usually follow certain established cycles, many peculiar effects can be observed from time to time. Every radio amateur should have some understanding of the known facts about radio-wave propagation so that he will stand some chance of interpreting the unusual conditions when they occur. The observant amateur is in an excellent position to make worth-while contributions to the science, provided he has sufficient background to understand his results. The serious amateur can, for example, coöperate with the National Bureau of Standards in its NBS-ARRL 28-Mc. Observing Project, about which more is said at the end of this chapter. Or he may develop a new theory of propagation for the very-high frequencies or the microwave region, as did the late Ross Hull. By making extensive observations of 56-Mc. conditions over a long-distance path and correlating the results with various weather conditions, Mr. Hull was able to establish the now-accepted theory of "tropospheric bending."

## What To Expect on the Various Amateur Bands

The 3.5-Mc., or "80-meter," band is a more useful band during the night than during the daylight hours. In the daytime, one can seldom hear signals from a distance of greater than 100 miles or so, but during the darkness hours distances up to several thousand miles are not unusual, and transoceanic contacts are regularly made during the winter months. During the summer, the static level is high in some parts of the world, and the sharp crashes of static often make reception difficult. The 3.5-Mc. band supports the majority of the traffic nets throughout the country, and it is also a great gathering place for "rag-chewers." Low power and simple antennas can be used here with good results.

The 7-Mc., or "40-meter," band has many of the same characteristics as 3.5, except that the distances that can be covered during the day and night hours are increased. During daylight, distances up to a thousand miles can be covered under good conditions, and during the dawn and dusk periods in winter it is possible to work stations as far as the other side of the world, the signals following the darkness path. The winter months are somewhat better than the summer ones. Rag-chewing, traffic handling and DX (working foreign countries) are popular activities on the band, in the order named. Here again antennas are not too important, although results will be improved in proportion to the effectiveness of the antenna system. In general, summer static is much less of a problem than on 80 meters, although it can be serious in the semitropical zones.

The 14-Mc., or "20-meter," band is probably the best one for long-distance work. During portions of the sunspot cycle it is open to some part of the world during practically all of the 24 hours, while at other times it is generally useful only during daylight hours and the dawn and dusk periods. DX activity is paramount, with rag-chewing next. Being less consistent, day by day, traffic handling is not too general, although many long-distance schedules are kept on the band. Effective antennas are more necessary than on the lower frequencies, but many amateurs enjoy excellent results with simple antennas and low power. Automobile ignition and other types of man-made interference begin to be a problem on this band.

The 28-Mc. band is generally considered to be a DX band during the daylight hours and a local rag-chewer's band during the hours of darkness. However, during parts of the sunspot cycle, the band is "open" into the late evening hours for DX communication. The band is even less consistent than 14 Mc., but this very fact is what makes it so fascinating for its many followers. It is not unusual for a foreign station to appear suddenly with a loud signal when only U.S. stations, or none at all, are being heard. High-performance antennas are almost a necessity for best results, but its small dimensions make the rotary beam a popular choice for the band. These antennas can be turned to direct the radiation in the desired direction, and they are used to provide useful gain on reception as well. A good antenna is far more important on this band than high power.

## Characteristics of Radio Waves

Radio waves differ from other forms of electromagnetic radiation (such as light and heat) in the manner in which they are generated and detected and in their wavelength. The wavelength spectrum of radio waves is greater than either heat or light, and ranges from approximately 30,000 meters to a small fraction of a centimeter. This corresponds to a frequency range of about 10 kc. to 1,000,000 Mc. They travel at the same velocity as light waves (about 186,000 miles per second in free space) and can be reflected, refracted and diffracted the way light and heat waves can.

The passage of radio energy through space is explained by a concept of traveling electrostatic and electromagnetic waves. The energy is evenly divided between the two types of fields, and the lines of force of these fields are at right angles to each other, in a plane perpendicular to the direction of travel. A simple representation of this is shown in Fig. 4-1.

### *Polarization*

The polarization of a radio wave is taken as the direction of the lines of force in the electrostatic field. If the plane of this field is perpendicular to the earth, the wave is said to be **vertically polarized;** if it is parallel to the earth, the wave is **horizontally polarized.** The longer waves, when traveling along the ground, usually maintain their polarization in the same plane as was generated at the antenna. The polarization of shorter waves may be altered during travel, however, and sometimes will vary quite rapidly.

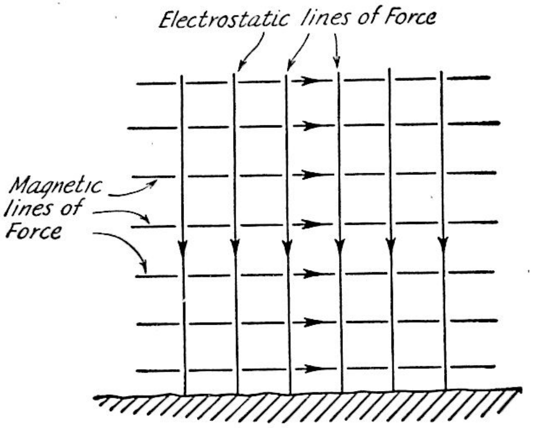



*Fig. 4-1* — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

### *Reflection*

Radio waves may be reflected from any sharply-defined discontinuity of suitable characteristics and dimensions encountered in the medium in which they are traveling. Any conductor (or any insulator having a dielectric constant differing from that of the medium) offers such a discontinuity if its dimensions are at least comparable to the wavelength. The surface of the earth and the boundaries between ionospheric layers are examples of such discontinuities. Objects as small as an airplane, a tree or even a man's body will readily reflect the shorter waves.

### *Refraction*

As in the case of light, a radio wave is bent when it moves obliquely into any medium having a refractive index different from that of the medium it leaves. Since the velocity of propagation differs in the two mediums, that part of the wave front that enters first travels faster if the new medium has a higher velocity of propagation. This tends to swing the wave front around, or "refract" it, in such a manner that the wave is directed in a new direction. If the wave front is one that is traveling obliquely away from the earth, and it encounters a medium with a higher velocity of propagation, the wave will be directed back toward the earth. If the new medium has a lower velocity of propagation, the opposite effect takes place, and the wave is directed away from the earth. Refraction may take place either in the ionosphere (ionized upper atmosphere) or the troposphere (lower atmosphere), or both.

### *Diffraction*

When a wave grazes the edge of an object in passing, it tends to be bent around that edge. This effect, called **diffraction,** results in a diversion of part of the energy of those waves which normally follow a straight or line-of-sight path, so that they may be received at some distance below the summit of an obstruction, or around its edges.

### *Types of Waves*

According to the altitude of the paths along which they are propagated, radio waves may be classified as **ionospheric waves, tropospheric waves** or **ground waves.**

The ionospheric wave (sometimes called the **sky wave**), is that part of the total radiation that is directed toward the ionosphere. Depending upon variable conditions in that region, as well as upon transmitting wavelength, the ionospheric wave may or may not be returned to earth by the effects of refraction and reflection.

The tropospheric wave is that part of the total radiation that undergoes refraction and reflection in regions of abrupt change of dielectric constant in the troposphere, such as the boundaries between air masses of differing temperature and moisture content.

The ground wave is that part of the total radiation that is directly affected by the presence of the earth and its surface features. The ground wave has two components. One is the **surface wave,** which is an earth-guided wave, and the other is the **space wave** (not to be confused with the ionospheric or sky wave). The space wave is itself the resultant of two components — the direct wave and the ground-reflected wave, as shown in Fig. 4-2.

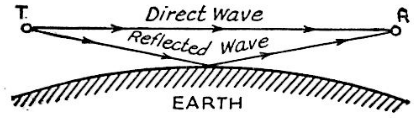



*Fig. 4-2* — Showing how both direct and reflected waves may be received simultaneously in v.h.f. transmission.

# Ionospheric Propagation

Communication between distant points by means of radio waves of frequencies ranging between 3 and 30 Mc. depends principally upon the ionospheric wave. Upon leaving the transmitting antenna, this wave travels upward from the earth's surface at such an angle that it would continue out into space were its path not bent sufficiently to bring it back to earth. The medium that causes such bending is the **ionosphere,** a region in the upper atmosphere, above a height of about 60 miles, where free ions and electrons exist in sufficient quantity to cause a change in the refractive index. This condition is believed to be the effect of ultraviolet radiation from the sun. The ionosphere is not a single region but is composed of a series of layers of varying densities of ionization occurring at different heights. Each layer consists of a central region of relatively dense ionization that tapers off in intensity both above and below.

### *Refraction, Absorption and Reflection*

For a given density of ionization, the degree of refraction becomes less as the wavelength becomes shorter (as the frequency increases). The bending, therefore, is less at high than at low frequencies, and if the frequency is raised to a sufficiently high value, a point is finally reached where the refractive bending becomes too slight to bring the wave back to earth, even though it may enter the ionized layer along a path that makes a very small angle with the boundary of the ionosphere.

The greater the density of ionization, the greater the bending at any given frequency. Thus, with an increase in ionization, the minimum wavelength that can be bent sufficiently for long-distance communication is lessened and the **maximum usable frequency** is increased.

The wave necessarily loses some of its energy in traveling through the ionosphere, this absorption loss increasing with wavelength and also with ionization density. Unusually high ionization, especially in the lower strata of the ionosphere, may cause complete absorption of the wave energy.

In addition to refraction, reflection may take place at the lower boundary of an ionized layer if it is sharply defined; i.e., if there is an appreciable change in ionization within a relatively short interval of travel. For waves approaching the layer at or near the perpendicular, the change in ionization must take place within a difference in height comparable to a wavelength; hence, ionospheric reflection is more apt to occur at longer wavelengths (lower frequencies).

### *Critical Frequency*

When the frequency is sufficiently low, a wave sent vertically upward to the ionosphere will be bent sharply enough to cause it to return to the transmitting point. The highest frequency at which such reflection can occur, for a given state of the ionosphere, is called the **critical frequency.** Although the critical frequency may serve as an index of transmission conditions, it is not the highest useful frequency, since other waves of a higher frequency that enter the ionosphere at angles smaller than 90 degrees (less than vertical) will be bent sufficiently to return to earth. The maximum usable frequency, for waves leaving the earth at very small angles to the horizontal, is in the vicinity of three times the critical frequency.

Besides being directly observable by special equipment, the critical frequency is of more practical interest than the ionization density because it includes the effects of absorption as well as refraction.

### *Virtual Height*

Although an ionospheric layer is a region of considerable depth it is convenient to assign to it a definite height, called the **virtual height.** This is the height from which a simple reflection would give the same effect as the gradual refraction that actually takes place, as illustrated in Fig. 4-3. The wave traveling upward is bent back over a path having an appreciable radius of turning, and a measurable interval of time is consumed in the turning process. The virtual height is the height of a triangle formed as shown, having equal sides of a total length proportional to the time taken for the wave to travel from $T$ to $R$.

### *Normal Structure of the Ionosphere*

The lowest normally useful layer is called the $E$ layer. The average height of the region of maximum ionization is about 70 miles. The ionization density is greatest around local noon; the layer is only weakly ionized at night, when it is not exposed to the sun's radiation. The air at this height is sufficiently dense so that free ions and electrons very quickly meet and recombine.

In the daytime there is a still lower ionized region, the $D$ layer. The $D$-layer intensity is proportional to the height of the sun and is greatest at noon. Low-frequency waves (80 meters) are almost completely absorbed by this layer while it exists, and only the high-angle radiation is reflected by the $E$ layer. (Lower-angle radiation travels farther through the $D$ layer and is absorbed.)

The second principal layer is the $F$ layer, which has a height of about 175 miles at night. At this altitude the air is so thin that recombination of ions and electrons takes place very slowly, inasmuch as particles can travel relatively great distances before meeting. The ionization decreases after sundown, reaching a minimum just before sunrise. In the daytime the $F$ layer splits into two parts, the $F_1$ and $F_2$ layers, with average virtual heights of, respectively, 140 miles and 200 miles. These layers are most highly ionized at about local noon, and merge again at sunset into the $F$ layer.

### *Cyclic Variations in the Ionosphere*

Since ionization depends upon ultraviolet radiation, conditions in the ionosphere vary with changes in the sun's radiation. In addition to the daily variation, seasonal changes result in higher critical frequencies in the $E$ layer in summer, averaging about 4 Mc. as against a winter average of 3 Mc. The $F$ layer shows little variation, the critical frequency being of the order of 4 to 5 Mc. in the evening. The $F_1$ layer, which has a critical frequency near 5 Mc. in summer, usually disappears entirely in winter. The critical frequencies for the $F_2$ are highest in winter (11 to 12 Mc.) and lowest in summer (around 7 Mc.). The virtual height of the $F_2$ layer, which is about 185 miles in winter, averages 250 miles in summer.

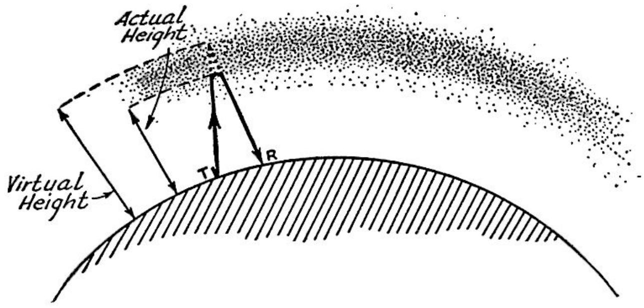



*Fig. 4-3* — Bending in the ionosphere, and the echo or reflection method of determining virtual height.

Seasonal transition periods occur in spring and fall, when ionospheric conditions are found highly variable.

There are at least two other regular cycles in ionization. One such cyclic period covers 28 days, which corresponds with the period of the sun's rotation. For a short time in each 28-day cycle, transmission conditions reach a peak. Usually this peak is followed by a fairly rapid drop to a lower level, and then a slow building-up to the next peak. The 28-day cycle is particularly evident in the 14- and 28-Mc. amateur bands.

The longest cycle yet observed covers about 11 years, corresponding to a similar cycle of sunspot activity. The effect of this cycle is to shift upward or downward the values of the critical frequencies for $F$- and $F_2$-layer transmission. The critical frequencies are highest during sunspot maxima and lowest during sunspot minima. It is during the period of minimum sunspot activity when long-distance transmissions occur on the lower frequencies. At such times the 28-Mc. band is seldom useful for long-distance work, while the 14-Mc. band performs well in the daytime but is not ordinarily useful at night. The most recent sunspot maximum is considered to have occurred in 1938.

### *Magnetic Storms and Other Disturbances*

Unusual disturbances in the earth's magnetic field (magnetic storms) usually are accompanied by disturbances in the ionosphere, when the layers apparently break up and expand. There is usually also an increase in absorption during such a period. Radio transmission is poor and there is a drop in critical frequencies so that lower frequencies must be used for communication. A magnetic storm may last for several days.

Unusually high ionization in the region of the atmosphere below the normal ionosphere may increase absorption to such an extent that sky-wave transmission becomes impossible on high frequencies. The length of such a disturbance may be several hours, with a gradual falling off of transmission conditions at the beginning and an equally gradual building up at the end of the period. **Fade-outs**, similar to the above in effect, are caused by sudden disturbances on the sun. They are characterized by very rapid ionization, with sky-wave transmission disappearing almost instantly, occur only in daylight, and do not last as long as the first type of absorption.

Magnetic storms frequently are accompanied by unusual auroral displays, creating an ionized "curtain" in the polar regions which can act as a reflector of radio waves. **Auroral reflection** is occasionally observed at frequencies as high as 54 Mc. It is characterized on 28 Mc. by a flutter on all signals which makes voice work difficult but not impossible. Directive antennas must be pointed toward the north and not in the direction of the station being worked.

### *Sporadic-E Layer Ionization*

Occasionally scattered patches of clouds of relatively dense ionization appear at heights approximately the same as that of the $E$ layer. The effect is to raise the critical frequency to a value perhaps twice that which is returned from any of the regular layers by normal refraction. Distances of about 500 to 1250 miles may be covered at 50 Mc. if the ionized cloud is situated midway between transmitter and receiver, or is of any very considerable extent. This effect, while infrequently observed in winter, is prevalent during the late spring and early summer, with no apparent correlation of the condition with the time of day.

The presence of sporadic-$E$ refraction on the 14- and 28-Mc. bands is indicated by an abnormally short distance between the transmitter and the point where the wave first is returned to earth as when, for example, 14-Mc. signals from a transmitter only 100 miles distant may arrive with an intensity usually associated with distances of this order on 7 and 3.5 Mc.

### *Scatter*

Scatter signals are heard on any band, but are more easily recognizable on the higher frequencies because of the extended skip zone. They are signals reflected from large discontinuities at a distance, such as sharp concentrations of ionization in any of the normal layers, sporadic-$E$ clouds or (rarely) large land objects. They result in one's hearing signals within the normal skip zone. Scatter signals are never very loud, and have a slight flutter characteristic. A further indication of scatter reflection is that, when beam antennas are used to indicate the direction of arrival of the wave, the ray path is not necessarily the direct route but can even be at right angles or in the opposite direction.

### *Meteor Trails*

Another phenomenon generally encountered in the 28-Mc. band, but also observed in the 14- and 50-Mc. bands, is one characterized by sudden bursts of intensity of a signal. These bursts last less than a second, generally, and are caused by reinforced reflection from the ionized trail of a meteor. The meteor, entering the earth's atmosphere at high velocity, heats by friction against the atmosphere and leaves a trail of ionized atmosphere. It takes a finite time for the ionized molecules to recombine, and during this time a small ionized cloud exists. If it is in the ray path of a signal, it may serve to reinforce the signal and cause the burst in intensity. When the meteor is moving in a direction somewhat parallel to the ray path, it can induce a rising or falling "whistle" on the signal, for a second or so. The effects of bursts and whistles can be observed at any time during the day or night, if there is any marked meteor activity, and during rare "meteor showers" the ionized clouds can serve in almost the same manner that sporadic-$E$ does to make long-distance work possible on 50 Mc.

### *Wave Angle*

The smaller the angle at which a wave leaves the earth, the less will be the bending required in the ionosphere to bring it back and, in general, the greater the distance between the point where it leaves the earth and that at which it returns. This is shown in Fig. 4-4. The vertical angle which the wave makes with a tangent to the earth is called the **wave angle** or **angle of radiation.**

### *Skip Distance*

Since greater bending is required to return the wave to earth when the wave angle is high, at the higher frequencies the refraction frequently is not enough to give the required bending unless the wave angle is smaller than a certain angle called the **critical angle.** This is illustrated in Fig. 4-4, where waves at angles of $A$ or less give useful signals while waves sent at higher angles penetrate the layer and are not returned. The distance between $T$ and $R_1$ is, therefore, the shortest possible distance over which communication by normal ionospheric refraction can be accomplished.

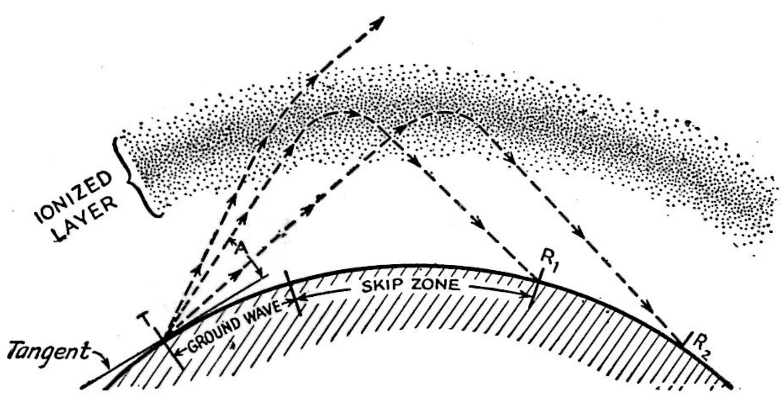



*Fig. 4-4* — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than $A$) are not bent enough to be returned to earth. As the angle is increased, the waves return to earth at increasingly greater distances.

The area between the end of the useful ground wave and the beginning of ionospheric-wave reception is called the **skip zone.** The extent of skip zone depends upon the frequency and the state of the ionosphere, and is greater the higher the transmitting frequency and the lower the critical frequency. Skip distance depends also upon the height of the layer in which the refraction takes place, the higher layers giving longer skip distances for the same wave angle. Wave angles at the transmitting and receiving points are usually, although not always, approximately the same for any given wave path.

It is readily possible for the ionospheric wave to pass through the $E$ layer and be refracted back to earth from the $F$, $F_1$ or $F_2$ layers. This is because the critical frequencies are higher in the latter layers, so that a signal too high in frequency to be returned by the $E$ layer can still come back from one of the others, depending upon the time of day and the existing conditions. Depending upon the wave angle and the frequency, it is sometimes possible to carry on communication via either the $E$ or $F_1$-$F_2$ layers on the same frequency.

### *Multihop Transmission*

On returning to the earth the wave can be reflected upward and travel again to the ionosphere. There it may once more be refracted, and again bent back to earth. This process may be repeated several times. Multihop propagation of this nature is necessary for transmission over great distances because of the limited heights of the layers and the curvature of the earth, since at the lowest useful wave angles (of the order of a few degrees, waves at lower angles generally being absorbed rapidly at high frequencies by being in contact with the earth) the maximum one-hop distance is about 1250 miles for refraction from the $E$ layer and around 2500 miles for the $F_2$ layer. However, ground losses absorb some of the energy from the wave on each reflection (the amount of the loss varying with the type of ground and being least for reflection from sea water). Thus, when the distance permits, it is better to have one hop rather than several, since the multiple reflections introduce losses that are higher than those caused by the ionosphere alone.

### *Fading*

Two or more parts of the wave may follow slightly different paths in traveling to the receiving point, in which case the difference in path lengths will cause a phase difference to exist between the wave components at the receiving antenna. The field strength therefore may have any value between the numerical sum of the components (when they are all in phase) and zero (when there are only two components and they are exactly out of phase). Since the paths change from time to time, this causes a variation in signal strength called **fading.** Fading can also result from the combination of single-hop and multihop waves, or the combination of a ground wave with an ionospheric or tropospheric wave. Such a condition gives rise to an area of severe fading near the limiting distance of the ground wave, better reception being obtained at both shorter and longer distances where one component or the other is considerably stronger. Fading may be rapid or slow, the former type usually resulting from rapidly-changing conditions in the ionosphere, the latter occurring when transmission conditions are relatively stable.

It frequently happens that transmission conditions are different for waves of slightly different frequencies, so that in the case of voice-modulated transmission, involving sidebands differing slightly from the carrier in frequency, the carrier and various sideband components may not be propagated in the same relative amplitudes and phases they had at the transmitter. This effect, known as **selective fading,** causes severe distortion of the signal.

### *Tropospheric Propagation*

Changes in refractive index of air masses in the lower atmosphere often permit work over greater-than-normal distances on 28 Mc. and higher frequencies. The effect can be observed on 28 Mc., but it is generally more marked on 50 and 144 Mc. The subject is treated in detail in Chapter Eleven.

## ● OPTIMUM WAVE ANGLES

One of the requirements in high-frequency radio transmission is to send a wave to the ionosphere in such a way that it will have the best chance of being returned to earth. This is chiefly a matter of the angle at which the wave enters the layer, although in some cases polarization may be of importance. The desirable conditions change considerably with variations in frequency.

The desirable conditions for waves of different frequencies can be summarized as follows, in terms of the various amateur bands:

### *3.5 Mc.*

Waves at all angles of radiation usually will be reflected, so that no energy is lost by high-angle radiation. However, the lower-angle waves will, in general, give the greatest distances. Polarization on this band is not of great importance.

### *7 Mc.*

Under most conditions, angles of radiation up to about 45 degrees will be returned to earth; during the sunspot maximum still higher angles are useful. It is best to concentrate the radiation below 45 degrees. Polarization is not important, except that losses probably will be higher with vertical polarization.

### *14 Mc.*

For long-distance transmission, most of the energy should be concentrated at angles below about 20 degrees. Higher angles are useful for comparatively short distances (300–400 miles), although 30 degrees is about the maximum useful angle. Aside from the probable higher losses with vertical polarization, the polarization may be of any type.

### *28 Mc.*

Angles of 10 degrees or less are most useful. As in the case of 14 Mc., polarization is not important.

## PREDICTION CHARTS

The National Bureau of Standards offers prediction charts three months in advance, for use in predicting and studying long-distance communication on the usable frequencies above 3.5 Mc. By means of these charts, it is possible to predict with considerable accuracy the maximum usable frequency that will hold over any path on the earth during a monthly period. The particular great-circle path is drawn by the operator on a piece of transparent paper, and the prediction charts include maps that make this a simple matter. **Control points** are then marked on this great-circle route at distances of 2000 kilometers from each station. By moving the transparent paper over the prediction chart and observing where the control points fall, the maximum usable frequency can be obtained for any hour of the 24. The charts are based on ionosphere soundings made at a number of stations throughout the world, coupled with considerable statistical data. The charts are conservative enough to enable the amateur to anticipate and plan his best operating times, particularly on the 14- and 28-Mc. bands. Amateurs who work on 50 Mc. and are interested in the occasional $F_2$ "openings" in this band watch the charts with great interest. They can be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., for 15 cents a copy or $1.50 per year on subscription. They are called "CRPL-D Basic Radio Propagation Predictions."

## N.B.S.-A.R.R.L. 28-MC. OBSERVING PROJECT

Any amateur with 28-Mc. receiving and transmitting equipment can apply to the National Bureau of Standards and ask to take part in the 28-Mc. Observing Project. This service provides valuable information for the NBS in their propagation work, and only requires that the amateur make regular monthly reports on a minimum number of schedules or receiving observations. Complete report forms are furnished by the Bureau, together with a simplified 28-Mc. prediction chart that shows what the Bureau has predicted for each particular month. The program is not confined to amateurs in the United States, and stations throughout the world are taking an active part. Address your application to Central Radio Propagation Laboratory, National Bureau of Standards, Washington 25, D. C.

CHAPTER 5

# High-Frequency Receivers

A good receiver in the amateur station makes the difference between mediocre contacts and solid QSOs, and its importance cannot be emphasized too much. In the v.h.f. bands that are not too crowded, **sensitivity** (the ability to bring in weak signals) is the most important factor in a receiver. In the more crowded amateur bands, good sensitivity must be combined with **selectivity** (the ability to distinguish between signals separated by only a small frequency difference) for best results and general ease of reception. Using only a simple receiver, old and experienced operators can copy signals that would be missed entirely by newer amateurs, but their success is because of their experience and not the receiving equipment. On the other hand, a less-experienced operator can use modern techniques to obtain the same degree of success, provided he understands the operation of his more advanced type of receiver and how to get the most out of it.

A number of signals may be picked up by the receiving antenna, and the receiver must be able to separate them and allow the operator to copy the one he wants. This ability is called "selectivity." To receive weak signals, the receiver must furnish enough **amplification** to amplify the minute signal power delivered by the antenna up to a useful amount of power that will operate a loudspeaker or set of headphones. Before the amplified signal can operate the 'speaker or 'phones, however, it must be converted to audio-frequency power by the process of **detection.** The sequence of amplification is not too important — some of the amplification can take place (and usually does) before detection, and some can be used after detection.

There are two major differences between receivers for 'phone reception and for c.w. reception. A 'phone signal has sidebands that make the signal take up about 6 or 8 kc. in the band, and the audio quality of the received signal is impaired if the passband of the receiver is less than this amount. On the other hand, a c.w. signal occupies only a few hundred cycles at the most, and consequently the passband of a c.w. receiver can be small. In either case, if the passband of the receiver is more than is necessary, signals adjacent to the desired one can be heard, and the selectivity of the receiver is said to be poor. The other difference is that the detection process delivers directly the audio frequencies present as modulation on a 'phone signal, but there is no modulation on a c.w. signal and additional technique is required to make the signal audible. It is necessary to introduce a second radio frequency, differing from the signal frequency by a suitable audio frequency, into the detector circuit to produce an audible beat. The frequency difference, and hence the **beat-note,** is generally of the order of 500 to 1000 cycles, since these tones are within the range of optimum response of both the ear and the headset. If the source of the second radio frequency is a separate oscillator, the system is known as **heterodyne** reception; if the detector itself is made to oscillate and produce the second frequency, it is known as an **autodyne** detector. Modern superheterodyne receivers (described later) generally use a separate oscillator to generate the beat-note. Summing up the two differences, 'phone receivers can't use as much selectivity as c.w. receivers, and c.w. receivers require some kind of beating oscillator to give an audible signal. Broadcast receivers can receive only 'phone signals because no beat oscillator is included. On the other hand, communications receivers include beat oscillators and often some means for varying the selectivity.

## Receiver Characteristics

### *Sensitivity*

Confusion exists among some radio men when talking about the "sensitivity" of a receiver. In commercial circles it is defined as the strength of the signal (in microvolts) at the input of the receiver that is required to produce a specified audio power output at the 'speaker or headphones. This is a perfectly-satisfactory definition for broadcast and communications receivers operating below about 20 Mc., where general atmospheric and man-made electrical noises normally mask any noise generated by the receiver itself.

Another commercial definition of sensitivity measures the merit of a receiver by defining the sensitivity as the signal at the input of the receiver required to give an audio output some stated amount (generally 10 db.) above the noise output of the receiver. This is a much more useful sensitivity measure for the amateur, since it indicates how well a weak signal will be reproduced and is not merely a measure of the over-all gain, or amplification, of the receiver. However, it is still not an absolute method for comparing two receivers, because the passband width of the receiver plays a large part in the result.

The random motion of the molecules in the antenna and receiver circuits generates small voltages called **thermal-agitation noise** voltages. The frequency of this noise is random and the noise exists across the entire radio spectrum. Its amplitude increases with the temperature of the circuits. Only the noise in the antenna and first stage of a receiver is normally significant, since the noise developed in later stages is masked by the amplified noise from the first stage. Since the only noise that is amplified is that which falls within the passband of the receiver, the noise appearing in the output of a receiver is less when the passband is reduced (the effect of the "tone control" of a broadcast receiver). Similar noise is generated by the current flow within the first tube itself; this effect can be combined with the thermal noise and called **receiver noise.** Since the passband of two receivers plays an important part in the sensitivity measured on a signal-to-noise basis as described in the preceding paragraph, such a sensitivity measurement puts more emphasis on passband width than on the all-important "front-end" design of the receiver.

The limit of a receiver's ability to detect weak signals is the thermal noise generated in the input circuit. Even if a perfect noise-free tube were developed and used throughout the receiver, the limit to reception would be the thermal noise. (Atmospheric-and-man-made noise is a *practical* limit below 20 Mc., but we are looking for a measure of comparison of receivers.) The degree to which a receiver approaches this ideal is called the **noise figure** of the receiver, and it is expressed as the ratio of noise power at the input of the receiver required to increase the noise output of the receiver 3 db. Since the noise power passed by the receiver is dependent on the passband (which is the same for the receiver noise and the noise introduced to the receiver), the figure is one that shows how far the receiver departs from the ideal. The ratio is generally expressed in db., and runs around 6 to 12 db. for a good receiver, although figures of 3 to 4 db. have been obtained with special techniques. Comparisons of noise figures can be made by the amateur with simple equipment, as described in Chapter Sixteen.

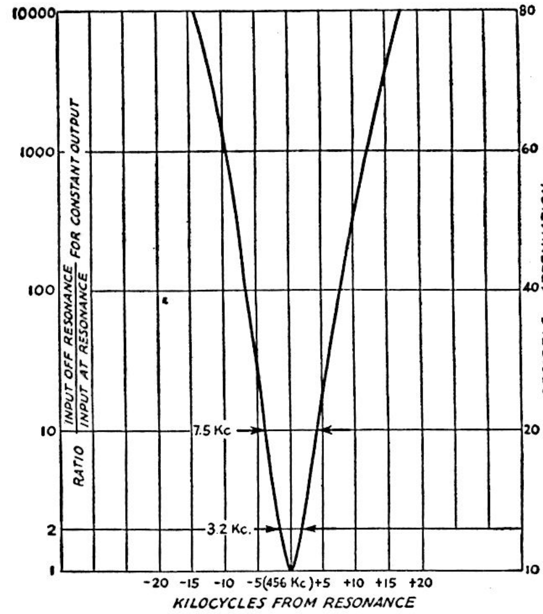



*Fig. 5-1* — **Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.**

### *Selectivity*

Selectivity is the ability of a receiver to discriminate against signals of frequencies differing from that of the desired signal. The over-all selectivity will depend upon the selectivity of the individual tuned circuits and the number of such circuits.

The selectivity of a receiver is shown graphically by drawing a curve that gives the ratio of signal strength required at various frequencies off resonance to the signal strength at resonance, to give constant output. A **resonance curve** of this type (taken on a typical communications-type superheterodyne receiver) is shown in Fig. 5-1. The **bandwidth** is the width of the resonance curve (in cycles or kilocycles) of a receiver at a specified ratio; in Fig. 5-1, the bandwidths are indicated for ratios of response of 2 and 10 ("2 times down" and "10 times down").

A receiver is more selective if the bandwidth (or passband) is less, but the bandwidth must be sufficient to pass the signal and its sidebands if faithful reproduction of the signal is desired. In the crowded amateur bands, it is generally advisable to sacrifice fidelity for selectivity, since the added selectivity reduces adjacent-channel interference and also the noise passed by the receiver. If the selectivity curve has steep sides, it is said to have good **skirt selectivity,** and this feature is very useful in listening to a weak signal that is adjacent to a strong one. Good skirt selectivity can only be obtained by using a large number of tuned circuits.

### *Stability*

The stability of a receiver is its ability to give constant output, over a period of time, from a signal of constant strength and frequency, and also its ability to remain tuned to a signal under varying conditions of gain-control setting, temperature, supply-voltage changes and mechanical shock and distortion. In other words, it means the ability "to stay put" on a given signal. The term "unstable" is also applied to a receiver that breaks into oscillation or a regenerative condition with some settings of its controls that are not specifically intended to control such a condition. This type of instability is sometimes encountered in high-gain amplifiers.

### *Fidelity*

Fidelity is the relative ability of the receiver to reproduce in its output the modulation (keying, voice, etc.) carried by the incoming signal. For exact reproduction the bandwidth must be great enough to accommodate the carrier and all of the sidebands before detection, and all of the frequency components of the modulation after detection. For perfect fidelity, the relative amplitudes of the various components must not be changed by passing through the receiver. However, fidelity plays a very minor rôle in amateur communication, where the important requirement is to transmit intelligence and not "high-fidelity" signals.

## Detection and Detectors

Detection is the process of recovering the modulation from a signal. Any device that is "nonlinear" (i.e., whose output is not *exactly* proportional to its input) will act as a detector. It can be used as a detector if an impedance for the desired modulation frequency is connected in the output circuit, so that the detector output can develop across this impedance.

Detector sensitivity is the ratio of desired detector output to the input. Detector linearity is a measure of the ability of the detector to reproduce the exact form of the modulation on the incoming signal. The resistance or impedance of the detector is the resistance or impedance it presents to the circuits it is connected to. The input resistance is important in receiver design, since if it is relatively low it means that the detector will consume power, and this power must be furnished by the preceding stage. The signal-handling capability means the ability of the detector to accept signals of a specified amplitude without overloading or distortion.

### *Diode Detectors*

The simplest detector is the diode rectifier. A galena, silicon or germanium **crystal** is an imperfect form of diode (a small current can pass in the reverse direction), and the principle of detection in a crystal is similar to that in a vacuum-tube diode.

Circuits for both half-wave and full-wave diodes are given in Fig. 5-2. The simplified half-wave circuit at 5-2A includes the r.f. tuned circuit, $L_2C_1$, a coupling coil, $L_1$, from which the r.f. energy is fed to $L_2C_1$, and the diode, $D$, with its load resistance, $R_1$, and by-pass condenser, $C_2$. The flow of rectified r.f. current causes a d.c. voltage to develop across the terminals of $R_1$, and this voltage varies with the modulation on the signal. The $-$ and $+$ signs show the polarity of the voltage. The variation in amplitude of the r.f. signal with modulation causes corresponding variations in the value of the d.c. voltage across $R_1$. The

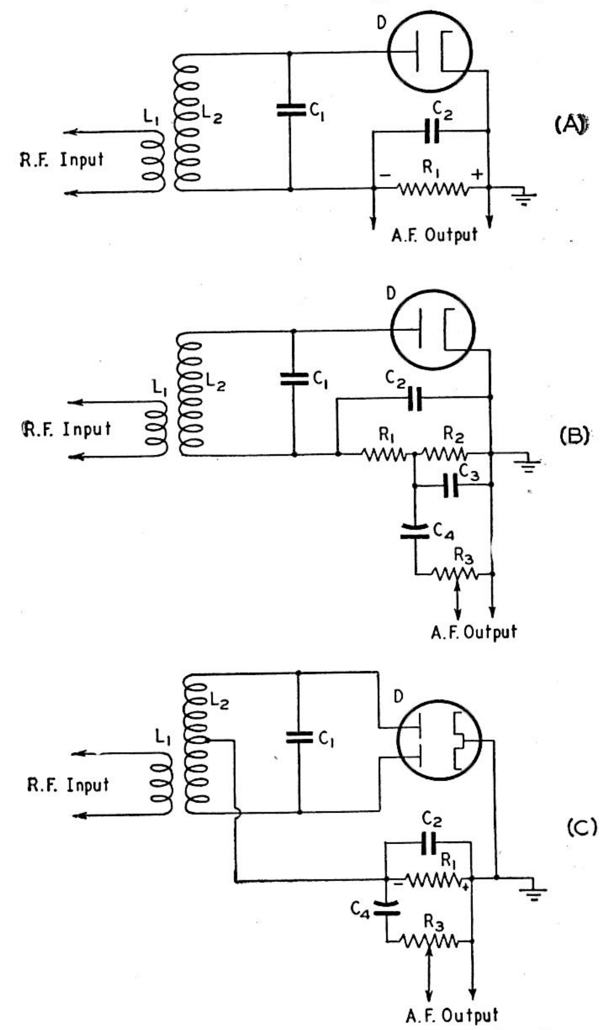



*Fig. 5-2* — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, $L_2C_1$, is tuned to the signal frequency; typical values for $C_2$ and $R_1$ in A and C are 250 $\mu\mu$fd. and 250,000 ohms, respectively; in B, $C_2$ and $C_3$ are 100 $\mu\mu$fd. each; $R_1$, 50,000 ohms; and $R_2$, 250,000 ohms. $C_4$ is 0.1 $\mu$fd. and $R_3$ may be 0.5 to 1 megohm.

load resistor, $R_1$, usually has a rather high value of resistance, so that a fairly large voltage will develop from a small rectified-current flow.

The progress of the signal through the detector or rectifier is shown in Fig. 5-3. A typical modulated signal as it exists in the tuned circuit is shown at A. When this signal is applied to the rectifier tube, current will flow only during the part of the r.f. cycle when the plate is positive with respect to the cathode, so that the output of the rectifier consists of half-cycles of r.f. still modulated as in the original signal. These current pulses flow in the load circuit comprised of $R_1$ and $C_2$, the resistance of $R_1$ and the capacity of $C_2$ being so proportioned that $C_2$ charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across $R_1$ is smoothed out, as shown in C. $C_2$ thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a d.c. component that varies in the same way as the modulation on the original signal. When this varying d.c. voltage is applied to a following amplifier through a coupling condenser ($C_4$ in Fig. 5-2B), only the *variations* in voltage are transferred, so that the final output signal is a.c., as shown in D.

In the circuit at 5-2B, $R_1$ and $C_2$ have been divided for the purpose of providing a more effective filter for r.f. It is important to prevent the appearance of any r.f. voltage in the output of the detector, because it may cause overloading of a succeeding amplifier tube. The audio-frequency variations can be transferred to another circuit through a coupling condenser, $C_4$ in Fig. 5-2B, to a load resistor, $R_3$, which usually is a "potentiometer" so that the volume can be adjusted to a desired level.

Coupling to the potentiometer (gain control) through a condenser also avoids any flow of d.c. through the gain control. The flow of d.c. through a high-resistance gain control often tends to make the control noisy (scratchy) after a short while.

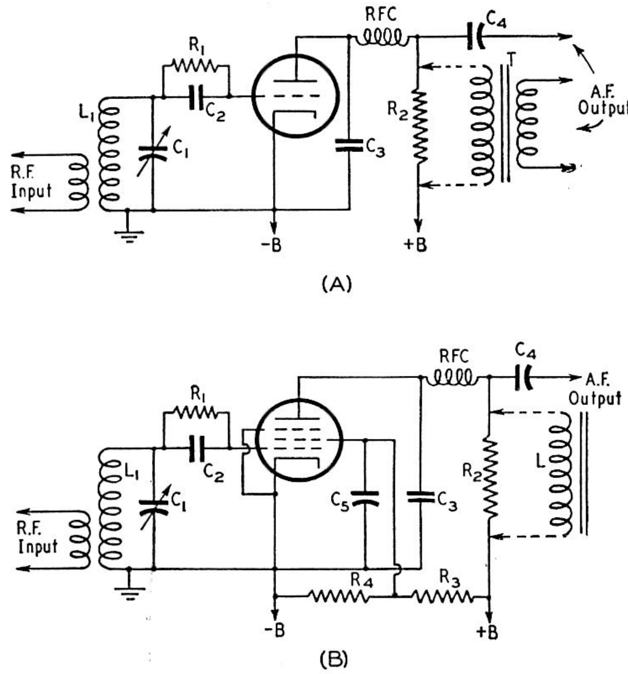



*Fig. 5-4* — **Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. $L_1C_1$ is a circuit tuned to the signal frequency. The grid leak, $R_1$, may be connected directly from grid to cathode instead of across the grid condenser as shown. The operation with either connection will be the same. Representative values for components are:**

| *Component* | *Circuit A* | *Circuit B* |
|---|---|---|
| $C_2$ | 100 to 250 $\mu\mu$fd. | 100 to 250 $\mu\mu$fd. |
| $C_3$ | 0.001 to 0.002 $\mu$fd. | 250 to 500 $\mu\mu$fd. |
| $C_4$ | 0.1 $\mu$fd. | 0.1 $\mu$fd. |
| $C_5$ | | 0.5 $\mu$fd. or larger. |
| $R_1$ | 1 to 2 megohms. | 1 to 5 megohms. |
| $R_2$ | 50,000 ohms. | 100,000 to 250,000 ohms. |
| $R_3$ | | 50,000 ohms. |
| $R_4$ | | 20,000 ohms. |
| L | | 500-henry choke. |
| RFC | 2.5 mh. | 2.5 mh. |
| T | Audio transformer. | |

**The plate voltage in A should be about 50 volts for best sensitivity. In B, the screen voltage should be about 30 volts and the plate voltage from 100 to 250.**

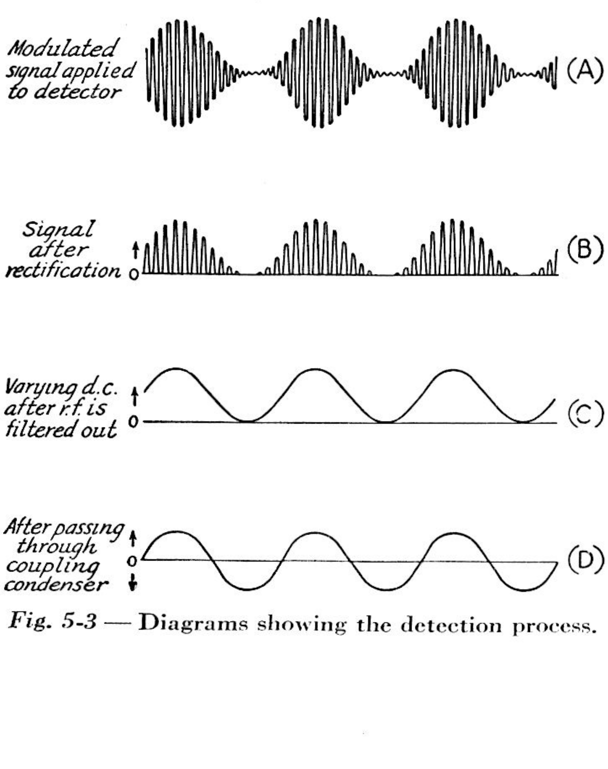



*Fig. 5-3* — Diagrams showing the detection process.

The full-wave diode circuit at 5-2C differs in operation from the half-wave circuit only in that both halves of the r.f. cycle are utilized. The full-wave circuit has the advantage that very little r.f. voltage appears across the load resistor, $R_1$, because the midpoint of $L_2$ is at the same potential as the cathode, or "ground" for r.f., and r.f. filtering is easier than in the half-wave circuit.

The reactance of $C_2$ must be small compared to the resistance of $R_1$ at the radio frequency being rectified, but at audio frequencies must be relatively large compared to $R_1$. This condition is satisfied by the values shown. If the capacity of $C_2$ is too large, response at the higher audio frequencies will be lowered.

Compared with other detectors, the sensitivity of the diode is low. Since the diode consumes power, the $Q$ of the tuned circuit is reduced, bringing about a reduction in selectivity. The linearity is good, however, and the signal-handling capability is high.

### *Grid-Leak Detectors*

The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuits of Fig. 5-4, the grid corresponds to the diode plate and the rectifying action is exactly the same as just described. The d.c. voltage from rectified-current flow through the grid leak, $R_1$, biases the grid negatively with respect to cathode, and the audio-frequency variations in voltage across $R_1$ are amplified through the tube just as in a normal a.f. amplifier. In the plate circuit, $R_2$ is the plate load resistance, $C_3$ is a by-pass condenser and $RFC$ an r.f. choke to eliminate r.f. in the output circuit. $C_4$ is the output coupling condenser. With a triode, the load resistor, $R_2$, may be replaced by an audio transformer, $T$, in which case $C_4$ is not used.

Since audio amplification is added to rectification, the grid-leak detector has considerably greater sensitivity than the diode. The sensitivity can be further increased by using a screen-grid tube instead of a triode, as at 5-4B. The operation is equivalent to that of the triode circuit. The screen by-pass condenser, $C_5$, should have low reactance for both radio and audio frequencies. $R_3$ and $R_4$ constitute a voltage divider on the plate supply to furnish the proper d.c. voltage to the screen. In both circuits, $C_2$ must have low r.f. reactance and high a.f. reactance compared to the resistance of $R_1$; the same applies to $C_3$ with respect to $R_2$. The reactance of $RFC$ will be high for r.f. and low for audio frequencies.

Because of the high plate resistance of the screen-grid tube, transformer coupling from the plate circuit of a screen-grid detector is not satisfactory. An impedance ($L$ in Fig. 5-4B) can be used in place of a resistor, with a gain in sensitivity because a high value of load impedance can be developed with little loss of plate voltage as compared to the voltage drop through a resistor. The coupling coil, $L$, for a screen-grid detector should have an inductance of the order of 300 to 500 henrys.

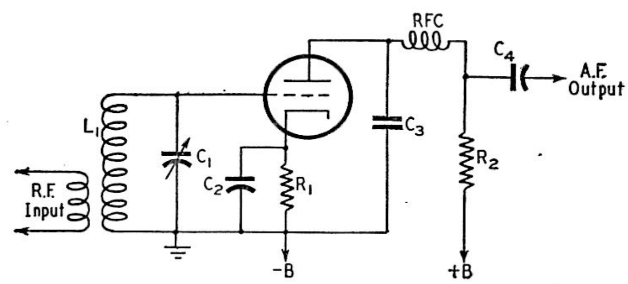
The sensitivity of the grid-leak detector is higher than that of any other type. Like the diode, it "loads" the tuned circuit and reduces its selectivity. The linearity is rather poor, and the signal-handling capability is limited. The signal-handling capability can be improved by reducing $R_1$ to 0.1 megohm, but the sensitivity will be decreased.

### *Plate Detectors*

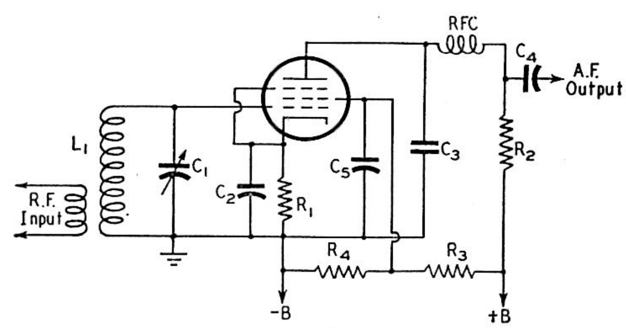
The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of the tube, as contrasted to the grid rectification just described. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cut-off point, so that the application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in signal amplitude in a fashion similar to the rectified current in a diode detector.

Circuits for triodes and pentodes are given in Fig. 5-5. $C_3$ is the plate by-pass condenser, and, with $RFC$, prevents r.f. from appearing in the output. $R_1$ is the cathode resistor which provides the operating grid bias, and $C_2$ is a by-pass for both radio and audio frequencies across $R_1$. $R_2$ is the plate load resistance across which a voltage appears as a result of the rectifiying action described above. $C_4$ is the output coupling condenser. In the pentode circuit at B, $R_3$ and $R_4$ form a voltage divider to supply the proper potential (about 30 volts) to the screen, and $C_5$ is a by-pass condenser between screen and cathode. $C_5$ must have low reactance for both radio and audio frequencies.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance even of a triode is very high when the bias is set near the plate-current cut-off point. Impedance coupling may be used in place of the resistance




(A)




(B)

*Fig. 5-5* — Circuits for plate detection. A, triode; B, pentode. The input circuit, $L_1C_1$, is tuned to the signal frequency. Typical values for the other components are:

| *Component* | *Circuit A* | *Circuit B* |
|---|---|---|
| $C_2$ | 0.5 μfd. or larger. | 0.5 μfd. or larger. |
| $C_3$ | 0.001 to 0.002 μfd. | 250 to 500 μμfd. |
| $C_4$ | 0.1 μfd. | 0.1 μfd. |
| $C_5$ | | 0.5 μfd. or larger. |
| $R_1$ | 25,000 to 150,000 ohms. | 10,000 to 20,000 ohms. |
| $R_2$ | 50,000 to 100,000 ohms. | 100,000 to 250,000 ohms. |
| $R_3$ | | 50,000 ohms. |
| $R_4$ | | 20,000 ohms. |
| RFC | 2.5 mh. | 2.5 mh. |

Plate voltages from 100 to 250 volts may be used. Effective screen voltage in B should be about 30 volts.

coupling shown in Fig. 5-5. The same order of inductance is required as with the pentode grid-leak detector described previously.

The plate detector is more sensitive than the diode since there is some amplifying action in the tube, but less so than the grid-leak detector. It will handle considerably larger signals than the grid-leak detector, but is not quite so tolerant in this respect as the diode. Linearity, with the self-biased circuits shown, is good. Up to the overload point the detector takes no power from the tuned circuit, and so does not affect its $Q$ and selectivity.

### *Infinite-Impedance Detector*

The circuit of Fig. 5-6 combines the high signal-handling capabilities of the diode detector with low distortion (good linearity), and, like the plate detector, does not load the tuned circuit it connects to. The circuit resembles that of the plate detector, except that the load resistance, $R_1$, is connected between cathode and ground and thus is common to both grid and plate circuits, giving negative feed-back for the audio frequencies. The cathode resistor is by-passed for r.f. ($C_2$) but not for audio, while the plate circuit is by-passed to ground for both audio and radio frequencies. $R_2$ forms, with $C_3$, an *RC* filter to isolate the plate from the "B" supply at a.f. An r.f. filter, consisting of a series r.f. choke and a shunt condenser, can be connected between the cathode and $C_4$ to eliminate any r.f. that might otherwise appear in the output.

The plate current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across $R_1$ similarly increases with signal, because of the increased plate current. Because of this and the fact that the initial drop across $R_1$ is large, the grid cannot be driven positive with respect to the cathode by the signal, hence no grid current can be drawn.

## REGENERATIVE DETECTORS

By providing controllable r.f. feed-back or regeneration in a triode or pentode detector circuit, the incoming signal can be amplified many times, thereby greatly increasing the sensitivity of the detector. Regeneration also increases the effective $Q$ of the circuit and increases the selectivity because the maximum regenerative amplification takes place only at the frequency to which the circuit is tuned. The grid-leak type of detector is most suitable for the purpose. Except for the regenerative connection, the circuit values are identical with those previously described for this type of detector, and the same considerations apply. The amount of regeneration must be controllable, because maximum regenerative amplification is secured at the critical point where the circuit is just about to oscillate, and the critical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned. In the oscillating condition, a regenerative detector can be detuned slightly from an incoming c.w. signal to give *autodyne* reception.

Fig. 5-7 shows the circuits of regenerative detectors of various types. The circuit of A is for a triode tube, with a variable by-pass condenser, $C_3$, in the plate circuit to control regeneration. When the capacity is small the tube does not regenerate, but as it increases toward maximum its reactance becomes smaller until a critical value is reached where there is sufficient feed-back to cause oscillation. If $L_2$ and $L_3$ are wound end-to-end in the same direction, the plate connection is to the outside of the plate or "tickler" coil, $L_3$, when the grid connection is to the outside of $L_2$.

The circuit of 5-7B is for a pentode tube, regeneration being controlled by adjustment of the screen-grid voltage. The tickler, $L_3$, is in the plate circuit. The portion of the control resistor between the rotating contact and ground is by-passed by a large condenser (0.5 μfd. or more) to filter out scratching noise when the arm is rotated. The feed-back is adjusted by varying the number of turns on $L_3$ or the coupling between $L_2$ and $L_3$, until the tube just goes into oscillation at a screen potential of approximately 30 volts.

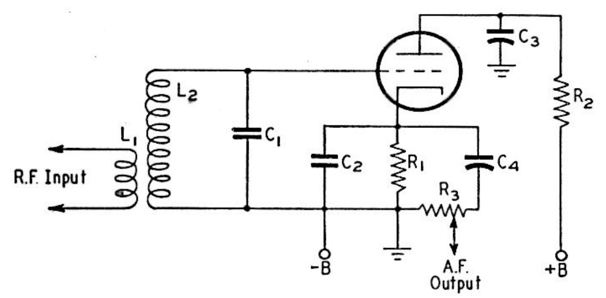



**Fig. 5-6 — The infinite-impedance detector. The input circuit, $L_2C_1$, is tuned to the signal frequency. Typical values for the other components are:**

| | |
|---|---|
| $C_2$ — 250 μμfd. | $R_1$ — 0.15 megohm. |
| $C_3$ — 0.5 μfd. | $R_2$ — 25,000 ohms. |
| $C_4$ — 0.1 μfd. | $R_3$ — 0.25-megohm volume control. |

**A tube having a medium amplification factor (about 20) should be used. Plate voltage should be 250 volts.**

Circuit C is identical with B in principle of operation, except that the oscillating circuit is of the Hartley type. Since the screen and plate are in parallel for r.f. in this circuit, only a small amount of "tickler" — that is, relatively few turns between the cathode tap and ground — is required for oscillation.

### *Smooth Regeneration Control*

The ideal regeneration control would permit the detector to go into and out of oscillation smoothly, would have no effect on the frequency of oscillation, and would give the same value of regeneration regardless of frequency and the loading on the circuit. In practice, the effects of loading, particularly the loading that occurs when the detector circuit is coupled to an antenna, are difficult to overcome. Likewise, the regeneration is usually affected by the frequency to which the grid circuit is tuned.

In all circuits it is best to wind the tickler at the ground or cathode end of the grid coil, and to use as few turns on the tickler as will allow the detector to oscillate easily over the whole tuning range at the plate (and screen, if a pentode) voltage that gives maximum sensitivity. Should the tube break into oscillation suddenly as the regeneration control is advanced, making a click, the operation often can be made smoother by changing the grid-leak resistance to a higher or lower value. The wrong grid leak plus too-high plate and screen voltage are the most frequent causes of lack of smoothness in going into oscillation.

### *Antenna Coupling*

If the detector is coupled to an antenna, slight changes in the antenna constants (as when the wire swings in a breeze) affect the frequency of the oscillations generated, and thereby the beat frequency when c.w. signals are being received. The tighter the antenna coupling is made, the greater will be the feedback required or the higher will be the voltage necessary to make the detector oscillate. The antenna coupling should be the maximum that will allow the detector to go into oscillation smoothly with the correct voltages on the tube. If capacity coupling to the grid end of the coil is used, only a very small amount of capacity will be needed to couple to the antenna. Increasing the capacity increases the coupling.

At frequencies where the antenna system is resonant the absorption of energy from the oscillating detector circuit will be greater, with the consequence that more regeneration is needed. In extreme cases it may not be possible to make the detector oscillate with normal voltages, causing so-called "dead spots." The remedy for this is to loosen the antenna coupling to the point that permits normal oscillation and smooth regeneration control.

### *Body Capacity*

A regenerative detector occasionally shows a tendency to change frequency slightly as the hand is moved near the dial. This condition (**body capacity**) can be caused by poor design of the receiver, or by the antenna if the detector is coupled directly to it. If body capacity is present when the antenna is disconnected, it can be eliminated by better shielding, and sometimes by r.f. filtering of the 'phone leads. Body capacity that is present only when the antenna is connected is caused by resonance effects in the antenna, which tend to raise the whole detector circuit above ground potential. A good, short ground connection should be made to the receiver and the length of the antenna varied electrically (by adding a small coil or variable condenser in the antenna lead) until the effect is minimized. Loosening the coupling to the antenna circuit also will help.

### *Hum*

Hum at the power-supply frequency may be present in a regenerative detector, especially when it is used in an oscillating condition for c.w. reception, even though the plate supply itself is free from ripple. The hum may result from the use of a.c. on the tube heater, but effects of this type normally are troublesome

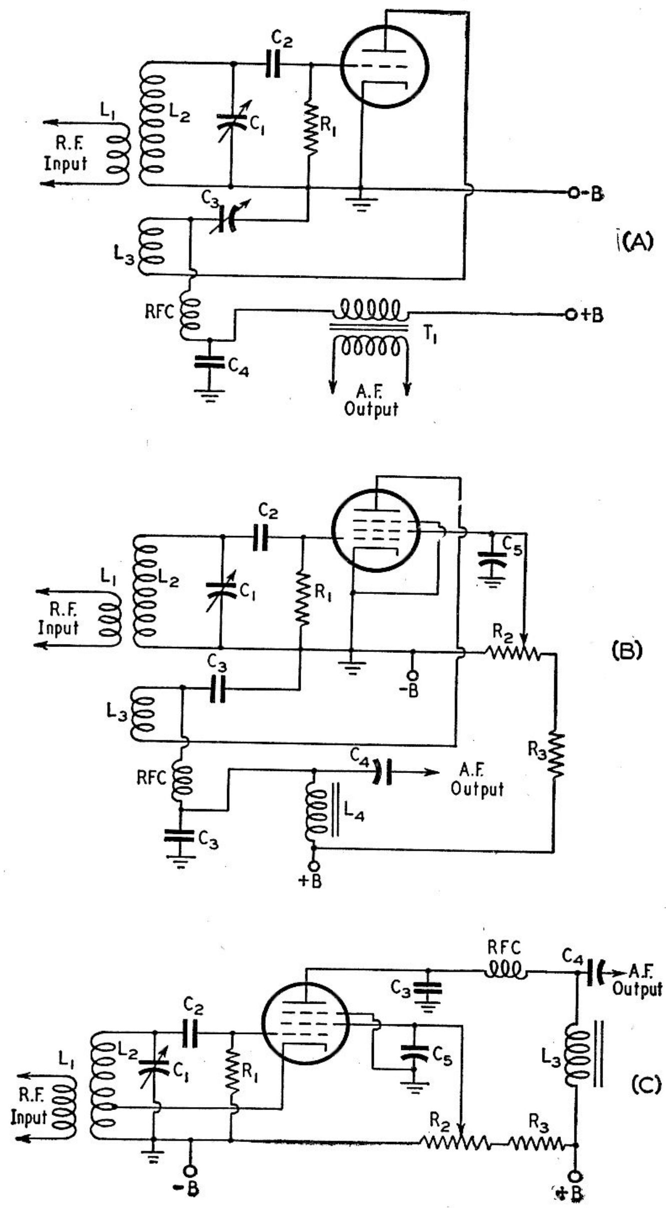



*Fig. 5-7* — Triode and pentode regenerative detector circuits. The input circuit, $L_2C_1$, is tuned to the signal frequency. The grid condenser, $C_2$, should have a value of about 100 $\mu\mu$fd. in all circuits; the grid leak, $R_1$, may range in value from 1 to 5 megohms. The tickler coil, $L_3$, ordinarily will have from 10 to 25 per cent of the number of turns on $L_2$; in C, the cathode tap is about 10 per cent of the number of turns on $L_2$ above ground. Regeneration-control condenser $C_3$ in A should have a maximum capacity of 100 $\mu\mu$fd. or more; by-pass condensers $C_3$ in B and C are likewise 100 $\mu\mu$fd. $C_5$ is ordinarily 1 $\mu$fd. or more; $R_2$, a 50,000-ohm potentiometer; $R_3$, 50,000 to 100,000 ohms. $L_4$ in B ($L_3$ in C) is a 500-henry inductance, $C_4$ is 0.1 $\mu$fd. in both circuits. $T_1$ in A is a conventional audio transformer for coupling from the plate of a tube to a following grid. *RFC* is 2.5 mh. In A, the plate voltage should be about 50 volts for best sensitivity. Pentode circuits require about 30 volts on the screen; plate potential may be 100 to 250 volts.

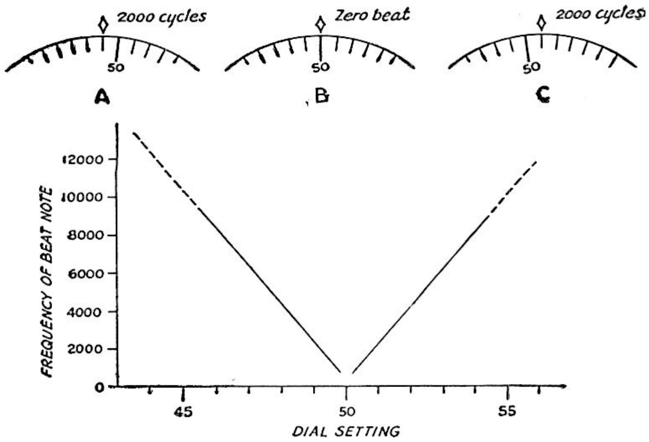



*Fig. 5-8* — As the tuning dial of a receiver is turned past a c.w. signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.

only when the circuit of Fig. 5-7C is used, and then only at 14 Mc. and higher frequencies. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater-transformer winding, is good practice to reduce hum, and the heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, will have a rather extensive electrostatic field which may cause hum if the detector tube, grid lead, and grid condenser and leak are not electrostatically shielded. This type of hum is easily recognizable because of its rather high pitch (a result of harmonics in the power-supply system).

Antenna resonance effects frequently cause a hum of the same nature as that just described which is most intense at the various resonance points, and hence varies with tuning. For this reason it is called **tunable hum.** It is prone to occur with a rectified-a.c. plate supply, when the receiver is put "above ground" by the antenna, as described in a preceding paragraph. The effect is associated with the nonlinearity of the rectifier tube in the plate supply. Elimination of antenna resonance effects as described and by-passing the rectifier plates to cathode (using by-pass condensers of the order of 0.001 $\mu$fd.) usually will cure it.

### *Tuning*

For c.w. reception, the regeneration control is advanced until the detector breaks into a "hiss," which indicates that the detector is oscillating. Further advancing the regeneration control after the detector starts oscillating will result in a slight decrease in the strength of the hiss, indicating that the sensitivity of the detector is decreasing.

The proper adjustment of the regeneration control for best reception of c.w. signals is where the detector just starts to oscillate, when it will be found that c.w. signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. As the receiver is tuned through a signal the tone first will be heard as a very high pitch, then will go down through "zero beat" (the region where the frequencies of the incoming signal and the oscillating detector are so nearly alike that the difference or beat is less than the lowest audible tone) and rise again on the other side, finally disappearing at a very high pitch. This behavior is shown in Fig. 5-8. It will be found that a low-pitched beat-note cannot be obtained from a strong signal because the detector **"pulls in"** or **"blocks"**; that is, the signal tends to control the detector in such a way that the latter oscillates at the signal frequency, despite the fact that the circuit may not be tuned exactly to resonance. This phenomenon, commonly observed when an oscillator is coupled to a source of a.c. voltage of approximately the frequency at which the oscillator is operating, is called **"locking-in"**; the more stable of the two frequencies assumes control over the other. "Blocking" usually can be corrected by advancing the regeneration control until the beat-note occurs again. If the regenerative detector is preceded by an r.f. amplifier stage, the blocking can be eliminated by reducing the gain of the r.f. stage. If the detector is coupled to an antenna, the blocking condition can be satisfactorily eliminated by advancing the regeneration control or loosening the antenna coupling.

The point just after the detector starts oscillating is the most sensitive condition for c.w. reception. Further advancing the regeneration control makes the receiver less prone to blocking by strong signals, but also less capable of receiving weak signals.

If the detector is in the oscillating condition and a 'phone signal is tuned in, a steady audible beat-note will result. While it is possible to listen to 'phone if the receiver can be tuned to exact zero beat, it is more satisfactory to reduce the regeneration to the point just before the receiver goes into oscillation. This is also the most sensitive operating point.

## Audio-Frequency Amplifiers

The ordinary detector does not produce very much audio-frequency power output — usually not enough to give satisfactory sound volume, even in headphone reception. Consequently, audio-frequency amplifiers are used after the detector to increase the power level. One amplifier usually is sufficient for headphones, but two stages generally are used where the receiver is to operate a loudspeaker. A few milliwatts of a.f. power are sufficient for headphones, but a loudspeaker requires a watt or more for good room volume.

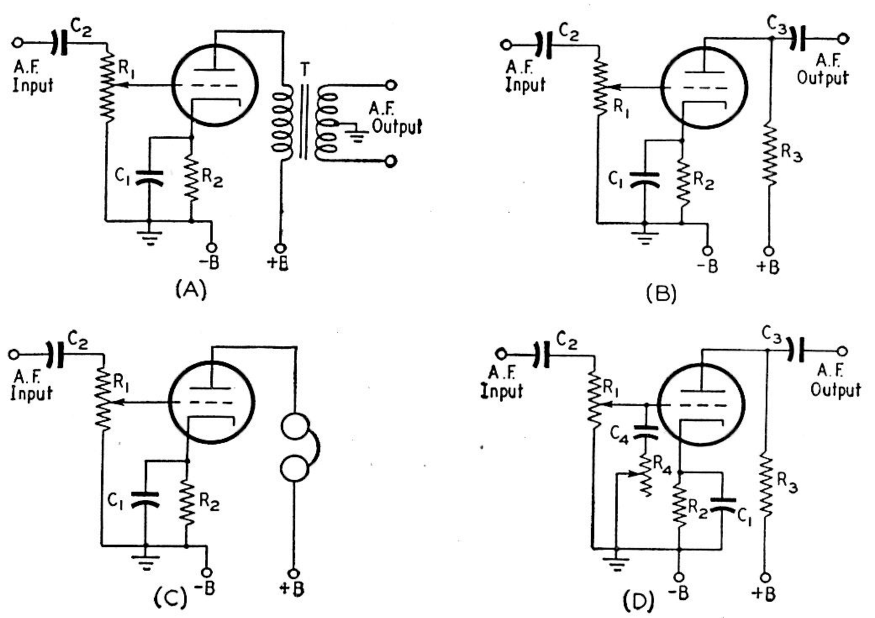



*Fig. 5-9* — **Audio-amplifier circuits used for voltage amplification and to provide power for headphone output. The tubes are operated as Class A amplifiers. A simple tone-control circuit ($R_4C_4$) is shown in D.**

### *Headset and Voltage Amplifiers*

The circuits shown in Fig. 5-9 are typical of those used for voltage amplification and for providing sufficient power for operation of headphones. Triodes usually are preferred to pentodes because they are better suited to working into an audio transformer or headset, the input impedances of which are of the order of 20,000 ohms.

In these circuits, $R_2$ is the cathode bias resistor and $C_1$ the cathode by-pass condenser. The grid resistor, $R_1$, gives volume-control action. Its value ordinarily is from 0.25 to 1 megohm. $C_2$ is the input coupling condenser, already discussed under detectors; it is, in fact, identical to $C_4$ in Figs. 5-4 and 5-5, if the amplifier is coupled to a detector. In 5-9D, $C_4$ and $R_4$ are a simple **"tone-control"** circuit. As $R_4$ is made smaller, $C_4$ by-passes more of the high audio frequencies. $R_4$ should be large compared to the reactance of $C_4$ at the highest audio frequency.

In all receivers using tubes with indirectly-heated cathodes, the negative grid bias of audio amplifiers usually is secured from the voltage drop in a cathode resistor. The cathode resistor must be by-passed by a condenser having low reactance at the lowest audio frequency to be amplified, compared to the resistance of the cathode resistor (10 per cent or less). In battery-operated receivers, which use filament-type tubes, a separate grid-bias battery generally is used.

### *Power Amplifiers*

A popular type of power amplifier is the single tetrode, operated Class A or AB; the circuit diagram is given in Fig. 5-10A. The grid resistor, $R_1$, may be a potentiometer for volume control, as shown at $R_1$ in Fig. 5-9. The output transformer, $T$, should have a turns ratio suitable for the loudspeaker used; many of the small loudspeakers now available are furnished complete with output transformer.

When greater volume is needed, a pair of tetrodes or pentodes may be connected in push-pull, as shown in Fig. 5-10B. Transformer coupling to the voltage-amplifier stage is the simplest method of obtaining push-pull input for the amplifier grids. The interstage transformer, $T_1$, has a center-tapped secondary with a secondary-to-primary turns ratio of about 2 to 1. An output transformer, $T_2$, with a center-tapped primary must be used. No by-pass condenser is needed across the cathode resistor, $R$, in Class A operation since the a.f. current does not flow through the resistor as it does in single-tube circuits.

### *Headphones and Loudspeakers*

Two types of headphones are in general use, **the magnetic** and **crystal** types. They are shown in cross-section in Fig. 5-11. In the magnetic type the signal is applied to a coil or pair of coils having a great many turns of fine wire wound on a permanent magnet. (Headphones having one coil are known as the "single-pole" type, while those having two coils, as shown in Fig. 5-11, are called "double-

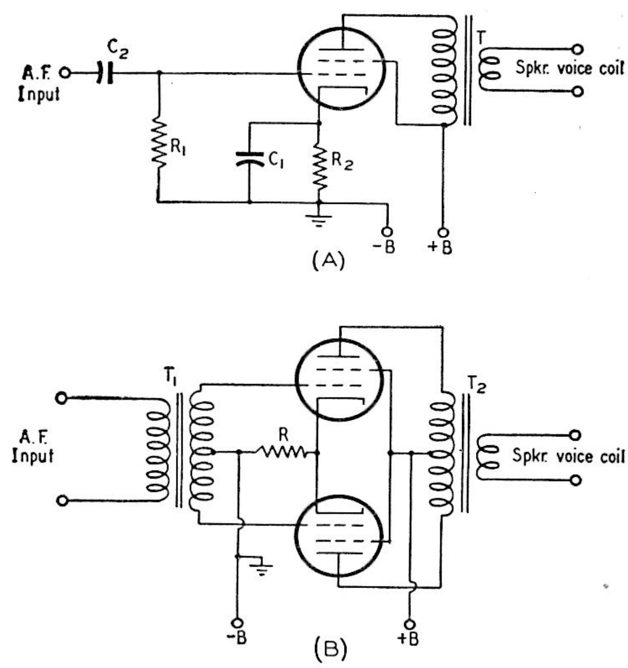



*Fig. 5-10* — **Power-output audio-amplifier circuits. Either Class A or AB amplification may be used.**

pole.") A thin circular diaphragm of iron is placed close to the open ends of the magnet. It is tightly clamped by the earpiece assembly around its circumference, and the center is drawn toward the permanent magnet under some tension. When an alternating current flows through the windings the field set up by the current alternately aids and opposes the steady field of the permanent magnet, so that

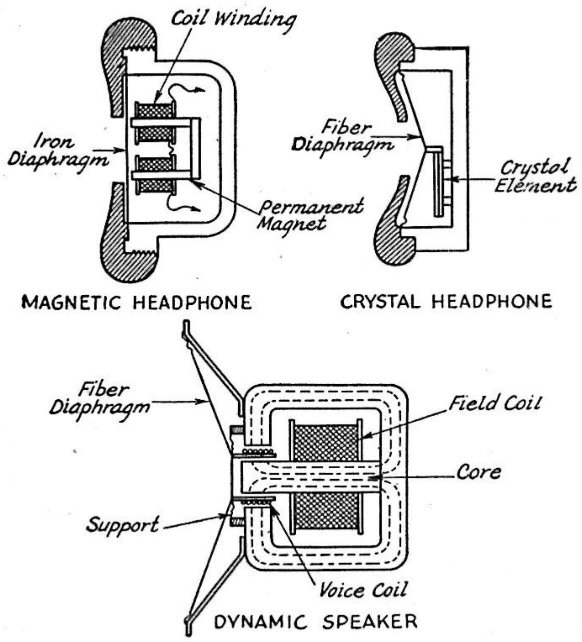



*Fig. 5-11* — Headphone and loudspeaker construction.

the diaphragm alternately is drawn nearer to and allowed to spring farther away from the magnet. Its motion sets the air into corresponding vibration. Although the d.c. resistance of the coils may be of the order of 2000 ohms, the a.c. impedance of a magnetic-type headset will be of the order of 20,000 ohms at 1000 cycles.

In the crystal headphone, two piezoelectric crystals made of Rochelle salts are cemented together in such a way that the pair tend to be bent in one direction when a voltage of a certain polarity is applied and to bend in the other direction when the polarity is reversed. The crystal unit is rigidly mounted to the earpiece, with the free end coupled to a diaphragm. When an alternating voltage is applied, the alternate bending as the polarity of the applied voltage reverses makes the diaphragm vibrate back and forth. The impedance is several times that of the magnetic type.

Magnetic-type headsets tend to give maximum response at frequencies of the order of 500 to 1000 cycles, with a considerable reduction of response at frequencies both above and below this region. The crystal type has a "flatter" frequency-response curve, and is particularly good at reproducing the higher audio frequencies. The peaked response curve of the magnetic type is advantageous in code reception, since it tends to reduce interference from signals having beat tones lying outside the region of maximum response, while the crystal type is better for the reception of voice and music. Magnetic headsets can be used in circuits in which d.c. is flowing, such as the plate circuit of a vacuum tube, provided the current is not too large to be carried safely by the wire in the coils; the limit is a few milliamperes. Crystal headsets must be used only on a.c. (since a steady d.c. voltage will damage the crystal unit), and consequently must be coupled to the tube through a device, such as a condenser, that isolates the d.c. voltage but permits the passage of an alternating current.

The most common type of loudspeaker is the **dynamic** type, shown in cross-section in Fig. 5-11. The signal is applied to a small coil (the **voice coil**) which is free to move in the gap between the ends of a magnet. The magnet is made in the form of a cylindrical coil, slightly smaller than the form on which the voice coil is wound, with the magnetic circuit completed through a pole piece which fits around the outside of the voice coil, leaving just enough clearance for free movement of the coil. The path of the flux through the magnet is as shown by the dotted lines in the figure. The voice coil is supported so that it is free to move along its axis but not in other directions, and is fastened to a fiber or paper conical diaphragm. When current is sent through the coil it moves in a direction determined by the polarity of the current, and thus moves back and forth when an alternating voltage is applied. The motion is transmitted by the diaphragm to the air, setting up sound waves.

The type of 'speaker shown in Fig. 5-11 obtains its fixed magnetic field by electromagnetic means, direct current being sent through the **field coil** for this purpose. Other types use permanent magnets to replace the electromagnet, and hence do not require a source of d.c. power. The voice coils of dynamic 'speakers have few turns and therefore low impedance, values of 3 to 15 ohms being representative.

# Tuning and Band-Changing Methods

### *Band-Changing*

The resonant circuits that are tuned to the frequency of the incoming signal constitute a special problem in the design of amateur receivers, since the amateur frequency assignments consist of groups or bands of frequencies at widely-spaced intervals. The same *LC* combination cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-minimum capacity ratio required, and also because the tuning would be excessively critical with such a large frequency range. It is necessary, therefore, to provide a means for

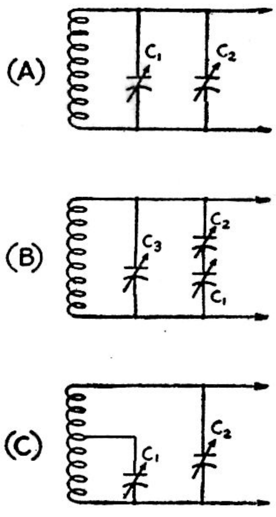



*Fig. 5-12* — Essentials of the three basic bandspread tuning systems.

changing the circuit constants for various frequency bands. As a matter of convenience the same tuning condenser usually is retained, but new coils are inserted in the circuit for each band.

One method of changing inductances is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. Another· is to use coils wound on forms with contacts (usually pins) which can be plugged in and removed from a socket.

### *Bandspreading*

The tuning range of a given coil and variable condenser will depend upon the inductance of the coil and the change in tuning capacity. For ease of tuning, it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called **bandspreading.** Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each band. Several of these methods are shown in Fig. 5-12.

In A, a small **bandspread condenser,** $C_1$ (15- to 25-$\mu\mu$fd. maximum capacity), is used in parallel with a condenser, $C_2$, which is usually large enough (100 to 140 $\mu\mu$fd.) to cover a 2-to-1 frequency range. The setting of $C_2$ will determine the minimum capacity of the circuit, and the maximum capacity for bandspread tuning will be the maximum capacity of $C_1$ plus the setting of $C_2$. The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate bandspread. In practicable circuits it is almost impossible, because of the nonharmonic relation of the various bands, to get full bandspread on all bands with the same pair of condensers, especially when the coils are wound to give continuous frequency coverage on $C_2$, which is variously called the **band-setting** or **main-tuning** condenser. $C_2$ must be reset each time the band is changed.

The method shown at B makes use of condensers in series. The tuning condenser, $C_1$, may have a maximum capacity of 100 $\mu\mu$fd. or more. The minimum capacity is determined principally by the setting of $C_3$, which usually has low capacity, and the maximum capacity by the setting of $C_2$, which is of the order of 25 to 50 $\mu\mu$fd. This method is capable of close adjustment to practically any desired degree of bandspread. Either $C_2$ and $C_3$ must be adjusted for each band or separate preadjusted condensers must be switched in.

The circuit at C also gives complete spread on each band. $C_1$, the bandspread condenser, may have any convenient value of capacity; 50 $\mu\mu$fd. is satisfactory. $C_2$ may be used for continuous frequency coverage ("general coverage") and as a band-setting condenser. The effective maximum-minimum capacity ratio depends upon the capacity of $C_2$ and the point at which $C_1$ is tapped on the coil. The nearer the tap to the bottom of the coil, the greater the bandspread, and vice versa. For a given coil and tap, the bandspread will be greater if $C_2$ is set at larger capacity. $C_2$ may be mounted in the plug-in coil form and preset, if desired. This requires a separate condenser for each band, but eliminates the necessity for resetting $C_2$ each time the band is changed.

### *Ganged Tuning*

The tuning condensers of the several r.f. circuits may be coupled together mechanically and operated by a single control. However, this operating convenience involves more complicated construction, both electrically and mechanically. It becomes necessary to make the various circuits **track** — that is, tune to the same frequency at each setting of the tuning control.

True tracking can be obtained only when the inductance, tuning condensers, and circuit inductances and minimum and maximum capacities are identical in all "ganged" stages. A small **trimmer** or **padding** condenser may be connected across the coil, so that variations in minimum capacity can be compensated. The fundamental circuit is shown in Fig. 5-13, where $C_1$ is the trimmer and $C_2$ the tuning condenser. The use of the trimmer necessarily

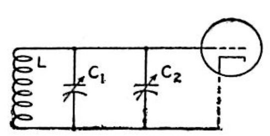



*Fig. 5-13* — Showing the use of a trimmer condenser, to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

increases the minimum circuit capacity, but it is a necessity for satisfactory tracking. Midget condensers having maximum capacities of 15 to 30 $\mu\mu$fd. are commonly used.

The same methods are applied to bandspread circuits that must be tracked. The circuits are identical with those of Fig. 5-12. If both general-coverage and bandspread tuning are to be available, an additional trimmer condenser must be connected across the coil in each circuit shown. If only amateur-band tuning is desired, however, then $C_3$ in Fig. 5-12B, and $C_2$ in Fig. 5-12C, serve as trimmers.

The coil inductance can be adjusted by starting with a larger number of turns than necessary and removing a turn or fraction of a turn at a time until the circuits track satisfactorily. An alternative method, provided the inductance is reasonably close to the correct value initially, is to make the coil so that the last turn is variable with respect to the whole

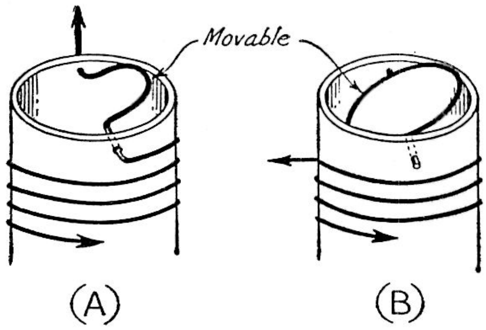



*Fig. 5-14* — Methods of adjusting the inductance for ganging. The half-turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the axis of the loop is perpendicular to the axis of the coil, and will give maximum reduction of the coil inductance when rotated 90°. The loop can be a solid disk of metal and give exactly the same effect.

coil, or to use a single short-circuited turn the position of which can be varied with respect to the coil. The application of these methods is shown in Fig. 5-14.

Still another method for trimming the inductance is to use an adjustable brass (or copper) or powdered-iron core. The brass core acts like a single shorted turn, and the inductance of the coil is decreased as the brass core, or **"slug,"** is moved into the coil. The powdered-iron core has the opposite effect, and *increases* the inductance as it is moved into the coil. The $Q$ of the coil is not affected materially by the use of the brass slug, provided the brass slug has a clean surface or is silverplated. The use of the powdered-iron core will actually raise the $Q$ of a coil, provided the iron core is of a type suitable for the frequency in use. Good powdered-iron cores can be obtained for use up to about 50 Mc.

# The Superheterodyne

For many years (up to about 1932) practically the only type of receiver to be found in amateur stations consisted of a regenerative detector and one or more stages of audio amplification. Receivers of this type can be made quite sensitive but they are lacking in stability and selectivity, particularly on the higher frequencies. Strong signals block them easily and, in our present crowded bands, they are seldom used except in emergencies. They have been replaced by **superheterodyne** receivers, generally called "superhets."

### *The Superheterodyne Principle*

In a superheterodyne receiver, the frequency of the incoming signal is changed to a new radio frequency, the **intermediate frequency** (abbreviated **"i.f."**), then amplified, and finally detected. The frequency is changed by means of the heterodyne process, the output of a tunable oscillator (the **high-frequency,** or **local, oscillator**) being combined with the incoming signal in a **mixer** or **converter** stage (**first detector**) to produce a beat frequency equal to the intermediate frequency. The audio-frequency signal is obtained at the **second detector.** C.w. signals are made audible by autodyne or heterodyne reception at the second detector.

As a numerical example, assume that an intermediate frequency of 455 kc. is chosen and that the incoming signal is on 7000 kc. Then the high-frequency oscillator frequency may be set to 7455 kc., in order that the beat frequency (7455 minus 7000) will be 455 kc. The high-frequency oscillator could also be set to 6545 kc. and give the same difference frequency. To produce an audible c.w. signal at the second detector of, say, 1000 cycles, the autodyning or heterodyning oscillator would be set to either 454 or 456 kc.

The frequency-conversion process permits r.f. amplification at a relatively low frequency, the i.f. High selectivity and gain can be obtained at this frequency, and this selectivity and gain are constant. The separate oscillators can be designed for stability and, since the h.f. oscillator is working at a frequency considerably removed from the signal frequency, its stability is practically unaffected by the incoming signal.

### *Images*

Each h.f. oscillator frequency will cause i.f. response at two signal frequencies, one higher and one lower than the oscillator frequency. If the oscillator is set to 7455 kc. to tune to a 7000-kc. signal, for example, the receiver can respond also to a signal on 7910 kc., which likewise gives a 455-kc. beat. The resultant undesired signal of the two frequencies is called the **image.**

The radio-frequency circuits of the receiver (those used before the frequency is converted to the i.f.) normally are tuned to the desired signal, so that the selectivity of the circuits reduces or eliminates the response to the image signal. The ratio of the receiver voltage output from the desired signal to that from the image is called the **signal-to-image ratio,** or **image ratio.**

The image ratio depends upon the selectivity of the r.f. tuned circuits preceding the mixer tube. Also, the higher the intermediate frequency, the higher the image ratio, since raising the i.f. increases the frequency separation between the signal and the image and places the latter further away from the resonance peak of the signal-frequency input circuits. Most receiver designs represent a compromise between economy (few r.f. stages) and image rejection (large number of r.f. stages).

### *Other Spurious Responses*

In addition to images, other signals to which the receiver is not ostensibly tuned may be heard. Harmonics of the high-frequency oscillator may beat with signals far removed from the desired frequency to produce output at the intermediate frequency; such spurious responses can be reduced by adequate selectivity before the mixer stage, and by using sufficient shielding to prevent signal pick-up by any means other than the antenna. When a strong signal is received, the harmonics generated by rectification in the second detector may, by stray coupling, be introduced into the r.f. or mixer circuit and converted to the intermediate frequency, to go through the receiver in the same way as an ordinary signal. These "birdies" appear as a heterodyne beat on the desired signal, and are principally bothersome when the frequency of the incoming signal is not greatly different from the intermediate frequency. The cure is proper circuit isolation and shielding.

Harmonics of the beat oscillator also may be converted in similar fashion and amplified through the receiver; these responses can be reduced by shielding the beat oscillator and operating it at low output level.

### *The Double Superheterodyne*

At high and very-high frequencies it is difficult to secure an adequate image ratio when the intermediate frequency is of the order of 455 kc. To reduce image response the signal frequently is converted first to a rather high (1500, 5000, or even 10,000 kc.) intermediate frequency, and then — sometimes after further amplification — reconverted to a lower i.f. where higher adjacent-channel selectivity can be obtained. Such a receiver is called a **double superheterodyne.**

## ● FREQUENCY CONVERTERS

The first detector or mixer resembles an ordinary detector. A circuit tuned to the intermediate frequency is placed in the plate circuit of the mixer, to offer a high impedance to the i.f. voltage that is developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are by-passed to ground, since they are not wanted in the output. The i.f. tuned circuit should have low impedance for these frequencies, a condition easily met if they do not approach the intermediate frequency.

The **conversion efficiency** of the mixer is the ratio of i.f. output voltage from the plate circuit to r.f. signal voltage applied to the grid. High conversion efficiency is desirable. The mixer tube noise also should be low if a good signal-to-noise ratio is wanted, particularly if the mixer is the first tube in the receiver.

The mixer should not require too much r.f. power from the h.f. oscillator, since it may be difficult to supply the power and yet maintain good oscillator stability. Also, the conversion efficiency should not depend too critically on the oscillator voltage (that is, a small change in oscillator output should not change the gain), since it is difficult to maintain constant output over a wide frequency range.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called **pulling.** If the mixer and oscillator could be completely isolated, mixer tuning would have no effect on the oscillator frequency; but in practice this is a difficult condition to attain. Pulling should be minimized, because the stability of the whole receiver depends critically upon the stability of the h.f. oscillator. Pulling decreases with separation of the signal and h.f.-oscillator frequencies, being less with high intermediate frequencies. Another type of pulling is caused by regulation in the power supply. Strong signals cause the supply voltage to change, and this in turn shifts the oscillator frequency.

### *Circuits*

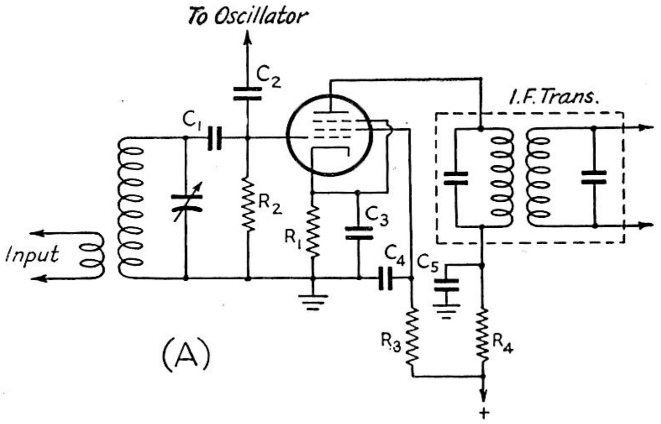
If the first detector and high-frequency oscillator are separate tubes, the first detector is called a "mixer." If the two are combined in one envelope (as is often done for reasons of economy or efficiency), the first detector is called a "converter." In either case the function is the same, however.

Typical mixer circuits are shown in Fig. 5-15. The variations are chiefly in the way in which the oscillator voltage is introduced. In 5-15A, a pentode functions as a plate detector; the oscillator voltage is capacity-coupled to the grid of the tube through $C_2$. Inductive coupling may be used instead. The conversion gain and input selectivity generally are good, so long as the sum of the two voltages (signal and oscillator) impressed on the mixer grid does not exceed the grid bias. It is desirable to make the oscillator voltage as high as possible without exceeding this limitation. The oscillator power required is negligible. If the signal frequency is only 5 or 10 times the i.f., it may be difficult to develop enough oscillator voltage at the grid (because of the selectivity of the tuned input circuit). However, the circuit is a sensitive one and makes a good mixer, particularly with high-$G_m$ tubes like the 6AC7 and 6AK5. A good triode also works well in the circuit, and

**TABLE 5-I**
**Circuit and Operating Values for Converter Tubes**

| *Tube* | *Plate Volts* | *Screen Volts* | *Cathode Resistor* | *Screen Resistor* | *Grid Leak* | *Grid I* |
|---|---|---|---|---|---|---|
| 6BE6 | 250 | 100 | 100[1,2] | 22,000 | 22,000 | 0.5 ma. |
| 6K8 | 250 | 100 | 240[1] | 27,000 | 47,000 | 0.15-0.2 |
| 6SA7 | 250 | 100 | 0[1]<br>160[2] | 18,000 | 22,000 | 0.5 |
| 6SB7Y | 250 | 100 | 0[1]<br>75[2] | 12,000<br>15,000 | 22,000 | 0.35 |

[1]Self-excited. [2]Separate excitation.



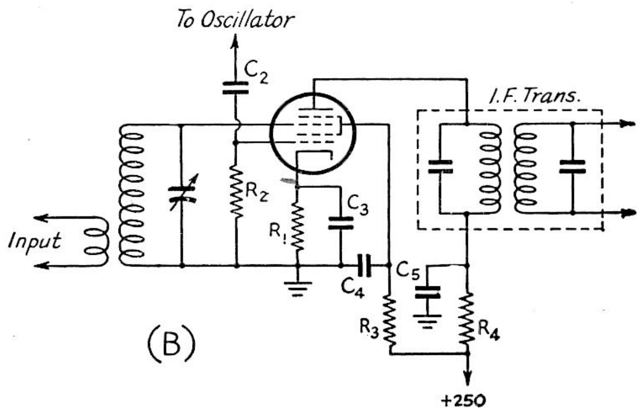




*Fig. 5-15* — Typical circuits for separately-excited mixers. Grid injection of a pentode mixer is shown at A, and separate excitation of a pentagrid converter is given in B. Typical values for B will be found in Table 5-I — the values below are for the pentode mixer of A.

| | |
|---|---|
| $C_1$ — 10 to 50 $\mu\mu$fd. | $R_2$ — 1.0 megohm. |
| $C_2$ — 5 to 10 $\mu\mu$fd. | $R_3$ — 0.47 megohm. |
| $C_3$, $C_4$, $C_5$ — 0.001 $\mu$fd. | $R_4$ — 1500 ohms. |
| $R_1$ — 6800 ohms. | |

Positive supply voltage can be 250 volts with a 6AC7, 150 with a 6AK5.

tubes like the 7F8 (one section), the 6J6 (one section) and the 6J4 work well. When a triode is used, care should be taken to see that the signal frequency is short-circuited in the plate circuit, and this is done by mounting the tuning capacitor of the i.f. transformer directly from plate to cathode.

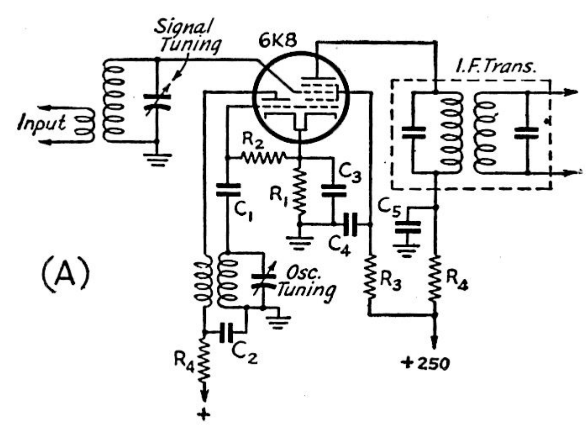
It is difficult to avoid "pulling" in a triode or pentode mixer, however, and a pentagrid converter tube used as a mixer provides much better isolation. A typical circuit is shown in Fig. 5-15B, and tubes like the 6SA7, 7Q7 or 6BE6 are commonly used. The oscillator voltage is introduced into the electron stream of the tube through an "injection" grid. Measurement of the rectified current flowing in $R_2$ is used as a check for proper oscillator-voltage amplitude. Tuning of the signal-grid circuit can have little effect on the oscillator frequency because the injection grid is isolated from the signal grid by a screen grid that is at r.f. ground potential. The pentagrid mixer is not quite as sensitive as a triode or pentode mixer, but its splendid isolating characteristics make it a very useful circuit.

Many receivers use pentagrid converters, and two typical circuits are shown in Fig. 5-16. The circuit shown in Fig. 5-16A, which is suitable for the 6K8, 7D7, 7J7 or 7S7, is for a "triode-hexode" converter. A triode oscillator tube is mounted in the same envelope with a hexode, and the control grid of the oscillator portion is connected internally to an injection grid in the hexode. The isolation between oscillator and converter tube is reasonably good, and very little pulling results, except on signal frequencies that are quite large compared with the i.f.

The pentagrid-converter circuit shown in Fig. 5-16B can be used with a tube like the 6SA7, 7Q7 or 6BE6. Generally the only care necessary is to adjust the feed-back of the oscillator circuit to give the proper oscillator r.f. voltage. This condition is checked by measuring the d.c. current flowing in grid resistor $R_2$.

A more stable receiver generally results, particularly at the higher frequencies, when separate tubes are used for the mixer and oscillator. Practically the same number of circuit components is required whether or not a combination tube is used, so that there is little difference to be realized from the cost standpoint.

Typical circuit constants for converter tubes are given in Table 5-I. The grid leak referred to is the oscillator grid leak or injection-grid return, $R_2$ of Figs. 5-15 and 5-16.



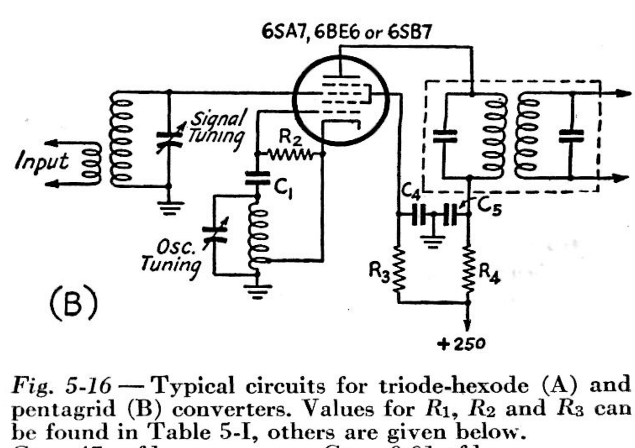




*Fig. 5-16* — Typical circuits for triode-hexode (A) and pentagrid (B) converters. Values for $R_1$, $R_2$ and $R_3$ can be found in Table 5-I, others are given below.

| | |
|---|---|
| $C_1$ — 47 $\mu\mu$fd. | $C_3$ — 0.01 $\mu$fd. |
| $C_2$, $C_4$, $C_5$ — 0.001 $\mu$fd. | $R_4$ — 1000 ohms. |

## THE HIGH-FREQUENCY OSCILLATOR

Stability of the receiver is dependent chiefly upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to changes in voltage, loading, and mechanical shock. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. These ends can be attained by the use of high-grade insulating materials and circuit components, suitable electrical design, and careful mechanical construction.

In addition, the oscillator must be capable of furnishing sufficient r.f. voltage and power for the particular mixer circuit chosen, at all frequencies within the range of the receiver, and its harmonic output should be as low as possible to reduce the possibility of spurious response.

It is desirable to make the $L/C$ ratio in the oscillator tuned circuit low (high-$C$), since this results in increased stability. Particular care should be taken to insure that no part of the oscillator circuit can vibrate mechanically. This calls for short leads and "solid" mechanical construction. The chassis and panel materials should be heavy and rigid enough so that pressure on the tuning dial will not cause torsion and a shift in the frequency. Care in mechanical construction of a receiver is repaid many times over by increased frequency stability.

### *Circuits*

Several oscillator circuits are shown in Fig. 5-17. The point at which output voltage is taken for the mixer is indicated in each case by $X$ or $Y$. Circuits A and B will give about the same results, and require only one coil. However, in these two circuits the cathode is above ground potential for r.f., which often is a cause of hum modulation of the oscillator output at 14 Mc. and higher frequencies when indirectly-heated-cathode tubes with a.c. on the heaters are used. The circuit of Fig. 5-17C reduces hum because the cathode is grounded. It is a simple circuit to adjust, and it is also the best circuit to use with filament-type tubes. With filament-type tubes, the other two circuits would require r.f. chokes to keep the filament above r.f. ground.

Besides the use of a fairly high $C/L$ ratio in the tuned circuit, it is necessary to adjust the feed-back to obtain optimum results. Too much feed-back will cause the oscillator to "squeg," or operate at several frequencies simultaneously; too little feed-back will cause the output to be low. In the tapped-coil circuits (A, B), the feed-back is increased by moving the tap toward the grid end of the coil; in $C$, by increasing the number of turns on $L_2$ or by moving $L_2$ closer to $L_1$.

The oscillator plate voltage should be as low as is consistent with adequate output. Low plate voltage will cause reduced tube heating and thereby reduce frequency drift. The oscillator and mixer circuits should be well isolated, preferably by shielding, since coupling other than by the means intended may result in pulling.

If the h.f.-oscillator frequency is affected by changes in plate voltage, it is good practice to use a voltage-regulated plate supply employing a VR tube except, of course, in receivers operated from batteries. Changes in plate-supply voltage are caused not only by variations in the line voltage but by poor regulation in the power supply. When a.v.c. is used, the controlled tubes draw less current from the power supply as the signal increases, and this change in power-supply load causes the power-supply voltage to vary if it isn't regulated. The use of Class AB audio amplification may also cause severe changes in the power-supply voltage.

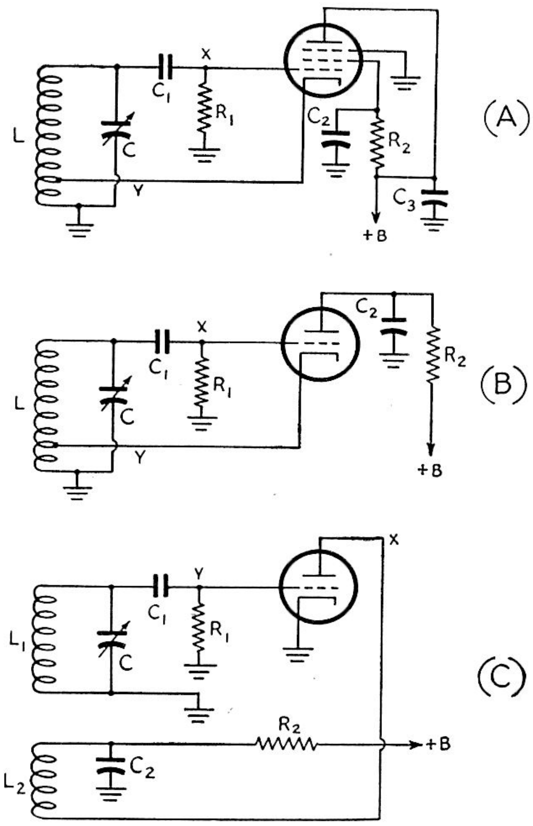



*Fig. 5-17* — High-frequency oscillator circuits. A, pentode grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points $X$ and $Y$. In A and B, coupling from $Y$ will reduce pulling effects, but gives less voltage than from $X$; this type is best adapted to mixer circuits with small oscillator-voltage requirements. Typical values for components are as follows:

| | *Circuit A* | *Circuit B* | *Circuit C* |
|---|---|---|---|
| $C_1$ — | 100 μμfd. | 100 μμfd. | 100 μμfd. |
| $C_2$ — | 0.1 μfd. | 0.1 μfd. | 0.1 μfd. |
| $C_3$ — | 0.1 μfd. | | |
| $R_1$ — | 47,000 ohms. | 47,000 ohms. | 47,000 ohms. |
| $R_2$ — | 47,000 ohms. | 10,000 to 25,000 ohms. | 10,000 to 25,000 ohms. |

The plate-supply voltage should be 250 volts. In circuits B and C, $R_2$ is used to drop the supply voltage to 100–150 volts; it may be omitted if voltage is obtained from a voltage divider in the power supply.

# A Two-Tube Superheterodyne Receiver

By using multipurpose tubes, it is possible to build a superheterodyne receiver using only two tubes. One tube is used as a pentagrid converter, and the other tube (a dual triode) acts as an autodyne detector working at the intermediate frequency, and as a stage of audio amplification.

A receiver of this type is shown in Figs. 5-18 through 5-22. The circuit diagram is given in Fig. 5-19. A 6K8 is used to convert the frequency of the incoming signal to the fixed or intermediate frequency, and the two triode sections of a 6SN7 serve as the regenerative detector and audio amplifier respectively. $L_1C_1$ is the r.f. circuit, tuned to the signal, and $L_2$ is the antenna coupling coil. $C_7$ is a by-pass condenser across the 1.5-volt battery used to bias the signal grid of the 6K8 and the half of the 6SN7 used as an audio amplifier. The high-frequency oscillator tank circuit is $L_3C_3C_4$, with $C_3$ for band-setting and $C_4$ for bandspread.

The i.f. tuned circuit (or regenerative detector circuit) is $L_5C_5$. This must be a high-$C$ circuit if stability better than that of an ordinary regenerative detector is to be secured. The frequency to which it is tuned should be in the vicinity of 1600 kc. $L_5$ and its tickler coil, $L_6$, are wound on a small form, and $L_5$ is tuned by a fixed mica condenser of the low-drift type. Since these condensers are rated with a capacity tolerance of 5 per cent, it is sufficient to wind $L_5$ as specified under Fig. 5-19. The resulting resonant frequency will be in the correct region. No manual tuning is necessary, and therefore the frequency of this circuit need not be adjusted. $C_2$ is the regeneration-control condenser, isolated from the d.c. supply by the choke, *RFC*. Only enough turns need be used on $L_6$ to make the detector oscillate readily when $C_2$ is at half capacity or more.

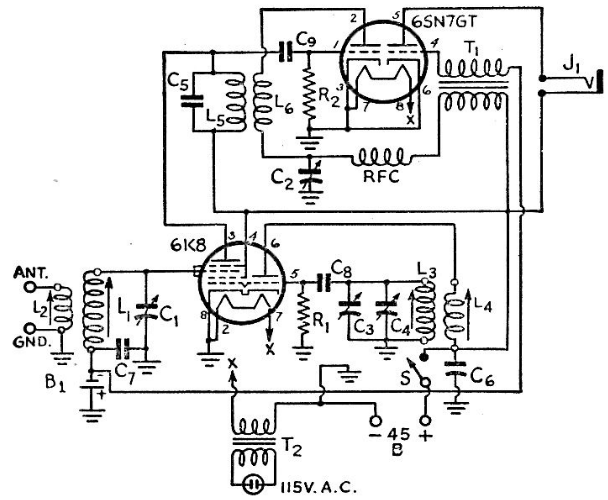



*Fig. 5-19* — **Circuit diagram of the two-tube superheterodyne receiver.**

$C_1$, $C_2$, $C_3$ — 100-μμfd. variable (Millen 19100).
$C_4$ — 15-μμfd. variable (Millen 20015).
$C_5$ — 240-μμfd. silvered mica.
$C_6$ — 0.01-μfd. paper.
$C_7$ — 0.0047-μfd. mica.
$C_8$, $C_9$ — 100-μμfd. mica.
$R_1$ — 47,000 ohms, ½ watt.
$R_2$ — 1 megohm, ½ watt.
$L_1$, $L_2$, $L_3$, $L_4$ — See coil table.
$L_5$ — 55 turns No. 30 d.s.c., close-wound on ¾-inch diam. form (National PRF-2); inductance 40 uh.
$L_6$ — 18 turns No. 30 d.s.c., close-wound on same form as $L_5$; see Fig. 5-20.
$B_1$ — 1.5-volt bias battery.
$J_1$ — Open-circuit jack.
RFC — 2.5-mh. r.f. choke.
S — S.p.s.t. toggle switch.
$T_1$ — Interstage audio transformer (Stancor A-4205).
$T_2$ — 6.3-volt filament transformer.

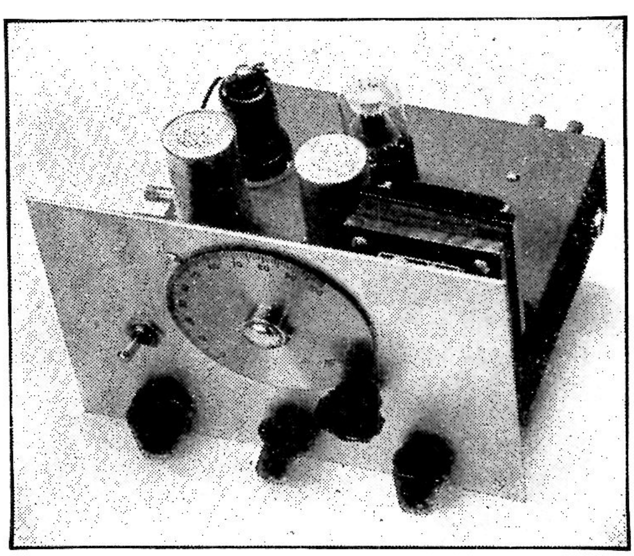


*Fig. 5-18* — **Panel view of the two-tube superheterodyne receiver. The panel is cut from a sheet of ⅛-inch aluminum. It is 6 inches high and 8 inches wide. The controls along the bottom, from left to right, are mixer tuning, oscillator padder and i.f. regeneration. The "B" switch is to the left of the tuning dial.**

The second section of the 6SN7 is transformer-coupled to the detector, and battery bias is used. The headphone output is taken from $J_1$, in the plate circuit.

Looking at the top of the chassis from in front, the r.f. or input circuit is at the left, with $C_1$ below the chassis and $L_1L_2$ just behind it. The 6K8 is directly to the rear of the coil. The h.f.-oscillator padding condenser, $C_3$, underneath, the socket for $L_3L_4$ and the 6SN7 are in line at the center of the chassis. At the right, underneath the audio transformer, $T_1$, is the i.f. regeneration-control condenser, $C_2$. The bandspread tuning condenser, $C_4$, is mounted on the panel with its shaft 3⅞ inches from the bottom edge of the panel. The audio transformer should be set back far enough so that there will be sufficient space for the bearing for the vernier knob of the National Type G dial. The "B" switch, $S$, is to the left of the dial.

A pair of terminals set in the left-hand edge

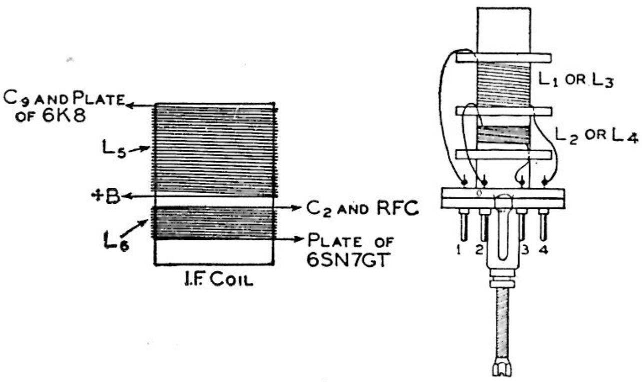



*Fig. 5-20* — How the coils for the two-tube superheterodyne receiver are wound. In both cases both windings are in the same direction. In the case of the i.f. coil at the left, the top end of the upper winding, $L_5$, is connected to $C_9$ and Pin 3 of the 6K8 socket, the lower end of $L_5$ is connected to Pin 4 of the 6K8, the upper end of the lower winding, $L_6$, is connected to the stator of $C_2$ and the lower end of $L_6$ goes to Pin 2 on the 6SN7GT socket.

In the case of the plug-in coils, the coil sockets and plug-in form bases are wired so that the upper end of $L_3$ connects to the stator of $C_3$, the lower end of this winding to the chassis, the upper end of the lower winding, $L_4$, to $C_6$ and the lower end of $L_4$ goes to Pin 6 on the 6K8 socket. When the coil is plugged into the mixer stage, the upper end of the top winding should go to the stator of $C_1$, the lower end to $C_7$ and the biasing battery, the upper end of the lower winding to the chassis and the lower end of the bottom winding to the antenna terminal.

of the chassis provide connections for antenna and ground, while another pair at the rear are for the "B"-battery connections. The antenna and B+ terminals must be insulated from the chassis. A jack in the right-hand side is provided for headphones, and 115 volts a.c. for the heater transformer, $T_2$, is plugged in at the rear. The headphone jack is insulated from the chassis by means of fiber washers. $T_2$ is placed under the chassis near the headphone jack.

Referring to the bottom view of Fig. 5-22, the biasing battery is to the left below $C_1$. It is a pen-light flashlight cell soldered between the coil-socket terminal and ground. Immediately below it is the by-pass condenser, $C_7$. $C_6$ is soldered between the socket terminal for $L_4$ and ground. The r.f. choke is supported at one end by a small fiber lug strip and soldered to $C_2$ at the other. The i.f. transformer, $L_5L_6$, is between the two tube sockets. $L_5$ is connected between the proper tube-socket terminals and $C_5$ is soldered across these same terminals. $C_9$ is fastened directly between the two tube sockets and $C_8$ between the 6K8 socket and the proper terminal of the socket for $L_3$. Clearance holes are drilled in the chassis for wiring to the switch, to the stator terminal of $C_4$, and to the grid cap of the 6K8. The rotor terminal of $C_4$ is grounded to the panel by a lug fastened under one of the mounting pillars. Two holes also are provided for the leads to $T_1$.

Coils for the receiver are wound on Millen shielded ½-inch diameter forms, Type 74001, which are provided with slug-type inductance trimmers.

The method of winding is indicated in Fig. 5-20; if the connections to the circuit are made as shown, there will be no trouble in obtaining the necessary oscillation. Both coils on each form should be wound in the same direction.

### *Adjustment*

To test the receiver, first try out the i.f. circuit. Connect the filament and "B" supplies and place both tubes in their sockets. Put a high-frequency coil in the r.f. socket, but do not insert a coil in the oscillator socket. The only test which needs be made is to see if the detector oscillates properly. Advance $C_2$ from minimum capacity until the detector goes into oscillation, which will be indicated by a soft hiss. This should occur at around half scale on the condenser. If it does not occur, check the coil ($L_5L_6$) connections and winding direction and, if these seem right, add a few turns to the tickler, $L_6$. If the detector oscillates with very low capacity at $C_2$, it will be advisable to take a few turns off $L_6$ until oscillation starts at about midscale.

**TWO-TUBE SUPERHET COIL DATA**

| | $L_1$ or $L_3$ | $L_2$ or $L_4$ |
|---|---|---|
| A. | 90 turns No. 30 d.s.c., closewound | 20 turns No. 30 d.s.c. |
| B. | 65 turns No. 26 d.s.c., closewound | 15 turns No. 26 d.s.c. |
| C. | 45 turns No. 22 d.s.c., closewound | 15 turns No. 26 d.s.c. |
| D. | 24 turns No. 22 enam., 1⅛ in. long | 15 turns No. 26 d.s.c. |
| E. | 20 turns No. 22 enam., 1⅛ in. long | 15 turns No. 26 d.s.c. |

| *Frequency Range* | *Coil at $L_1$-$L_2$* | *Coil at $L_3$-$L_4$* |
|---|---|---|
| 1700 to 3200 kc. | A | B |
| 3000 to 5700 kc. | B | C |
| 5400 to 10,000 kc. | C | D |
| 9500 to 14,500 kc. | E | D |

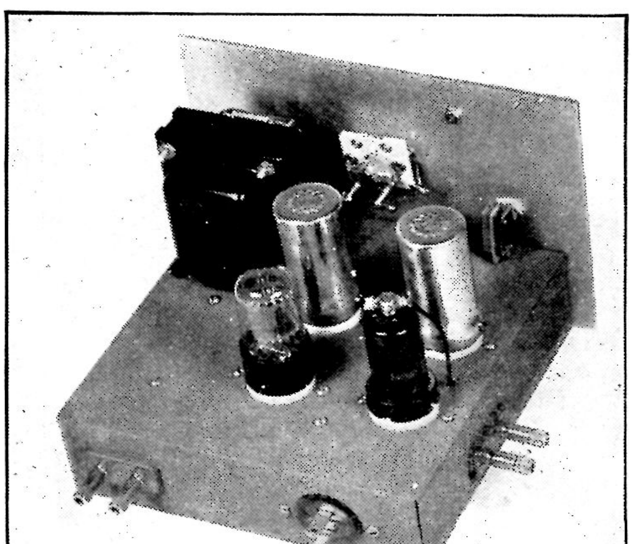


*Fig. 5-21* — A back-of-panel view of the two-tube superheterodyne receiver. The chassis is 7 × 7 × 2 inches.

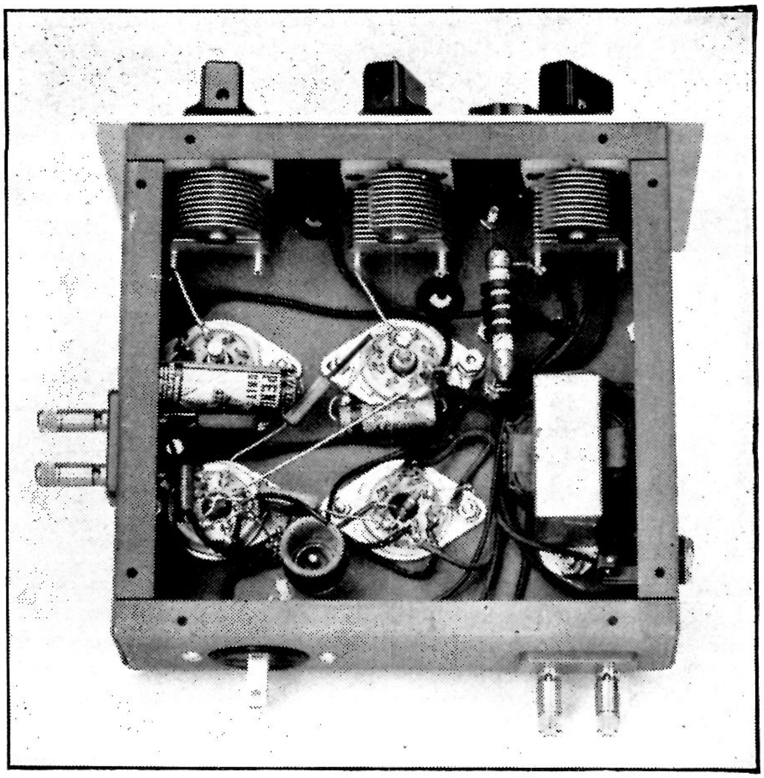


*Fig. 5-22* — **Bottom view of the two-tube superheterodyne receiver. The i.f. coil is between the two tube sockets near the rear of the chassis. The transformer to the right is the filament transformer.**

After the i.f. has been checked, plug in an oscillator coil for a range on which signals are likely to be heard at the time. The 5400–10,000-kc. range is usually a good one. The coils are arranged so that a minimum number is needed, even though two are used at a time. With Coil C in the r.f. socket and D in the oscillator socket, set $C_1$ at about half scale and turn $C_3$ slowly around midscale until a signal is heard. Then tune $C_1$ for maximum volume. Should no signals be heard, the probability is that the oscillator section of the 6K8 converter tube is not working, in which case the same method of testing is used as described above for the i.f. detector — check wiring, direction of windings of coils, and finally, add turns to the tickler, $L_4$, if necessary.

The same oscillator coil, D, is used for two frequency ranges. This is possible because the oscillator frequency is placed on the low-frequency side of the signal on the higher range. This gives somewhat greater stability at the highest frequency range. Some pulling — a change in beat-note as the r.f. tuning is varied by means of $C_1$ — will be observed on the highest frequency range, but it is not serious in the region of resonance with the incoming signal frequency.

The receiver will respond to signals either 1600 kc. lower or 1600 kc. higher than the oscillator frequency. The unwanted response is discriminated against by the selectivity of the r.f. circuit. On the three lower frequency ranges, when it is possible to find two tuning spots on $C_1$ at which incoming noise peaks up, the lower-frequency peak is the right one. The oscillator frequency is 1600 kc. higher than that of the incoming signal on these three ranges and 1600 kc. lower on the fourth range. The inductance of the coils to hit the desired ranges can be adjusted by means of the trimming slug in the coil forms.

The regeneration control may be set to give desired sensitivity and left alone while tuning; only when an exceptionally strong signal is encountered is it necessary to advance it more to keep the detector in oscillation. It should be set just on the edge of oscillation for 'phone reception.

The "B"-battery current is between 4 and 5 ma., so a standard 45-volt block will last hundreds of hours.

# The Intermediate-Frequency Amplifier

While the simple receiver just described is capable of highly-satisfactory results, it does not take full advantage of the superheterodyne principle. One major advantage of the superhet is that high gain and selectivity can be obtained by using a good i.f. amplifier. This can be a one-stage affair in simple receivers, or two or three stages in the more complex sets.

### *Choice of Frequency*

The selection of an intermediate frequency is a compromise between various conflicting factors. The lower the i.f. the higher the selectivity and gain, but a low i.f. brings the image nearer the desired signal and hence decreases the image ratio. A low i.f. also increases pulling of the oscillator frequency. On the other hand, a high i.f. is beneficial to both image ratio and pulling, but the selectivity and gain are lowered. The difference in gain is least important.

An i.f. of the order of 455 kc. gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 Mc. The image ratio is poor at 14 Mc. when the mixer is connected to the antenna, but adequate when there is a tuned r.f. amplifier between antenna and mixer. At 28 Mc. and on the very-high frequencies, the image ratio is very poor unless several r.f. stages are used. Above 14 Mc., pulling is likely to be bad unless very loose coupling can be used between mixer and oscillator.

With an i.f. of about 1600 kc., satisfactory image ratios can be secured on 14, 28 and 50 Mc., and pulling can be reduced to negligible proportions. However, the i.f. selectivity is considerably lower, so that more tuned circuits must be used to increase the selectivity. For frequencies of 28 Mc. and higher, the best solution is to use a double superheterodyne, choosing one high i.f. for image reduction (5 and 10 Mc. are frequently used) and a lower one for gain and selectivity.

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. The frequencies mentioned are fairly free of such interference.

### *Fidelity; Sideband Cutting*

Modulation of a carrier causes the generation of sideband frequencies numerically equal to the carrier frequency plus and minus the highest modulation frequency present. If the receiver is to give a faithful reproduction of modulation that contains, for instance, audio frequencies up to 5000 cycles, it must be capable of amplifying equally all frequencies contained in a band extending from 5000 cycles above to 5000 cycles below the carrier frequency. In a superheterodyne, where all carrier frequencies are changed to the fixed intermediate frequency, this means that the i.f. amplifier should amplify equally well all frequencies within that band. In other words, the amplification must be uniform over a band 10 kc. wide, with the i.f. at its center. The signal-frequency circuits usually do not have enough over-all selectivity to affect materially the "adjacent-channel" selectivity, so that only the i.f.-amplifier selectivity need be considered.

A 10-kc. band is considered sufficient for reasonably-faithful reproduction of music, but much narrower bandwidths can be used for communication work where intelligibility rather than fidelity is the primary objective. If the selectivity is too great to permit uniform amplification over the band of frequencies occupied by the modulated signal, the higher modulating frequencies are attenuated as compared to the lower frequencies; that is, the upper-frequency sidebands are "cut." While sideband cutting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of greater selectivity.

The selectivity of an i.f. amplifier, and hence the tendency to cut sidebands, increases with the number of amplifier stages and also is greater the lower the intermediate frequency. From the standpoint of communication, sideband cutting is not serious with two-stage amplifiers at frequencies as low as 455 kc.

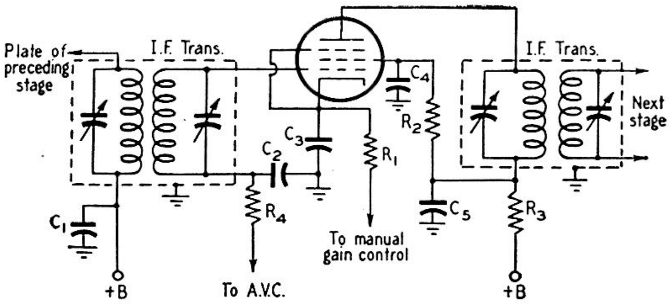



*Fig. 5-23* — **Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:**

**$C_1$ — 0.1 μfd. at 455 kc.; 0.01 μfd. at 1600 kc. and higher.**
**$C_2$ — 0.01 μfd.**
**$C_3$, $C_4$, $C_5$ — 0.1 μfd. at 455 kc.; 0.01 μfd. above 1600 kc.**
**$R_1$, $R_2$ — See Table 5-II. $R_3$ — 1800 ohms.**
**$R_4$ — 0.27 megohm.**

### *Circuits*

I.f. amplifiers usually consist of one or two stages. At 455 kc. two stages generally give all the gain usable, and also give suitable selectivity for good-quality 'phone reception.

A typical circuit arrangement is shown in Fig. 5-23. A second stage would simply duplicate the circuit of the first. The i.f. amplifier practically always uses a remote cut-off pentode-type tube operated as a Class A amplifier. For maximum selectivity, double-tuned transformers are used for interstage coupling, although single-tuned circuits or transformers with untuned primaries can be used for coupling, with a consequent loss in selectivity. All other things being equal, the selectivity of an i.f. amplifier is proportional to the number of tuned circuits in it. The use of too many high-$Q$ tuned circuits in an amplifier is not generally feasible, however, because of stability problems.

In Fig. 5-23, the gain of the stage is reduced by introducing a negative voltage to the lead marked "to a.v.c." or a positive voltage to $R_1$ at the point marked "to manual gain control." In either case, the voltage increases the bias on the tube and reduces the mutual conductance and hence the gain. When two or more stages are used, these voltages are generally obtained from common sources. The decoupling resistor, $R_3$, helps to isolate the amplifier from the power supply and thus prevents stray feed-back. $C_2$ and $R_4$ are part of the automatic volume-control circuit (described later); if no a.v.c. is used, the lower end of the i.f.-transformer secondary is simply connected to ground.

In a two-stage amplifier the screen grids of both stages may be fed from a common supply, either through a resistor ($R_2$) as shown, the screens being connected in parallel, or from a voltage divider across the plate supply. Separate screen voltage-dropping resistors are preferable for preventing undesired coupling between stages.

Typical values of cathode and screen resistors for common tubes are given in Table 5-II. The 6K7, 6SK7, 6SG7, 6BA6 and 7H7 are recommended for i.f. work.

When two stages are used the high gain will tend to cause instability and oscillation, so that good shielding, by-passing, and careful

**TABLE 5-II**
**Cathode and Screen-Dropping Resistors for R.F. or I.F. Amplifiers**

| *Tube* | *Plate Volts* | *Screen Volts* | *Cathode Resistor* | *Screen Resistor* |
|---|---|---|---|---|
| 6AB7 | 300 | | 200 ohms | 33,000 ohms |
| 6AC7 | 300 | | 160 | 62,000 |
| 6AK5 | 180 | 120 | 200 | 27,000 |
| 6AU6 | 250 | 150 | 68 | 33,000 |
| 6BA6 | 250 | 100 | 68 | 33,000 |
| 6J7 | 250 | 100 | 1200 | 270,000 |
| 6K7 | 250 | 125 | 240 | 47,000 |
| 6SG7 | 250 | 125 | 68 | 27,000 |
| 6SG7 | 250 | 150 | 200 | 47,000 |
| 6SH7 | 250 | 150 | 68 | 39,000 |
| 6SJ7 | 250 | 100 | 820 | 180,000 |
| 6SK7 | 250 | 100 | 270 | 56,000 |
| 7G7/1232 | 250 | 100 | 270 | 68,000 |
| 7H7 | 250 | 150 | 180 | 27,000 |

circuit arrangement to prevent stray coupling, with exposed r.f. leads well separated, are necessary.

### *I.F. Transformers*

The tuned circuits of i.f. amplifiers are built up as transformer units consisting of a metal-shield container in which the coils and tuning condensers are mounted. Both air-core and powdered-iron-core universal-wound coils are used, the latter having somewhat higher $Q$s and, hence, greater selectivity and gain per unit. In universal windings the coil is wound in layers with each turn traversing the length of the coil, back and forth, rather than being wound perpendicular to the axis as in ordinary single-layer coils. In a straight multilayer winding, the turns on adjacent layers at the edges of the coil have a rather large potential difference between them as compared to the difference between any two adjacent turns in the same layer; hence a fairly large capacity can exist between layers. Universal winding, with its "criss-crossed" turns, tends to avoid building up such potential differences, and hence reduces distributed-capacity effects.

Variable tuning condensers are of the midget type, air-dielectric condensers being preferable because their capacity is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the inductance (permeability tuning), in which case stability comparable to that of variable air-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier. Typical i.f.-transformer construction is shown in Fig. 5-24.

Besides the type of i.f. transformer shown in Fig. 5-24, special units to give desired selectivity characteristics are available. For higher-than-ordinary adjacent-channel selectivity **triple-tuned** transformers, with a third tuned circuit inserted between the input and output windings, are used. The energy is transferred from the input to the output windings via this **tertiary winding,** thus adding its selectivity to the over-all selectivity of the transformer. **Variable-selectivity** transformers also can be obtained. These usually are provided with a third (untuned) winding which can be connected to a resistor, thereby loading the tuned circuits and decreasing the $Q$ and selectivity to broaden the selectivity curve. The variation in selectivity is brought about by switching the resistor in and out of the circuit. Another method is to vary the coupling between primary and secondary, overcoupling being used to broaden the selectivity curve and undercoupling to sharpen it.

### *Selectivity*

The over-all selectivity of the i.f. amplifier will depend on the frequency and the number of stages. The following figures are indicative of the bandwidths to be expected with good-quality transformers in amplifiers so constructed as to keep regeneration at a minimum:

| | *Bandwidth in Kilocycles* | | |
|---|---|---|---|
| *Intermediate Frequency* | *2 times down* | *10 times down* | *100 times down* |
| One stage, 455 kc. (air core) | 8.7 | 17.8 | 32.3 |
| One stage, 455 kc. (iron core) | 4.3 | 10.3 | 20.4 |
| Two stages, 455 kc. (iron core) | 2.9 | 6.4 | 10.8 |
| Two stages, 1600 kc. | 11.0 | 16.6 | 27.4 |
| Two stages, 5000 kc. | 25.8 | 46.0 | 100.0 |

### *Tubes for I.F. Amplifiers*

Variable-$\mu$ (remote cut-off) pentodes are almost invariably used in i.f. amplifier stages, since grid-bias gain control is practically always applied to the i.f. amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest

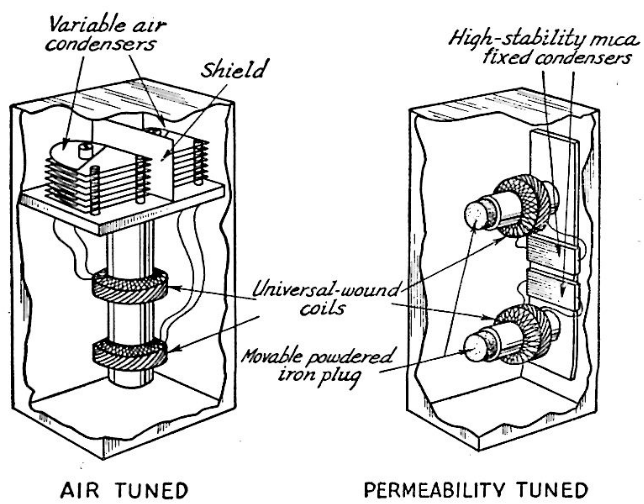



*Fig. 5-24* — Representative i.f.-transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the cores consist of finely-divided iron particles supported in an insulating binder, formed into cylindrical "plugs." The tuning capacity is fixed, and the inductances of the coils are varied by moving the iron plugs in and out,

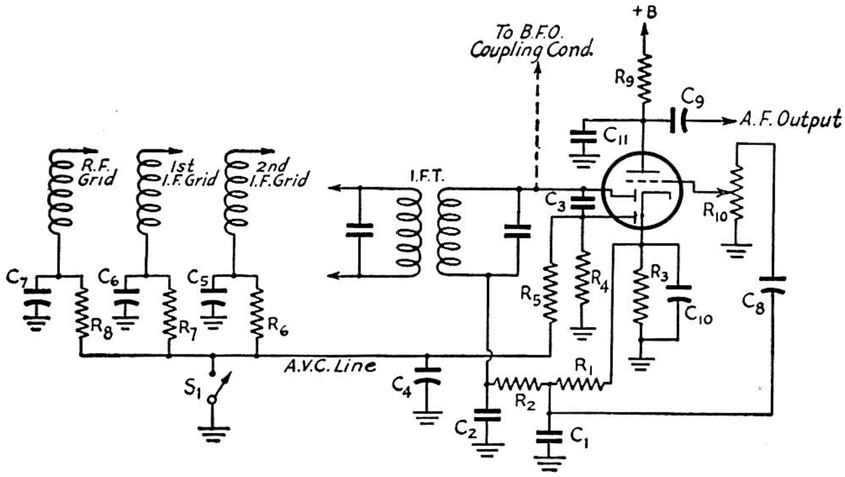



*Fig. 5-25* — **Automatic volume-control circuit using a dual-diode-triode as a combined a.v.c. rectifier, second detector and first audio-frequency amplifier.**

$R_1$ — 0.27 megohm.
$R_2$ — 50,000 to 250,000 ohms.
$R_3$ — 1800 ohms.
$R_4$ — 2 to 5 megohms.
$R_5$ — 0.5 to 1 megohm.
$R_6$, $R_7$, $R_8$, $R_9$ — 0.25 megohm.
$R_{10}$ — 0.5-megohm variable.
$C_1$, $C_2$, $C_3$ — 100 $\mu\mu$fd.
$C_4$ — 0.1 $\mu$fd.
$C_5$, $C_6$, $C_7$ — 0.01 $\mu$fd.
$C_8$, $C_9$ — 0.01 to 0.1 $\mu$fd.
$C_{10}$ — 5- to 10-$\mu$fd. electrolytic. $C_{11}$ — 270 $\mu\mu$fd.

gain. The choice of i.f. tubes has practically no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and r.f. amplifier (if the latter is used).

When single-ended tubes are used, care should be taken to keep the plate and grid leads well separated. With these tubes it is advisable to mount the screen by-pass condenser directly on the bottom of the socket, crosswise between the plate and grid pins, to provide additional shielding. The outside foil of the condenser should be connected to ground.

## THE SECOND DETECTOR AND BEAT OSCILLATOR

### *Detector Circuits*

The second detector of a superheterodyne receiver with an i.f. amplifier performs the same function as the detector in the simple receiver, but usually operates at a higher input level because of the relatively great amplification ahead of it. Therefore, the ability to handle large signals without distortion is preferable to high sensitivity. Plate detection is used to some extent, but the diode detector is most popular. It is especially adapted to furnishing automatic gain or volume control. The basic circuits have been described, although in many cases the diode elements are incorporated in a multipurpose tube that contains an amplifier section in addition to the diode.

### *The Beat Oscillator*

Any standard oscillator circuit may be used for the beat oscillator required for heterodyne reception. Special beat-oscillator transformers are available, usually consisting of a tapped coil with adjustable tuning; these are most conveniently used with circuits such as those shown at Fig. 5-17A and B, with the output taken from *Y*. A variable condenser of about 25-$\mu\mu$fd. capacity may be connected between cathode and ground to provide fine adjustment. The beat oscillator usually is coupled to the second-detector tuned circuit through a fixed condenser of a few $\mu\mu$fd. capacity.

The beat oscillator should be well shielded, to prevent coupling to any part of the circuit except the second detector and to prevent its harmonics from getting into the front end of the receiver and being amplified with desired signals. To this end, the plate voltage should be as low as is consistent with sufficient audio-frequency output. If the beat-oscillator output is too low, strong signals will not give a proportionately strong audio response.

When an oscillating second detector is used to give the audio beat-note, the detector must be detuned from the i.f. by an amount equal to the frequency of the beat note. The selectivity and signal strength will be reduced, while blocking will be pronounced because of the high signal level at the second detector.

## AUTOMATIC VOLUME CONTROL

### *Principles*

Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is a great advantage, especially in 'phone reception, since it tends to keep the output level of the receiver constant regardless of input-signal strength. It is readily accomplished in superheterodyne receivers by using the average rectified d.c. voltage, developed by the received signal across a resistance in a detector circuit, to vary the bias on the r.f. and i.f. amplifier tubes. Since this voltage is proportional to the average amplitude of the signal, the gain is reduced as the signal strength becomes greater. The control will be more complete as the number of stages to which the a.v.c. bias is applied is increased. Control of at least two stages is advisable.

### *Circuits*

A typical circuit using a diode-triode type tube as a combined a.v.c. rectifier, detector and first audio amplifier is shown in Fig. 5-25. One plate of the diode section of the tube is used for signal detection and the other for a.v.c. rectification. The a.v.c. diode plate

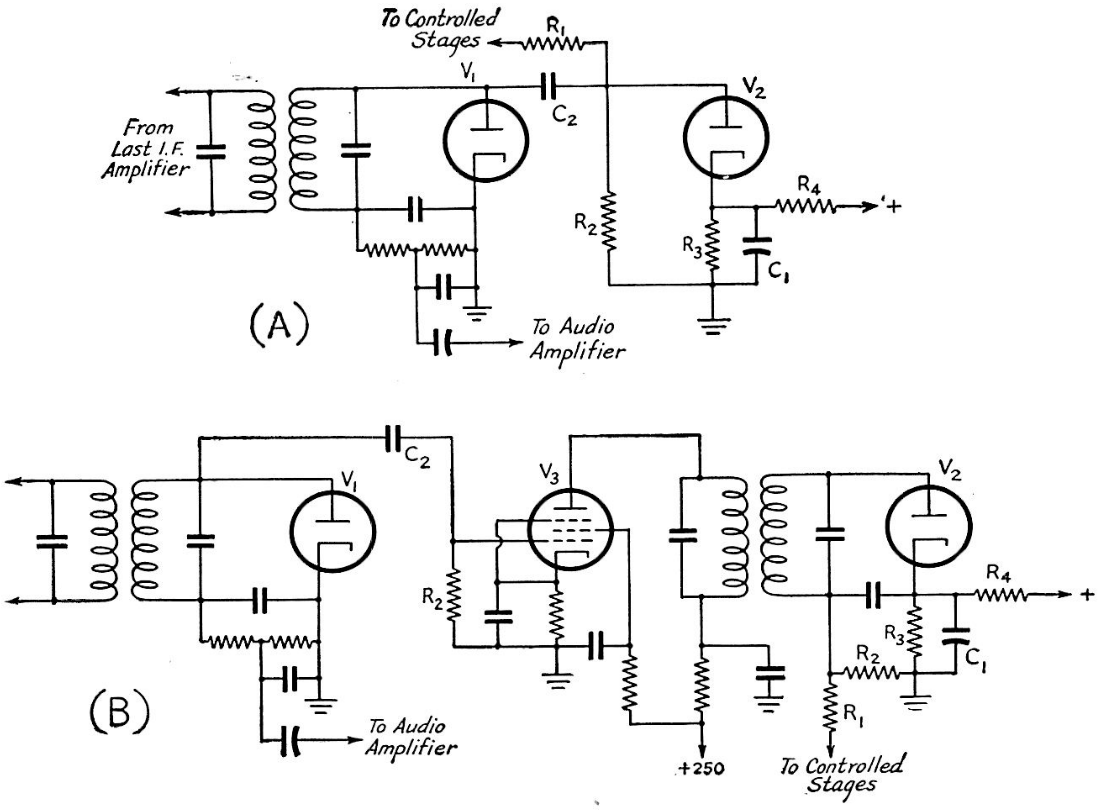



*Fig. 5-26* — Delayed a.v.c. is shown at A, and amplified and delayed a.v.c. is shown in B. The circuit at B gives excellent a.v.c. action over a wide range, with no impairment of sensitivity for weak signals. For either circuit, typical values are:

$C_1$ — 0.001 $\mu$fd. $R_1$, $R_2$ — 1.0 megohm.
$C_2$ — 100 $\mu\mu$fd. $R_3$, $R_4$ — Voltage divider.

Resistors $R_3$ and $R_4$ are carefully proportioned to give the desired delay voltage at the cathode of diode $V_2$. Bleeder current of 1 or 2 ma. is ample, and hence the bleeder can be figured on 1000 or 500 ohms per volt. The delay voltage should be in the vicinity of 3 or 4 for a simple receiver and 20 or 30 for a multitube high-gain affair.

is fed from the detector diode through the small coupling condenser, $C_3$. A negative bias voltage resulting from the flow of rectified carrier current is developed across $R_4$, the diode load resistor. This negative bias is applied to the grids of the controlled stages through the filtering resistors, $R_5$, $R_6$, $R_7$ and $R_8$. When $S_1$ is closed the a.v.c. line is grounded, thereby removing the a.v.c. bias from the amplifier without disturbing the detector circuit.

It does not matter which of the two diode plates is selected for audio and which for a.v.c. Frequently the two plates are connected together and used as a combined detector and a.v.c. rectifier. This could be done in Fig. 5-25. The a.v.c. filter and line would connect to the junction of $R_2$ and $C_2$, while $C_3$ and $R_4$ would be omitted from the circuit.

### Delayed A.V.C.

In Fig. 5-25 the audio-diode return is made directly to the cathode and the a.v.c. diode return to ground. This places negative bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since no rectification takes place in the a.v.c. diode circuit until the carrier amplitude is large enough to overcome the bias. Without this delay the a.v.c. would start working even with a very small signal. This is undesirable, because the full amplification of the receiver then could not be realized on weak signals. In the audio-diode circuit this fixed bias would cause distortion, and must be avoided; hence, the return is made directly to the cathode.

### Time Constant

The time constant of the resistor-condenser combinations in the a.v.c. circuit is an important part of the system. It must be high enough so that the modulation on the signal is completely filtered from the d.c. output, leaving only an average d.c. component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the a.v.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal, and in practice would cause frequency distortion. On the other hand, the time constant must not be too great or the a.v.c. will be unable to follow rapid fading. The capacitance and resistance values indicated in Fig. 5-25 will give a time constant that is satisfactory for high-frequency reception.

### C.W.

A.v.c. can be used for c.w. reception but the circuit is more complicated. The a.v.c. voltage must be derived from a rectifier that is isolated from the beat-frequency oscillator (otherwise the rectified b.f.o. voltage will reduce the receiver gain even with no signal coming through). This is generally done by using a separate a.v.c. channel connected to an i.f. amplifier stage ahead of the second detector (and b.f.o.). If the i.f. selectivity ahead of the a.v.c. rectifier isn't good, strong adjacent signals will develop a.v.c. voltages that will reduce the receiver gain while listening to weak signals. When clear channels are available, however, c.w. a.v.c. will hold the receiver output constant over a wide range of signal input.

### *Amplified A.V.C.*

The a.v.c. system shown in Fig. 5-25 will not hold the audio output of the receiver exactly constant, although the variation becomes less as more stages are controlled by the a.v.c. voltage. The variation also becomes less as the delay voltage is increased, although there will of course be variation in output if the signal intensity is below the delay-voltage level at the a.v.c. rectifier. In the circuit of Fig. 5-25, the delay voltage is set by the proper operating bias for the triode portion of the tube. However, a separate diode may be used, as shown in Fig. 5-26A. Since such a system requires a large voltage at the diode, a separate i.f. stage is sometimes used to feed the delayed a.v.c. diode, as in Fig. 5-26B. A system like this, sometimes called an amplified a.v.c. system, gives excellent control once the delay voltage is reached, and yet maintains full receiver sensitivity up to that point.

## Noise Reduction

### *Types of Noise*

In addition to tube and circuit noise, much of the noise interference experienced in reception of high-frequency signals is caused by domestic electrical equipment and by automobile ignition systems. The interference is of two types in its effects. The first is the "hiss" type, consisting of overlapping pulses similar in nature to the receiver noise. It is largely reduced by high selectivity in the receiver, especially for code reception. The second is the "pistol-shot" or "machine-gun" type, consisting of separated impulses of high amplitude. The "hiss" type of interference usually is caused by commutator sparking in d.c. and series-wound a.c. motors, while the "shot" type results from separated spark discharges (a.c. power leaks, switch and key clicks, ignition sparks, and the like).

### *Impulse Noise*

Impulse noise, because of the extremely short duration of the pulses as compared to the time between them, must have high pulse amplitude to contain much average energy. Hence, noise of this type strong enough to cause much interference generally has an instantaneous amplitude much higher than that of the signal being received. The general principle of devices intended to reduce such noise is that of allowing the signal amplitude to pass through the receiver unaffected, but making the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared to its time of duration the more successful the noise reduction, since more of the constituent energy can be suppressed.

In passing through selective receiver circuits, the time duration of the impulses is increased, because of the *Q* or flywheel effect of the circuits. Hence, the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good noise suppression.

### *Audio Limiting*

A considerable degree of noise reduction in code reception can be accomplished by amplitude-limiting arrangements applied to the audio output circuit of a receiver. Such limiters also maintain the signal output nearly constant without fading. These output-limiter systems are simple, and adaptable to most receivers. However, they cannot prevent noise peaks from overloading previous circuits.

## A Simple Audio Noise Limiter

The limiter shown in Fig. 5-27 is plugged into the receiver 'phone jack and the headphones are plugged into the limiter, so that no work on the receiver is required. The limiter will also keep the strength of c.w. signals at a constant comfortable level and will do much to relieve the operating fatigue resulting from long hours of listening to crackles, key clicks, blocking signals, and the like.

As can be seen from the wiring diagram in Fig. 5-28, two 1N34 crystal diodes, individually biased by 1½-volt dry cells, are used to short-circuit any signal coming through the 'phone circuit that has an amplitude greater than about 3 volts, peak-to-peak. Hence if the audio gain of the receiver is adjusted to give a signal of this amplitude — comfortable headphone volume — noise peaks of greater amplitude will be short-circuited and not heard in the headphones. A 6AL5 twin diode can be substituted for the two 1N34 crystals, but a heater supply will be required and it is generally more convenient to build the limiter as shown. No current is drawn from the two cells

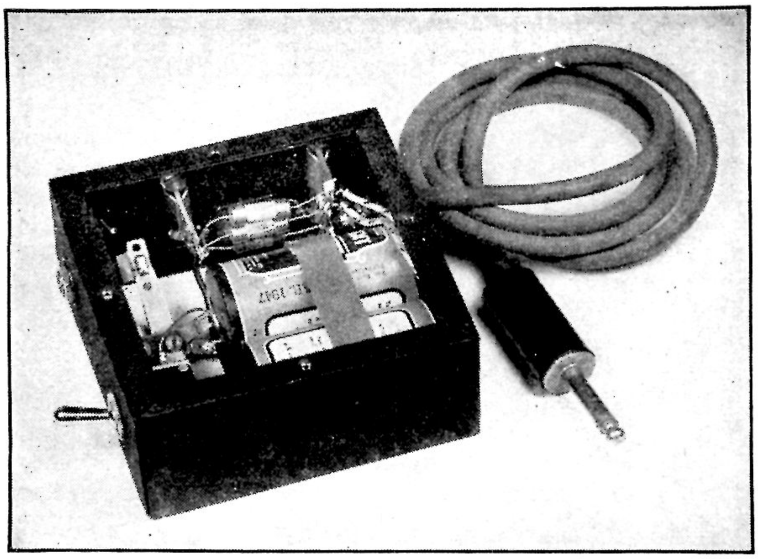


*Fig. 5-27* — A crystal-diode noise limiter for use between receiver and headphones. Built in a 4 × 4 × 2-inch box, it contains the limiter crystals, bias cells, headphone jack and on-off switch, and is provided with a cord and plug to connect to the receiver headphone output.

Although primarily intended for c.w. reception, the limiter also is highly effective on 'phone signals when the audio volume level is properly set and the r.f. gain is automatically controlled.

used for bias, and they will last their shelf life.

The limiter can be built in a 4 × 4 × 2-inch cabinet, as shown in Fig. 5-27. By removing the two sides of the cabinet, all of the components can be mounted in the frame. The two dry cells can be taped together and then held in place by heavy leads soldered to them, or special clips can be made of spring brass. The two 1N34 crystal diodes are best mounted on tie-points, and the pigtails of the diodes should be held in a pair of long-nose pliers while soldering to them, because too much heat from the soldering iron may decrease the effectiveness of the crystal. The pliers conduct away the heat that might otherwise reach the crystal.

## SECOND-DETECTOR NOISE-LIMITER CIRCUITS

The circuit of Fig. 5-29 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode, which becomes nonconducting above a predetermined signal level. The audio output of the detector must pass through the diode to the grid of the amplifier tube. The diode normally would be nonconducting with the connections shown were it not for the fact that it is given positive bias from a 30-volt source through the adjustable potentiometer, $R_3$. Resistors $R_1$ and $R_2$ must be fairly large in value to prevent loss of audio signal.

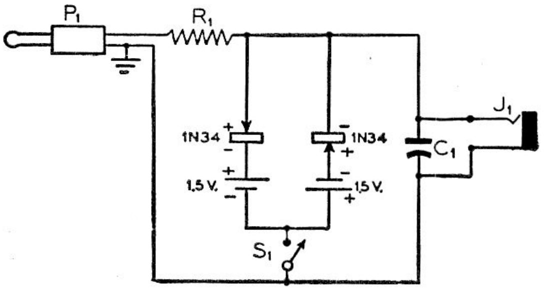



*Fig. 5-28* — **Practical noise-limiter circuit for headphone reception.**

**$C_1$ — 0.005 μfd.**
**$R_1$ — 15,000 ohms, ½ watt.**
**$J_1$ — Single-circuit jack.**
**$P_1$ — Headphone plug.**
**$S_1$ — S.p.s.t. toggle.**

The audio signal from the detector can be considered to modulate the steady diode current, and conduction will take place so long as the diode plate is positive with respect to the

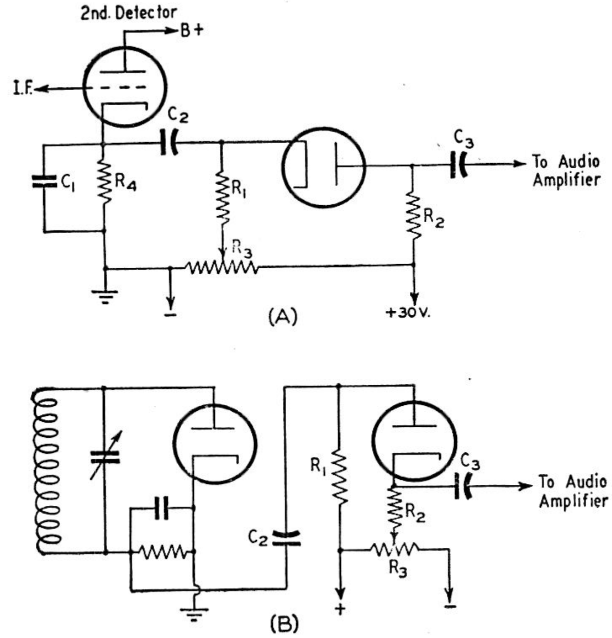



*Fig. 5-29* — **Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:**

**$R_1$ — 0.27 megohm.**
**$R_2$ — 47,000 ohms.**
**$R_3$ — 10,000 ohms.**
**$R_4$ — 20,000 to 50,000 ohms.**
**$C_1$ — 270 μμfd.**
**$C_2$, $C_3$ — 0.1 μfd.**

**All other diode-circuit constants in B are conventional.**

cathode. When the signal is sufficiently large to swing the cathode positive with respect to the plate, however, conduction ceases, and that portion of the signal is cut off from the audio amplifier. The point at which cut-off occurs can be selected by adjustment of $R_3$. By setting $R_3$ so that the signal just passes through the "valve," noise pulses higher in amplitude than the signal will be cut off. The circuit of Fig. 5-29A, using an infinite-impedance detector, gives a positive voltage on rectification. When the rectified voltage is negative, as it is from the usual diode detector,

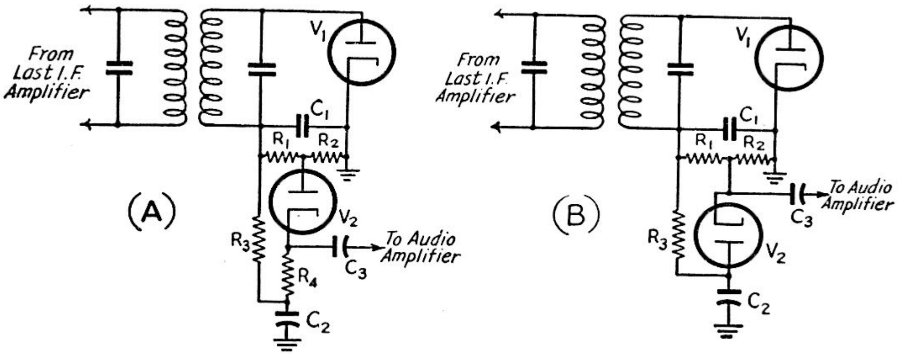



*Fig. 5-30* — **Self-adjusting series (A) and shunt (B) noise limiters. The functions of $V_1$ and $V_2$ can be combined in one tube like the 6H6 or 6AL5, or Type 1N34 crystals can be used.**
**$C_1$ — 100 $\mu\mu$fd.**
**$C_2$, $C_3$ — 0.05 $\mu$fd.**
**$R_1$ — 0.27 meg. in A; 47,000 ohms in B.**
**$R_2$ — 0.27 meg. in A; 0.15 meg. in B.**
**$R_3$ — 1.0 megohm.**
**$R_4$ — 0.82 megohm.**

the circuit arrangement shown in Fig. 5-29B must be used.

An audio signal of about ten volts is required for good limiting action. When a beat oscillator is used for c.w. reception the b.f.o. voltage should be small, so that incoming noise will not have a strong carrier to beat against and so produce large audio output.

Second-detector noise-limiting circuits that automatically adjust themselves to the receiver carrier level are shown in Fig. 5-30. In either circuit, $V_1$ is the usual diode second detector, $R_1R_2$ is the diode load resistor, and $C_1$ is an r.f. by-pass. A negative voltage proportional to the carrier level is developed across $C_2$, and this voltage cannot change rapidly because $R_3$ and $C_2$ are both large. In the circuit at A, diode $V_2$ acts as a conductor for the audio signal up to the point where its anode is negative with respect to the cathode. Noise peaks that exceed the maximum carrier-modulation level will drive the anode negative instantaneously, and during this time the diode does not conduct. The large time constant of $C_2R_3$ prevents any rapid change of this reference voltage. In the circuit at B, the diode $V_2$ is inactive until its cathode voltage exceeds its anode voltage. This condition will obtain under noise peaks and, when it does, the diode $V_2$ short-circuits the signal and no voltage is passed on to the audio amplifier. Practical values for these two circuits will be found in the six- and eight-tube superheterodynes described later in this chapter. Diode rectifiers such as the 6H6 and 6AL5, or the 1N34 germanium crystal diode, can be used for these types of noise limiters. Neither circuit is useful for c.w. reception, but they are both quite effective for 'phone work.

### *I.F. Noise Silencer*

In the circuit shown in Fig. 5-31, noise pulses are made to decrease the gain of an i.f. stage momentarily and thus silence the receiver for the duration of the pulse. Any noise voltage in excess of the desired signal's maximum i.f. voltage is taken off at the grid of the i.f. amplifier, amplified by the noise-amplifier stage, and rectified by the full-wave diode noise rectifier. The noise circuits are tuned to the i.f. The rectified noise voltage is applied as a pulse of negative bias to the No. 3 grid of the 6L7 i.f. amplifier, wholly or partially disabling this stage for the duration of the individual noise pulse, depending on the amplitude of the noise voltage. The noise-amplifier/rectifier circuit is biased by means of the "threshold control," $R_2$, so that rectification will not start until the noise voltage exceeds the desired signal amplitude. With automatic volume control the a.v.c. voltage can be applied to the grid of the noise amplifier, to augment this threshold bias. In a typical instance, this system improved the signal-to-noise ratio some 30 db. (power ratio of 1000) with heavy ignition interference, raising the signal-to-noise ratio from −10 db. without the silencer to +20 db. with the silencer.

## SIGNAL-STRENGTH AND TUNING INDICATORS

A useful accessory to the receiver is an indicator that will show relative signal strength. Not only is it an aid in giving reports to transmitting stations, but it is helpful also in aligning the receiver circuits, in conjunction with a test oscillator or other steady signal.

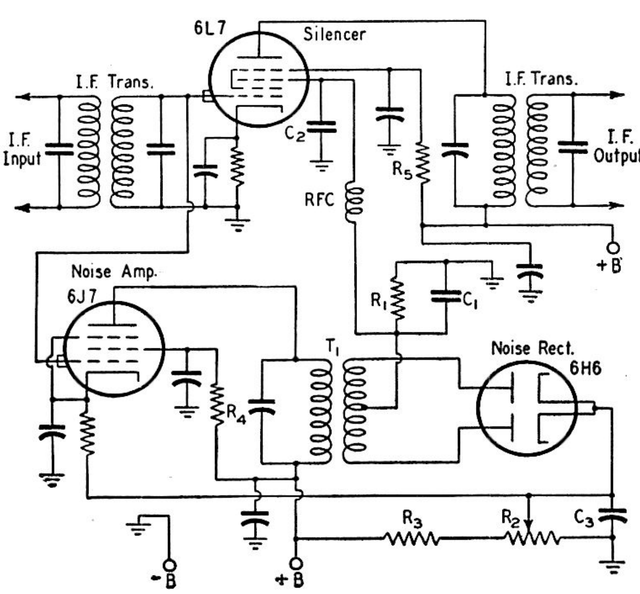



*Fig. 5-31* — **I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:**
**$C_1$ — 50–250 $\mu\mu$fd. (use smallest value possible without r.f. feed-back).**
**$C_2$ — 47 $\mu\mu$fd.** **$R_2$ — 5000-ohm variable.**
**$C_3$ — 0.1 $\mu$fd.** **$R_3$ — 22,000 ohms.**
**$R_1$ — 0.1 megohm.** **$R_4$, $R_5$ — 0.1 megohm.**
**$T_1$ — Special i.f. transformer for noise rectifier.**

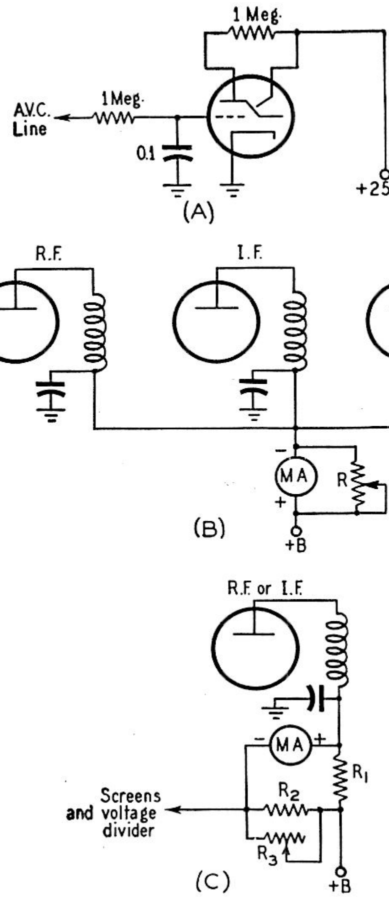



*Fig. 5-32* — **Tuning-indicator or S-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor *R* should have a maximum resistance several times that of the milliammeter. In C, representative values for the components are: $R_1$, 270 ohms; $R_2$, 330 ohms; $R_3$, 1000-ohm variable.**

Three types of indicators are shown in Fig. 5-32. That at A uses an electron-ray tube, several types of which are available. The grid of the triode section usually is connected to the a.v.c. line. The particular type of tube used depends upon the voltage available for its grid; where the a.v.c. voltage is large, a remote cut-off type (6G5, 6N5 or 6AD6G) should be used in preference to the more sensitive sharp cut-off type (6E5).

In B, a milliammeter is connected in series with the d.c. plate lead to one or more r.f. and i.f. tubes, the grids of which are controlled by a.v.c. voltage. Since the plate current of such tubes varies with the strength of the incoming signal, the meter will indicate relative signal intensity and may be calibrated in S-points. The scale range of the meter should be chosen to fit the number of tubes in use; the maximum plate current of the average remote cut-off r.f. pentode is from 7 to 10 milliamperes. The shunt resistor, *R*, enables setting the plate current to the full-scale value ("zero adjustment"). With this system the ordinary meter reads downward from full scale with increasing signal strength, which is the reverse of normal pointer movement (clockwise with increasing reading). Special instruments in which the zero-current position of the pointer is on the right-hand side of the scale are used in commercial receivers.

The system at C uses a 0–1 milliammeter in a bridge circuit, arranged so that the meter reading and the signal strength increase together. The current through the branch containing $R_1$ should be approximately equal to the current through that containing $R_2$. In some manufactured receivers this is brought about by draining the screen voltage-divider current and the current to the screens of three r.f. pentodes (r.f. and i.f. stages) through $R_2$, the sum of these currents being about equal to the maximum plate current of one a.v.c.-controlled tube. Typical values for this type of circuit are given. The sensitivity can be increased by increasing the resistance of $R_1$, $R_2$ and $R_3$. The initial setting is made with the manual gain control set near maximum; when $R_3$ should be adjusted to make the meter read zero with no signal.

# A Six-Tube General-Coverage and Bandspread Superheterodyne

A superhet receiver of simple construction, having a wide frequency range for general coverage as well as full bandspread for amateur-band reception, is shown in Figs. 5-33 to 5-38. The circuit uses six tubes and gives continuous coverage from about 2.4 to 47 Mc. An additional range, 0.55 to 1.6 Mc., covers the broadcast band. The receiver is intended for a.c. operation and requires a filament supply delivering 6.3 volts at 1.65 amp., and a "B" supply of 200 to 250 volts at 70 ma.

## *The Circuit*

The circuit diagram is given in Fig. 5-34. A 6K8 is used as a combined oscillator-mixer followed by a 6SK7 i.f. amplifier. The intermediate frequency is 456 kc. The oscillator is tuned 456 kc. higher than the signal on frequencies up to 27 Mc.; above 27 Mc. the oscillator is 456 kc. lower than the signal. Manual gain control is applied to the i.f. stage, and a.v.c. is applied to both the i.f. and mixer stages. One section of the 6H6 double diode is used as the second detector and the other section as a series-type noise limiter. One half of a 6SL7GT double triode is used as the b.f.o. tube and the other triode is used in the first audio stage. The b.f.o. is capacity-coupled to the detector by soldering one end of an insulated wire to the detector plate and wrapping several turns of the wire around the b.f.o. grid lead. This capacity is designated $C_{C1}$ in the di-

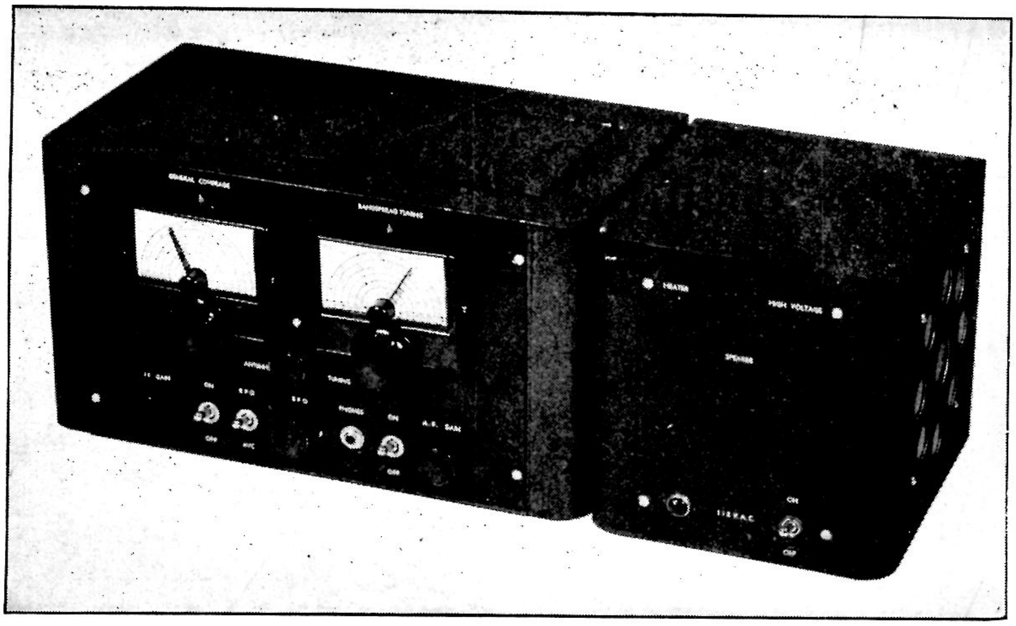


*Fig. 5-33* — A front view of the six-tube receiver. The general-coverage and the band-spread-tuning dials are to the left and right of the panel. The antenna trimmer condenser is centered below the main tuning dials. The i.f. gain control, noise-limiter switch, a.v.c.-b.f.o. switch, b.f.o. pitch control, headphone jack, stand-by switch, and a.f. gain control are shown from left to right across the bottom of the panel. A suitable power supply for the receiver is mounted in the 'speaker cabinet shown at the right of the receiver.

agram. A small mica by-pass condenser, $C_{21}$, connected between the first-audio grid and ground, prevents the b.f.o. signal from leaking into the audio system. This will occur without the by-pass condenser because of the capacity coupling that exists inside the 6SL7GT. The condenser can be connected from grid to ground, as shown, or from plate to ground.

Both headphone and loudspeaker output are available, with the headphone output taken from the plate circuit of the triode audio stage at $J_1$, and the loudspeaker output taken from the plate of the 6V6GT audio-amplifier stage. The 'phone jack is wired in a manner that grounds the 6V6 input grid when the 'phones are plugged in.

A VR-105 regulator tube is used to stabilize the voltage applied to the plates of the high-frequency oscillator and b.f.o. tubes, and to the screen grid of the i.f. amplifier. Positive bias for the 6SK7 is obtained from a bleeder network connected to the regulated supply.

### *Construction*

The parts arrangement is shown in Figs. 5-35 and 5-36 and the components are identified in the captions accompanying the photographs. The chassis on which the parts are mounted measures 2 × 7 × 11 inches, and the cabinet that houses the finished receiver measures 8 × 8¼ × 14½ inches. A panel measuring 8 × 12 inches is included as part of the cabinet, and the variable condensers are mounted on this panel. The sockets for the 6K8 and the r.f. coils are mounted on 1-inch-long metal pillars. Holes 1¼ inches in diameter are cut in the chassis just below the three sockets and the *below-deck* leads are passed through these holes. Although the No. 12 wire leads between the oscillator band-set condenser may appear to be unduly long, it is of no consequence since the leads are associated with the oscillator circuit rather than the mixer circuit. The dials used with these two condensers (shown in the front view) are Millen No. 10039. The bandspread condenser is a four-plate (two stator and two rotor) affair made by pulling out plates from the original. Inasmuch as bandspread is obtained by the parallel-condenser-and-coil method (this system allows adequate bandspread to be obtained without the bother of tapping the coils or using series condensers), it is necessary to adjust the $L/C$ ratio of the oscillator circuit until the desired degree of bandspread is obtained. A four-plate condenser of the type recommended will give full bandspread when used with the coils listed in the coil chart. The 3.5-Mc. band is broken into two sections — 3.5 to 3.8 Mc. and 3.7 to 4.0 Mc. This division, necessary if full bandspread is to be obtained on the higher-frequency bands, offers no operating disadvantage and, as a matter of fact, is a decided advantage because it gives full bandspread to the c.w. and 'phone portions of the 80-meter band.

**COIL TABLE FOR THE SIX-TUBE SUPERHETERODYNE**

| *Range* | *Turns* $L_1$ | $L_2$ | $L_3$ | $L_4$ |
|---|---|---|---|---|
| 550–1600 kc. | 12 | 150 | 12 | 67 |
| 2.4 to 6.45 Mc. (80 meters) | 12 | 31 | 10 | 29 |
| 6.4 to 13.8 Mc. (40 meters) | 10 | 17 | 7 | 12 |
| 13.8 to 27 Mc. (20 meters) | 9 | 10 | 7 | 5 |
| 23.5 to 47 Mc. (10–11 meters) | 3 | 3 | 3¼ | 1½ |

The broadcast-band coils are close-wound with No. 30 d.s.c. on 4-prong forms having a diam. of 1½ inches and a winding length of 2½ inches. $L_1$ is wound on a cardboard tube of ¾-inch diam., and is mounted inside the form. $L_2$ has a winding length of 2⅛ inches, and $L_4$ has a length of 15/16 inch. Coils for all other bands are wound on 1-inch diam. forms (Millen 45004) with No. 26 d.s.c. wire. The turns of $L_2$ and $L_4$, with exception of $L_4$ for 10 meters, are spaced to a length of 1 inch; $L_4$ for 10 meters is close-wound. Antenna and tickler coils, $L_1$ and $L_3$, are close-wound, spaced about 1/16 inch from the bottom of the main windings, except for 10 meters. The antenna coil for 10 meters is interwound with $L_2$, and the tickler coil is directly adjacent to $L_4$.

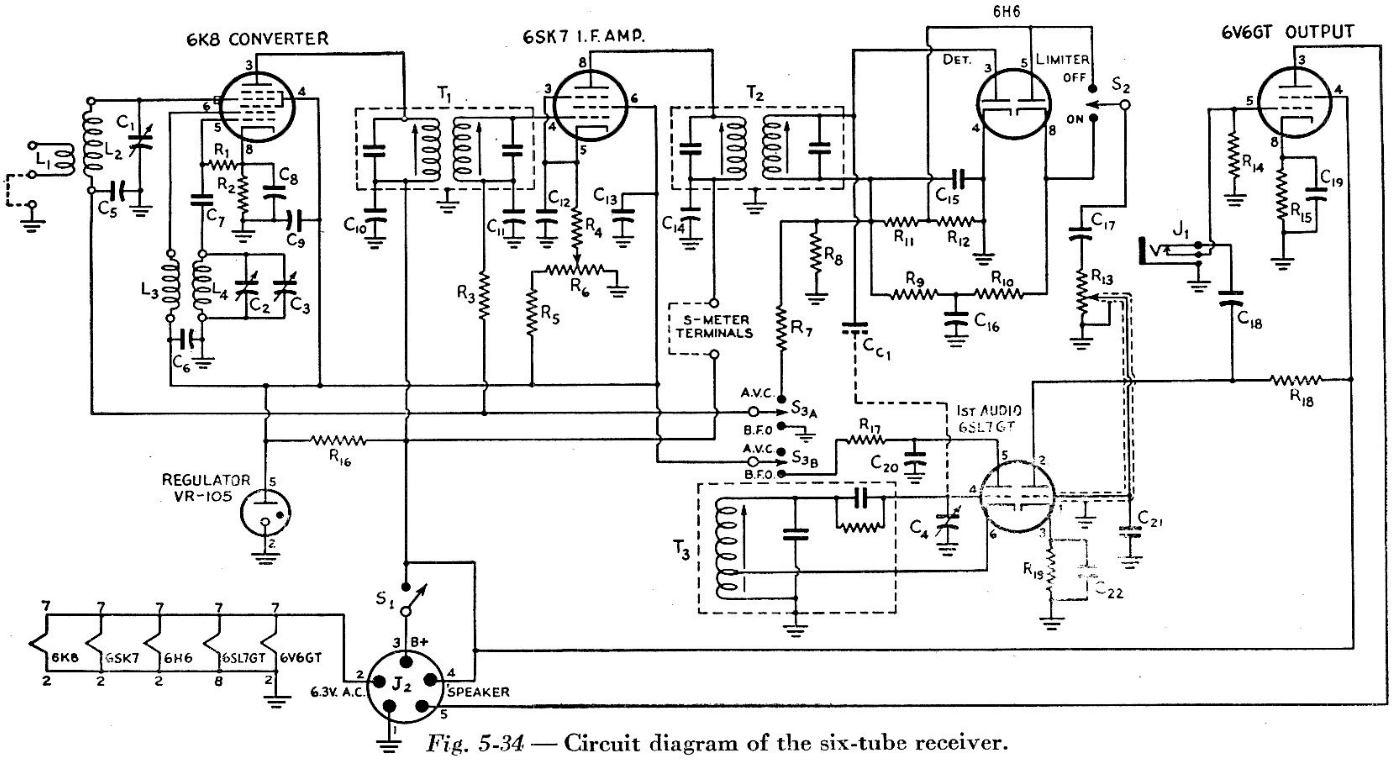



*Fig. 5-34* — Circuit diagram of the six-tube receiver.

$C_1$, $C_2$ — 140-$\mu\mu$fd. midget variable (Millen 22140).
$C_3$ — 25-$\mu\mu$fd. midget variable (Millen 20025).
$C_4$ — 35-$\mu\mu$fd. midget variable (Millen 20035).
$C_5$, $C_6$, $C_8$, $C_9$, $C_{10}$, $C_{11}$, $C_{12}$, $C_{13}$, $C_{14}$, $C_{16}$, $C_{18}$, $C_{20}$ — 0.01-$\mu$fd. paper, 400 volts.
$C_7$, $C_{15}$, $C_{21}$ — 100-$\mu\mu$fd. mica.
$C_{17}$ — 0.05-$\mu$fd. paper, 400 volts.
$C_{19}$, $C_{22}$ — 10-$\mu$fd. 25-volt electrolytic.
$C_{C1}$ — See text.
$R_1$, $R_5$, $R_{17}$ — 47,000 ohms.
$R_2$ — 220 ohms.
$R_3$, $R_{18}$ — 0.1 megohm.
$R_4$ — 180 ohms.
$R_6$ — 2000-ohm wire-wound potentiometer.
$R_7$, $R_8$, $R_9$ — 1.0 megohm.
$R_{10}$ — 0.82 megohm.
$R_{11}$, $R_{12}$ — 0.27 megohm.
$R_{13}$ — 1-megohm carbon potentiometer.
$R_{14}$ — 0.22 megohm.
$R_{15}$ — 220 ohms, 1 watt.
$R_{16}$ — 7500 ohms, 10 watts.
$R_{19}$ — 2200 ohms.
All resistors ½ watt unless otherwise noted.
$L_1$ through $L_4$ — See coil table.
$J_1$ — Closed-circuit jack.
$J_2$ — 5-prong chassis-mounting male plug.
$S_1$ — S.p.s.t. toggle switch.
$S_2$ — S.p.d.t. toggle switch.
$S_{3A\text{-}B}$ — D.p.d.t. toggle switch.
$T_1$ — 456-kc. interstage i.f. transformer, permeability-tuned (Millen 64455).
$T_2$ — 456-kc. diode transformer (Millen 64453).
$T_3$ — 456-kc. b.f.o. assembly (Millen 65456).

The mixer section was designed with short leads between the tuning condenser and r.f. coil. The tuning condenser, $C_1$, is mounted on the panel well down toward the coil socket. An "L"-shaped metal shield is mounted on the coil socket as shown in Fig. 5-35. The shield is 2½ inches wide, has a 2¼-inch base, and the vertical portion is 3¼ inches high. This shield was inserted between the mixer coil and the diode coupling transformer because of magnetic coupling that occurred between the two circuits when the receiver was operated at the lower frequencies. A 6-inch length of 150-ohm Twin-Lead is used between the antenna input terminals and the antenna winding, $L_1$. The by-pass con-

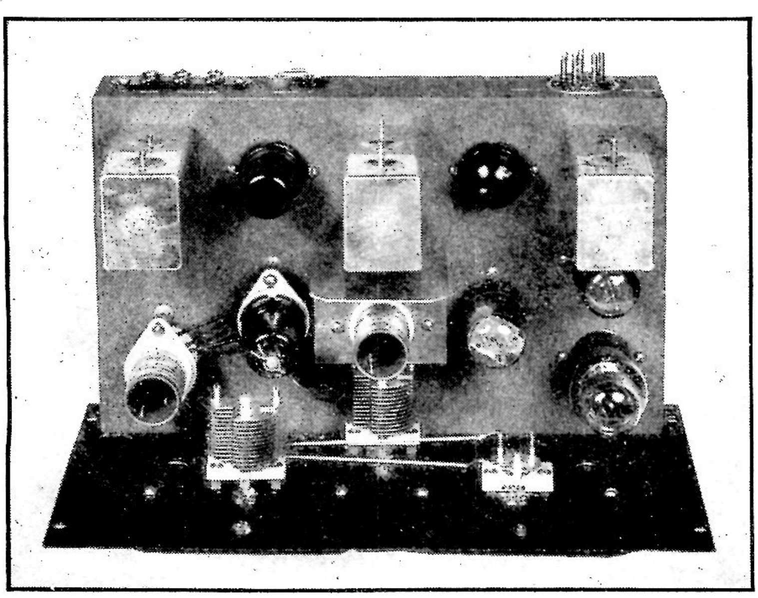


*Fig. 5-35* — A top view of the six-tube superheterodyne. Coupling transformers $T_1$ and $T_2$, the i.f. amplifier and double-diode tubes, and the b.f.o. assembly are in line across the rear of the chassis. The mixer and oscillator coils are to the right and left of the 6K8 converter tube. The sockets for the tube and the coils are mounted above the chassis. An aluminum shield is mounted on the mixer-coil socket and provides shielding between this coil and the diode-coupling transformer (see text). The 6SL7GT is to the right of the mixer coil, and the output and regulator tubes are at the right front of the chassis.

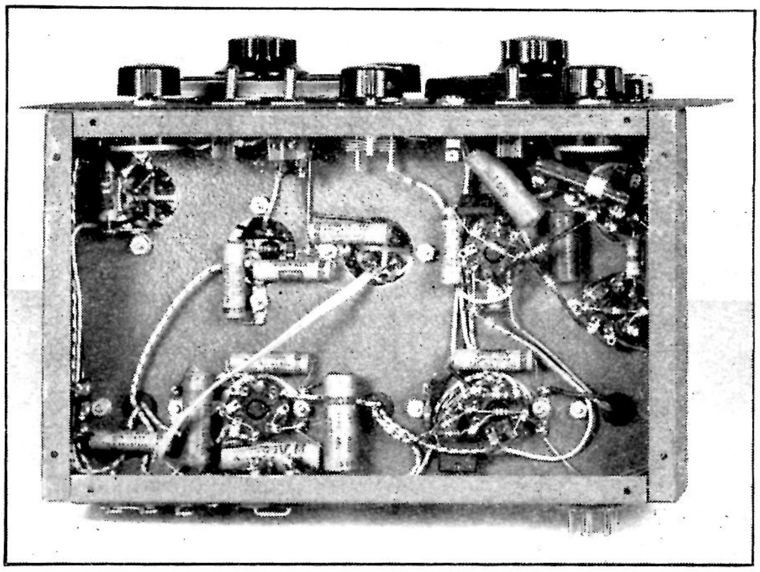


*Fig. 5-36* — This bottom view of the six-tube receiver shows how the by-pass and coupling condensers are closely grouped around the tube sockets. The chassis and panel are fastened together by the switches and controls mounted along the front wall of the chassis. The antenna terminals, S-meter tip-jacks, and the power-input connector, $J_2$, are at the rear.

densers for the r.f. section are mounted below the chassis and are connected with the shortest possible leads.

The wiring of the components associated with the rest of the receiver circuit does not call for any special treatment with the exception of the lead to the grid of the first audio amplifier — this lead is made with shielded wire. The by-pass condensers are kept as close as possible to the tubes and circuit points that they by-pass, and spare socket prongs are used as tie-points wherever possible. Only three independent tie-point strips are required: one is at the top left-hand corner (Fig. 5-36) for the connection between the 6SK7 cathode resistor and the lead to the tube, one at the lower left-hand corner for mounting of the 6SK7 grid-leak resistor, and one to the left of the double diode for mounting of the a.v.c. filter resistor, $R_7$. The pin-jacks, shown at the rear of the chassis, are shorted out, or eliminated entirely, if the use of an external S-meter unit is not contemplated.

### *Tuning*

The first step in putting the receiver into operation is to align the i.f. amplifier. This should preferably be done with the aid of a test oscillator, but if one is not available the circuits may be aligned on hiss or noise. The beat oscillator can also be used to furnish a signal for alignment, by coupling its output temporarily to the grid of the 6K8. A d.c. voltmeter can be used to measure the voltage developed across $R_{11}R_{12}$ during alignment.

When the i.f. is aligned, the mixer grid and oscillator coils for a band can be plugged in. $C_3$ should be set near maximum capacity and $C_2$ tuned from maximum capacity until a signal is heard. The capacity of $C_2$ should be further reduced until the same signal is heard a second time, this setting being the one that places the oscillator frequency on the high side of the incoming signal. The high-capacity setting of $C_2$ is used when the receiver is tuned to the 10- or 11-meter bands. The mixer tuning condenser, $C_1$, is adjusted for maximum signal strength after the oscillator adjustment has been completed. With $C_2$ set at the low-frequency end of an amateur band, further tuning should be done with $C_3$, and the band should be found to cover about ninety per cent of the dial. $C_3$ can of course be used for bandspread tuning outside as well as inside the amateur bands.

The bandspread condenser need not be tuned for coverage of the broadcast band. Of course, it can be used for vernier tuning in the event that critical tuning is required to separate stations operating on nearly the same frequency. Ordinarily, $C_3$ can be set at minimum capacity while the tuning of stations is done with $C_2$.

It is convenient to calibrate the receiver, using the scales provided for the purpose. Calibration points may be taken from incoming signals whose frequencies are known, from a calibrated test oscillator, or from the harmonics of a 100-kc. oscillator, as described in

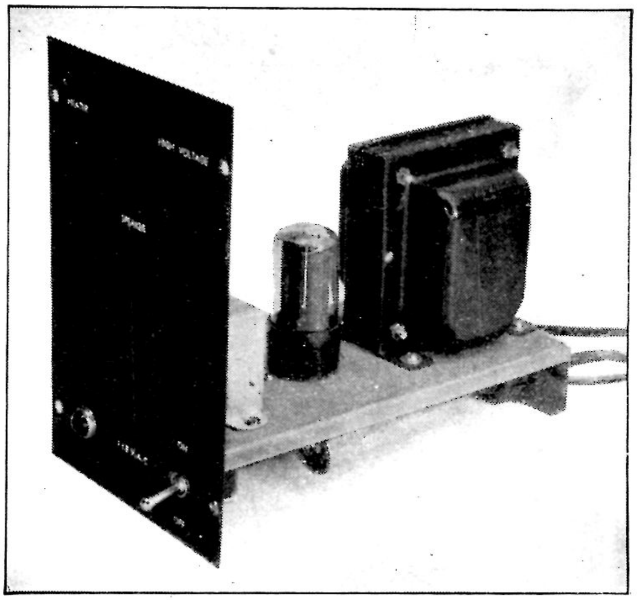


*Fig. 5-37* — A combination 'speaker-and-power-supply unit for the 6-tube superheterodyne receiver. The power transformer, filter choke and rectifier tube are mounted on an open-ended chassis measuring 1½ × 4 × 8 inches. The chassis and panel are fastened together by the toggle switch and the pilot-light assembly. The filter condensers are mounted below the chassis. A five-wire cable, at the rear of the chassis, connects the power supply and the 'speaker to the receiver. The 'speaker can be mounted on either the right or left side of the cabinet shown in Fig. 5-33.

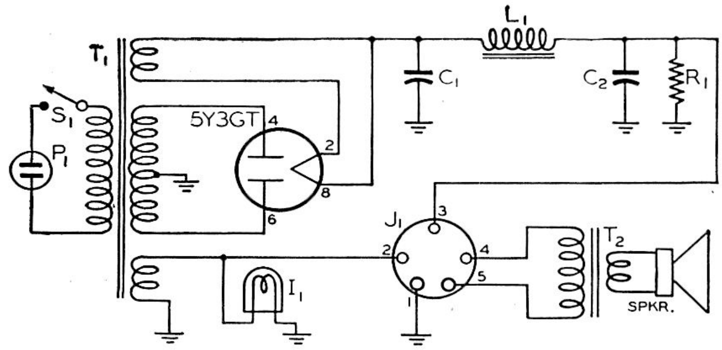



*Fig. 5-38* — Power-supply circuit diagram.
C1, C2 — 20-μfd. 450-volt paper electrolytic.
R1 — 47,000 ohms, 2 watts.
L1 — 8-henry 75-ma. filter choke (Stancor C-1355).
I1 — 6.3-volt pilot-lamp-and-socket assembly.
J1 — 5-prong socket on power cable.
P1 — 115-volt a.c. line plug.
S1 — S.p.s.t. toggle switch.
SPKR — Six-inch permanent magnet 'speaker.
T1 — Power transformer, 350 volts each side of center-tap, 70-ma. rating. Filament windings: 5 v. 3 amp.; 6.3 v. 3 amp. (Stancor P-4078).

Chapter Sixteen. The mixer need not be calibrated, since tuning of the mixer circuit has little effect on the oscillator frequency. However, a mixer calibration can be made to insure that the circuit is tuned to the desired signal rather than to an image.

### *Power Supply*

A power supply suitable for the 6-tube receiver is shown in Figs. 5-37 and 5-38. The arrangement of components can be determined from Fig. 5-37. An output transformer and 'speaker are both mounted in the 7½ × 8 × 8-inch cabinet that houses the power supply. This particular supply need not be used with the receiver if another supply of suitable design is available. It is only necessary that the supply deliver a well-filtered output of 250 volts at 70 ma.

## A Signal-Strength Indicator (S-Meter)

If your receiver has no built-in S-meter and you would like one for comparing signal strengths (and for help in aligning your receiver), the unit shown in Figs. 5-39 and 5-41 can be used. The wiring diagram, Fig. 5-40, is an adaptation of Fig. 5-32C, and uses a 0–1 milliammeter as the indicator. A variable shunt, $R_1$, allows the meter sensitivity to be regulated to suit the particular receiver, and $R_4$ is for setting the meter to zero with no signal. The meter can be connected in the plate circuit of any r.f. or i.f. amplifier controlled by the a.v.c.

The construction of the unit is apparent from the photographs and is not at all critical.

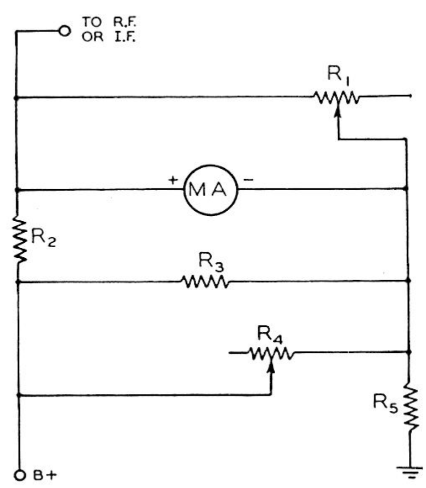



*Fig. 5-40* — Wiring diagram of the signal-strength indicator.
R1 — 100-ohm wire-wound potentiometer.
R2 — 220 ohms, ½ watt.
R3 — 680 ohms, ½ watt.
R4 — 1000-ohm wire-wound potentiometer.
R5 — 4700 ohms, 1 watt.
MA — 0–1 ma. d.c. meter.

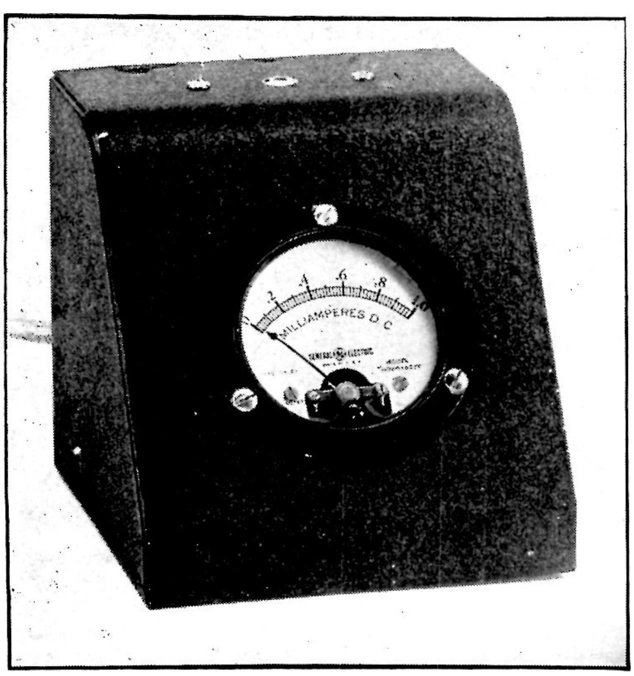



*Fig. 5-39* — Front view of the signal-strength indicator. The 0–1 milliammeter is mounted in a metal meter case. The zero-adjustment potentiometer, $R_4$, is mounted below the top of the cabinet by means of a "U"-shaped bracket; the potentiometer shaft is slotted so that it can be adjusted with a screwdriver. A new face, calibrated in S-units, can be pasted to the 0–1 ma. scale, or a calibration chart can be attached to the cabinet.

If possible and desirable, the meter and circuit can be built into the receiver instead of being made a separate unit.

### *S-Meter Calibration*

It is customary to calibrate in terms of S-units up to about midscale, and then in "decibels above S9" over the upper half of the scale. Although there are no standards, current practice is to use about 6-db. steps in the S-scale, and a 100-microvolt signal for "S9."

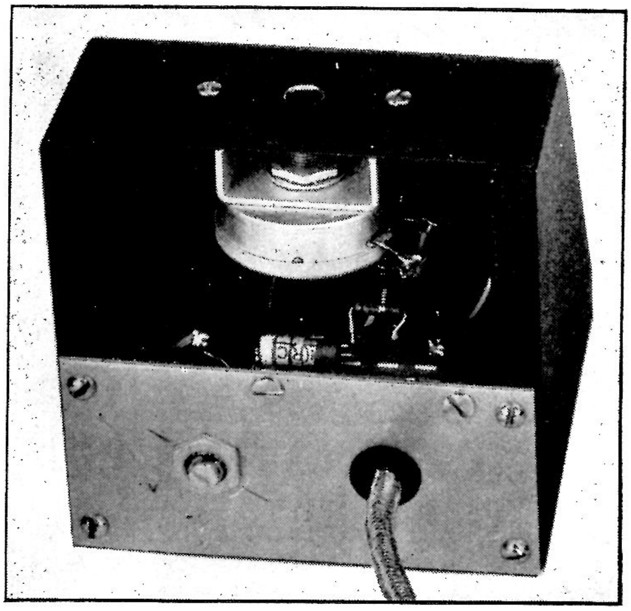


*Fig. 5-41* — This rear view of the S-meter shows the meter shunt, $R_1$, and a tie-point strip mounted on a metal strip attached to the rear side of the meter cabinet. Resistors $R_2$, $R_3$ and $R_5$ are mounted on the tie-point strip. A three-wire cable, running out of the case through a rubber grommet, connects the meter to the receiver.

Such a calibration requires an accurate r.f. signal generator, and relatively few amateurs have access to laboratory equipment of this type. Also, the scale will be accurate only on the radio frequency at which the calibration is made. On different bands — or even in different parts of the same band — the r.f. gain of the receiver will change and the calibration will not hold.

An S-meter is principally useful for making comparisons between signals on or near the same frequency. For this purpose it is entirely satisfactory to choose arbitrarily a signal that seems to you to be about the right strength to represent "S9," adjust the meter sensitivity to give a suitable reading on that signal, and then divide off the scale into equal intervals from zero to 9.

Alternatively, points can be taken by comparing with another receiver that does have a calibrated S-meter. The two receivers may be connected to the same antenna so that simultaneous measurements can be made on incoming signals, provided their antenna input impedances are not widely different. Local signals should be used to avoid fading effects.

# Improving Receiver Selectivity

## ● INTERMEDIATE-FREQUENCY AMPLIFIERS

As mentioned earlier in this chapter, one of the big advantages of the superheterodyne receiver is the improved selectivity that is possible. This selectivity is obtained in the i.f. amplifier, where the lower frequency allows more selectivity per stage than at the higher signal frequency. For 'phone reception, the limit to useful selectivity in the i.f. amplifier is the point where so many of the sidebands are cut that intelligibility is lost, although it is possible to remove completely one full set of sidebands without impairing the quality at all. Maximum receiver selectivity in 'phone reception requires excellent stability in both transmitter and receiver, so that they will both remain "in tune" during the transmission. The limit to useful selectivity in code work is around 50 or 100 *cycles* for hand-key speeds, but it is difficult to use this much selectivity because it requires remarkable stability in both transmitter and receiver, and to tune in a signal becomes a major problem.

### *Single-Signal Effect*

In heterodyne c.w. reception with a superheterodyne receiver, the beat oscillator is set to give a suitable audio-frequency beat-note when the incoming signal is converted to the intermediate frequency. For example, the beat oscillator may be set to 456 kc. (the i.f. being 455 kc.) to give a 1000-cycle beat-note. Now, if an interfering signal appears at 457 kc., or if the receiver is tuned to heterodyne the incoming signal to 457 kc., it will also be heterodyned by the beat oscillator to produce a 1000-cycle beat. Hence every signal can be tuned in at two places that will give a 1000-cycle beat (or any other low audio frequency). This **audio-frequency image** effect can be reduced if the i.f. selectivity is such that the incoming signal, when heterodyned to 457 kc., is greatly attenuated.

When this is done, tuning through a given signal will show a strong response at the desired beat-note on one side of zero beat only, instead of the two beat-notes on either side of zero beat characteristic of less-selective reception, hence the name: **single-signal reception.**

The necessary selectivity is difficult to obtain with nonregenerative amplifiers using ordinary tuned circuits unless a very low i.f. or a large number of circuits is used. In practice it is secured either by regenerative amplification or by a crystal filter.

### *Regeneration*

Regeneration can be used to give a pronounced single-signal effect, particularly when the i.f. is 455 kc. or lower. The resonance curve of an i.f. stage at critical regeneration (just below the oscillating point) is extremely sharp, a bandwidth of 1 kc. at 10 times down and 5 kc. at 100 times down being obtainable in one stage. The audio-frequency image of a given signal thus can be reduced by a factor of

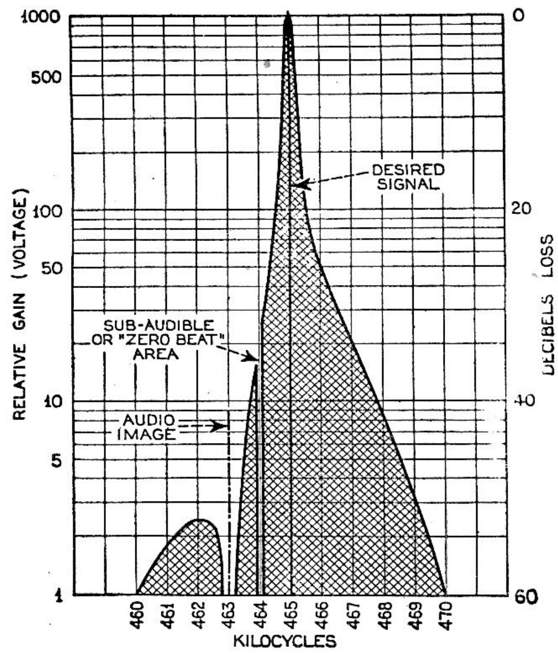



*Fig. 5-42* — **Graphical representation of single-signal selectivity. The shaded area indicates the over-all bandwidth, or region in which response is obtainable.**

nearly 100 for a 1000-cycle beat-note (image 2000 cycles from resonance).

Regeneration is easily introduced into an i.f. amplifier by providing a small amount of capacity coupling between grid and plate. Bringing a short length of wire, connected to the grid, into the vicinity of the plate lead usually will suffice. The feed-back may be controlled by the regular cathode-resistor gain control. When the i.f. is regenerative, it is preferable to operate the tube at reduced gain (high bias) and depend on regeneration to bring up the signal strength. This prevents overloading and increases selectivity.

The higher selectivity with regeneration reduces the over-all response to noise generated in the earlier stages of the receiver, just as does high selectivity produced by other means, and therefore improves the signal-to-noise ratio. The disadvantage is that the regenerative gain varies with signal strength, being less on strong signals, and the receiver selectivity varies accordingly.

### *Crystal Filters*

The most satisfactory method of obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i.f. amplifier. Compared to a good tuned circuit, the $Q$ of such a crystal is extremely high. The dimensions of the crystal are made such that it is resonant at the desired intermediate frequency. It is then used as a selective coupler between i.f. stages.

Fig. 5-42 gives a typical crystal-filter resonance curve. For single-signal reception, the audio-frequency image can be reduced by a factor of 1000 or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides great discrimination against signals very close to the desired signal in frequency, and, by reducing the bandwidth, reduces the response of the receiver to noise both from sources external to the receiver and in the radio-frequency stages of the receiver itself.

### *Crystal-Filter Circuits; Phasing*

Several crystal-filter circuits are shown in Fig. 5-43. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit, with the secondary side of $T_1$, the input transformer, balanced to ground either through a pair of condensers, $C$-$C$ (A), or by a center-tap on the secondary, $L_2$ (B). The bridge is completed by the crystal and the *phasing condenser*, $C_2$, which has a maximum capacity somewhat higher than the capacity of the crystal in its holder. When $C_2$ is set to balance the crystal-holder capacity, the resonance curve of the crystal circuit is practically symmetrical; the crystal acts as a series-resonant circuit of very high $Q$ and thus allows signals of the desired frequency to be fed through $C_3$ to $L_3L_4$, the output transformer. Without $C_2$, the holder capacity (with the crystal acting as a dielectric) would pass signals of undesired frequencies.

The phasing control has an additional function besides neutralization of the crystal-holder capacity. The holder capacity becomes a part of the crystal circuit and causes it to act as a parallel-tuned resonant circuit at a frequency slightly higher than its series-resonant frequency. Signals at the parallel-resonant frequency thus are prevented from reaching the output circuit. The phasing control, by varying the effect of the holder capacity, permits shifting the parallel-resonant frequency over a considerable range, providing adjustable rejection of interfering signals. The effect of rejection is illustrated in Fig. 5-42, where the audio image is reduced, by proper setting of the phasing control, far below the value that would be expected if the resonance curve were symmetrical.

### *Variable Selectivity*

In circuits such as A and B, Fig 5-43, variable selectivity is obtained by adjustment of the variable input impedance, which is effectively in series with the crystal resonator. This is accomplished by varying $C_1$ (the *selectivity control*), which tunes the balanced secondary circuit of $T_1$. When the secondary is tuned to i.f. resonance the parallel impedance of the $L_2C_1$ combination is maximum and is purely resistive. Since the secondary circuit is center-tapped, approximately one-fourth of this resistive impedance is in series with the crystal through $C_3$ and $L_4$. This lowers the $Q$ of the crystal circuit and makes its over-all selectivity minimum. At the same time, the voltage applied to the crystal circuit is maximum.

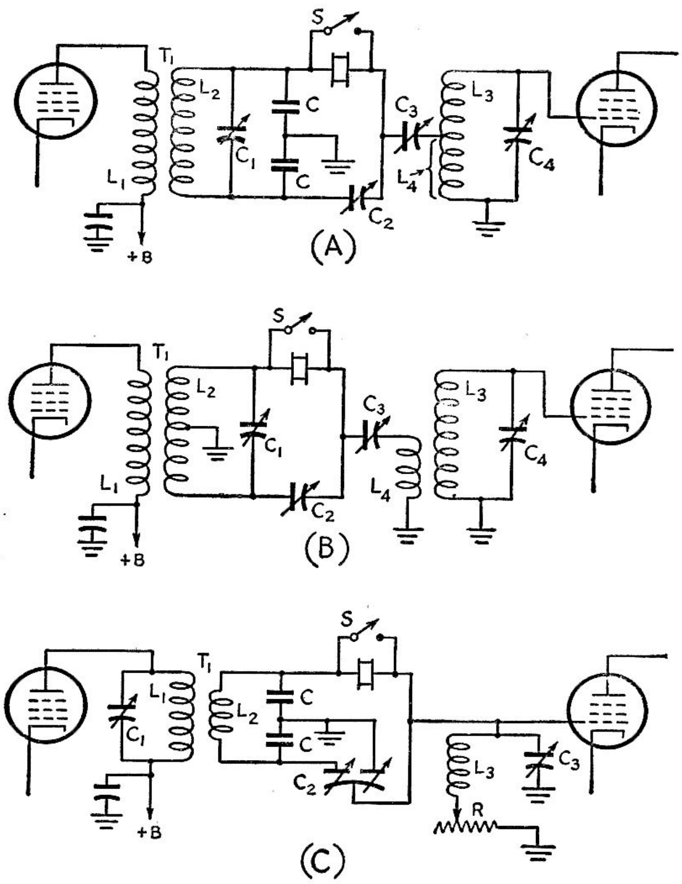



*Fig. 5-43* — Crystal-filter circuits of three types. All give variable bandwidth, with C having the greatest range of selectivity. Their operation is discussed in the text. Suitable circuit values are as follows: Circuit A, $T_1$, special i.f. input transformer with high-inductance primary, $L_1$, closely coupled to tuned secondary, $L_2$; $C_1$, 50-$\mu\mu$fd. variable; $C$, each 100-$\mu\mu$fd. fixed (mica); $C_2$, 10- to 15-$\mu\mu$fd. (max.) variable; $C_3$, 50-$\mu\mu$fd. trimmer; $L_3C_4$, i.f. tuned circuit, with $L_3$ tapped to match crystal-circuit impedance. In circuit B, $T_1$ is the same as in circuit A except that the secondary is center-tapped; $C_1$ is 100-$\mu\mu$fd. variable; $C_2$, $C_3$ and $C_4$, same as for circuit A; $L_2L_4$ is a transformer with primary, $L_4$, corresponding to tap on $L_3$ in A. In circuit C, $T_1$ is a special i.f. input transformer with tuned primary and low-impedance secondary; $C$, each 100-$\mu\mu$fd. fixed (mica); $C_2$, opposed stator phasing condenser, approximately 8-$\mu\mu$fd. maximum capacity each side; $L_3C_3$, high-$Q$ i.f. tuned circuit; $R$, 0 to 3000 ohms (selectivity control).

## RADIO-FREQUENCY AMPLIFIERS

While selectivity to reduce audio-frequency images can be built into the i.f. amplifier, discrimination against radio-frequency images can only be obtained in circuits ahead of the first detector. These tuned circuits and their associated vacuum tubes are called **radio-frequency amplifiers.** For top performance of a communications receiver on frequencies above 7 Mc., it is mandatory that it have one or two stages of r.f. amplification, for image rejection and improved sensitivity. (The improvement in sensitivity that can be obtained will be discussed later.)

Receivers with an i.f. of 455 kc. can be expected to have some r.f. image response at a signal frequency of 14 Mc. and higher if only one stage of r.f. amplification is used. (Regeneration in the r.f. amplifier will reduce image response, but regeneration is often a tricky thing to control.) With two stages of r.f. amplification and an i.f. of 455 kc., no images should be apparent at 14 Mc., but they will show up on 28 Mc. and higher. Three stages or more of r.f. amplification, with an i.f. of 455 kc., will reduce the images at 28 Mc., but it really takes four or more stages to do a good job. The better solution at 28 Mc. is to use a "triple-detection" superheterodyne, with one stage of r.f. amplification and a first i.f. of 1600 kc. or higher. A regular receiver with an i.f. of 455 kc. can be converted to a triple-detection superhet by connecting a "converter" (to be described later) ahead of the receiver.

For best selectivity, r.f. amplifiers should use high-$Q$ circuits and tubes with high input and output resistance. Variable-$\mu$ pentodes are practically always used, although triodes (neutralized or otherwise connected so that they won't oscillate) are often used on the higher frequencies because they introduce less noise. Pentodes are better where maximum image rejection is desired, because they have less loading effect on the circuits.

## CROSS-MODULATION

Since a one- or two-stage r.f. amplifier will have a passband measured in hundreds of kc. at 14 Mc. or higher, strong signals will be amplified through the r.f. amplifier even though it is not tuned exactly to them. If these signals are strong enough, their amplified magnitude may be measurable in volts after passing through several r.f. stages. If this undesired signal is strong enough after amplification in the r.f. stages to shift the operating point of a tube (by driving the grid into the positive region), the undesired signal will modulate the desired signal. This effect is called **cross-modulation,** and is often encountered in receivers with several r.f. stages that are working at high gain. It is readily detectable as a superimposed modulation on the signal being listened to, and often the effect is that a signal can be tuned in at several points. It can be reduced or eliminated by greater selectivity in the antenna and r.f. stages (difficult to obtain), the use of variable-$\mu$ tubes in the r.f. amplifier, reduced gain in the r.f. amplifier, or reduced antenna input to the receiver.

### *Gain Control*

To avoid cross-modulation and other overload effects in the first detector and r.f. stages, the gain of the r.f. stages is usually made adjustable. This is accomplished by using variable-$\mu$ tubes and varying the d.c. grid bias, either in the grid or cathode circuit. If the gain control is automatic, as in the case of a.v.c., the bias is controlled in the grid circuit. Manual control of r.f. gain is generally done in the cathode circuit. A typical r.f. amplifier stage with the two types of gain control is shown in Fig. 5-44.

### *Tracking*

In a simple receiver with no r.f. stage, it is no inconvenience to adjust the high-frequency oscillator and the mixer circuit independently, because the mixer tuning is broad and requires little attention over an amateur band. However, when r.f. stages are added ahead of the mixer, the selectivity of the r.f. stages and mixer makes it awkward to use a two-control receiver over an entire amateur band, even though the mixer and r.f. stages are ganged and require only one control. Hence most receivers with one or more r.f. stages gang all of the tuning controls to give a single-tuning-control receiver. Obviously there must exist a constant difference in frequency (the i.f.) between the oscillator and the mixer/r.f. circuits, and when this condition is achieved the circuits are said to **track.**

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 5-45. The **tracking capacity,** $C_5$, commonly consists of two condensers in parallel, a fixed one of somewhat less capacity than the value needed and a smaller variable in parallel to allow for adjustment to the exact proper value. In practice, the trimmer, $C_4$, is first set for the high-frequency end of the tuning range, and then the tracking condenser is set for the low-frequency end. The tracking capacity becomes larger as the percentage difference between the oscillator and signal frequencies becomes smaller (that is, as the signal frequency becomes higher). Typical circuit values are given in the tables under Fig. 5-45. The coils can be calculated quite closely by using the *ARRL Lightning Calculator*, but they will have to be trimmed in the circuit for best tracking.

In amateur-band receivers, tracking is simplified by choosing a bandspread circuit that gives practically straight-line-frequency tuning (equal frequency change for each dial division), and then adjusting the oscillator and mixer tuned circuits so that both cover the same total number of kilocycles. For example, if the i.f. is 455 kc. and the mixer circuit tunes from 7000 to 7300 kc. between two given points on the dial, then the oscillator must tune from 7455 to 7755 kc. between the same two dial readings. With the bandspread arrangement of Fig. 5-12C, the tuning will be practically straight-line-frequency if the capacity actually in use at $C_2$ is not too small; the same is true of 5-12A if $C_1$ is small compared to $C_2$.

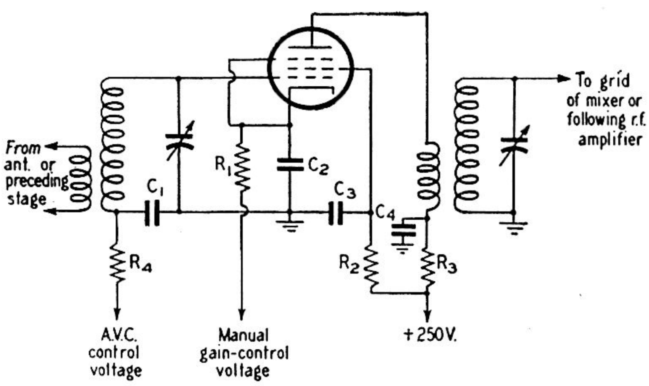



*Fig. 5-44* — Typical radio-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

$C_1$, $C_2$, $C_3$, $C_4$ — 0.01 $\mu$fd. below 15 Mc., 0.001 $\mu$fd. at 30 Mc.
$R_1$, $R_2$ — See Table 5-II.
$R_3$ — 1800 ohms.
$R_4$ — 0.22 megohm.

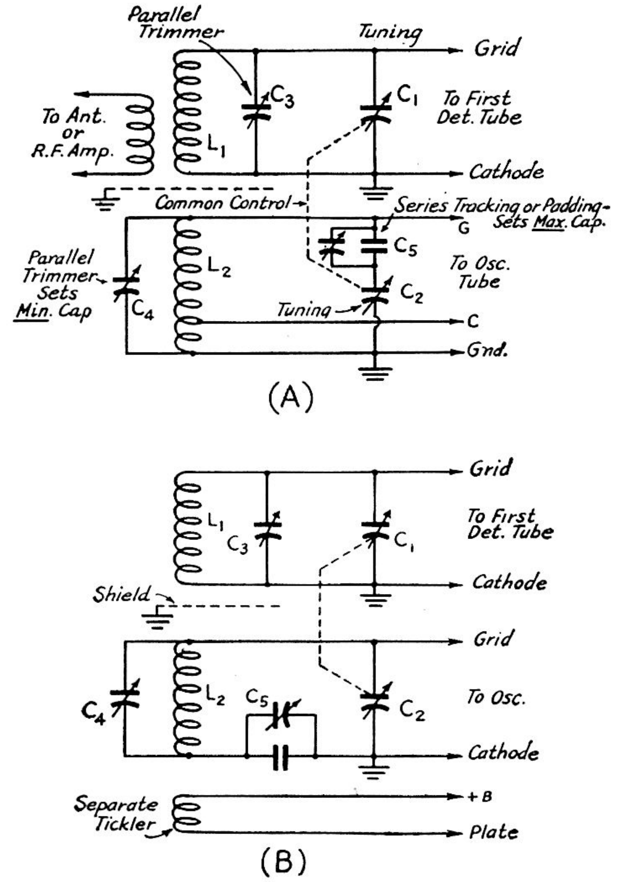



*Fig. 5-45* — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and $C_2$ having 140-$\mu\mu$fd. maximum, and the total minimum capacitance, including $C_3$ or $C_4$, being 30 to 35 $\mu\mu$fd.

| *Tuning Range* | $L_1$ | $L_2$ | $C_5$ |
|---|---|---|---|
| 1.7–4 Mc. | 50 $\mu$h. | 40 $\mu$h. | 0.0013 $\mu$fd. |
| 3.7–7.5 Mc. | 14 $\mu$h. | 12.2 $\mu$h. | 0.0022 $\mu$fd. |
| 7–15 Mc. | 3.5 $\mu$h. | 3 $\mu$h. | 0.0045 $\mu$fd. |
| 14–30 Mc. | 0.8 $\mu$h. | 0.78 $\mu$h. | None used |

Approximate values for 450- to 465-kc. i.f.s with a 2.5-to-1 tuning range, $C_1$ and $C_2$ being 350-$\mu\mu$fd. maximum, minimum including $C_3$ and $C_4$ being 40 to 50 $\mu\mu$fd.

| *Tuning Range* | $L_1$ | $L_3$ | $C_5$ |
|---|---|---|---|
| 0.5–1.5 Mc. | 240 $\mu$h. | 130 $\mu$h. | 425 $\mu\mu$fd. |
| 1.5–4 Mc. | 32 $\mu$h. | 25 $\mu$h. | 0.00115 $\mu$fd. |
| 4–10 Mc. | 4.5 $\mu$h. | 4 $\mu$h. | 0.0028 $\mu$fd. |
| 10–25 Mc. | 0.8 $\mu$h. | 0.75 $\mu$h. | None used |

## An Amateur-Band Eight-Tube Superheterodyne

An advanced type of amateur receiver incorporating one r.f. amplifier stage, variable i.f. selectivity and audio noise limiting is shown in Figs. 5-46, 5-48 and 5-49. As can be seen from the circuit in Fig. 5-47, a 6SG7 pentode is used for the tuned r.f. stage ahead of the 6K8 converter. An antenna compensator, $C_4$, controlled from the panel, allows one to trim up the r.f. stage when using different antennas that might modify the tracking. The cathode bias resistor of the r.f. stage is made as low as possible consistent with the tube ratings, to keep the gain and hence the signal-to-noise ratio of the stage high. The oscillator portion of the 6K8 mixer is tuned to the high-frequency side of the signal except on the 28-Mc. band, the usual custom nowadays in communications receivers. The oscillator tuning condenser, $C_{17}$, is of higher capacity than the r.f. and mixer tuning condensers, in the interest of better oscillator stability.

The i.f. amplifier is tuned to 455 kc., and the first stage is made regenerative by soldering a short length of wire to the plate terminal of the socket and running it near the grid terminal, as indicated by $C_{C1}$ in the diagram. Regeneration is controlled by reducing the gain of the tube, and $R_{12}$, a variable cathode-bias control, serves this function. The second i.f. stage uses a 6K7, selected because high gain is not necessary at this point.

Manual gain-control voltage is applied to the r.f. and second i.f. stages. It is not applied to the mixer because it might pull the oscillator frequency, and it is not tied in with the first i.f. amplifier because it would interlock with the regeneration control used for controlling the selectivity. However, the a.v.c. voltage is applied to the r.f. and both i.f. stages, with the result that the selectivity of the regenerative stage decreases with loud signals and gives a measure of automatic selectivity control. Using a negative-voltage power supply for the manual gain control is more expensive than the familiar cathode control, but it allows a wide range of control with less dissipation in the components. The a.v.c. is of the delayed type, the a.v.c. diode being biased about 1½ volts by the cathode resistor of the diode-triode detector-audio stage.

The second-detector-and-first-audio is the usual diode-triode combination and uses a 6SQ7. A 1N34 crystal diode is used as a noise limiter, and is left in the circuit all of the time. As is common with this type of circuit, it has little or no effect when the b.f.o. is on, but it is of considerable help to 'phone reception on the bands where automobile ignition is a factor. The constructor can satisfy himself on its operation when first building the receiver and working on it out of the case. By leaving one end of the 1N34 floating and touching it to the proper point in the circuit, a marked drop in ignition noise will be noted.

The b.f.o. is capacity-coupled to the detector by soldering one end of an insulated wire to the a.v.c. diode plate and wrapping several turns of the wire around the b.f.o. grid lead. This capacity is designated $C_{C2}$ in the diagram. The wire was connected to the a.v.c. diode plate lead for wiring convenience — the a.v.c. coupling condenser, $C_{32}$, passing the b.f.o. voltage without introducing appreciable attenuation.

Headphone output is obtained from the plate circuit of the 6SQ7 at $J_1$, and loudspeaker output is available from the 6F6 audio-amplifier stage. High-impedance or crystal headphones are recommended for maximum headphone output.

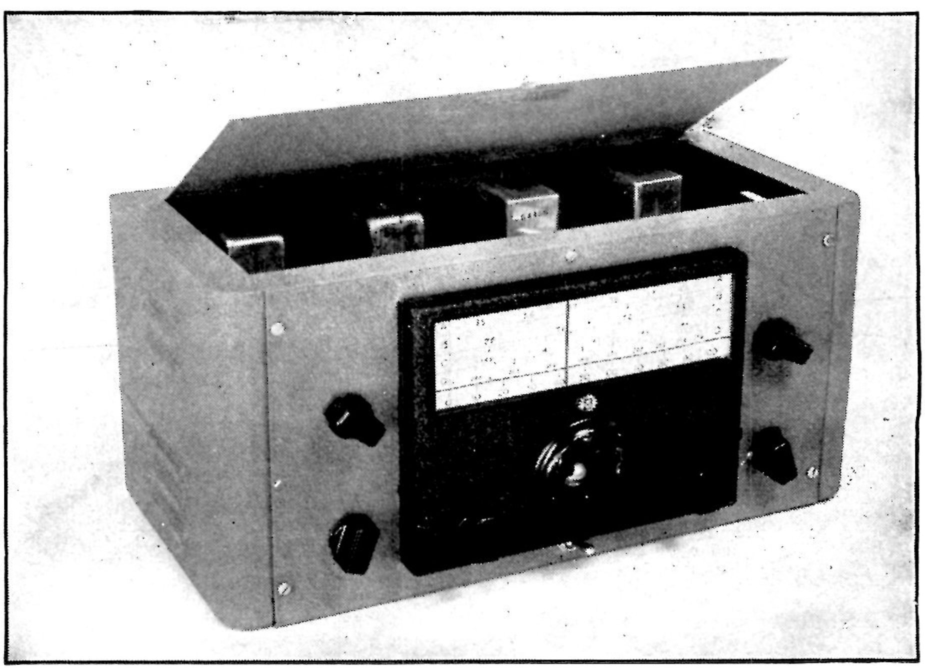


Fig. 5-46 — An amateur-band eight-tube receiver. The knobs on the left control audio volume (upper) and b.f.o. pitch, and the two on the right handle r.f. and i.f. gain (upper) and i.f. regeneration. The knob to the left of the large tuning knob is fastened to the *MAN.-A.V.C.-B.F.O* switch, and the one on the right is for the antenna trimmer. The toggle switch under the dial throws high negative bias on the r.f. stage during transmission periods.

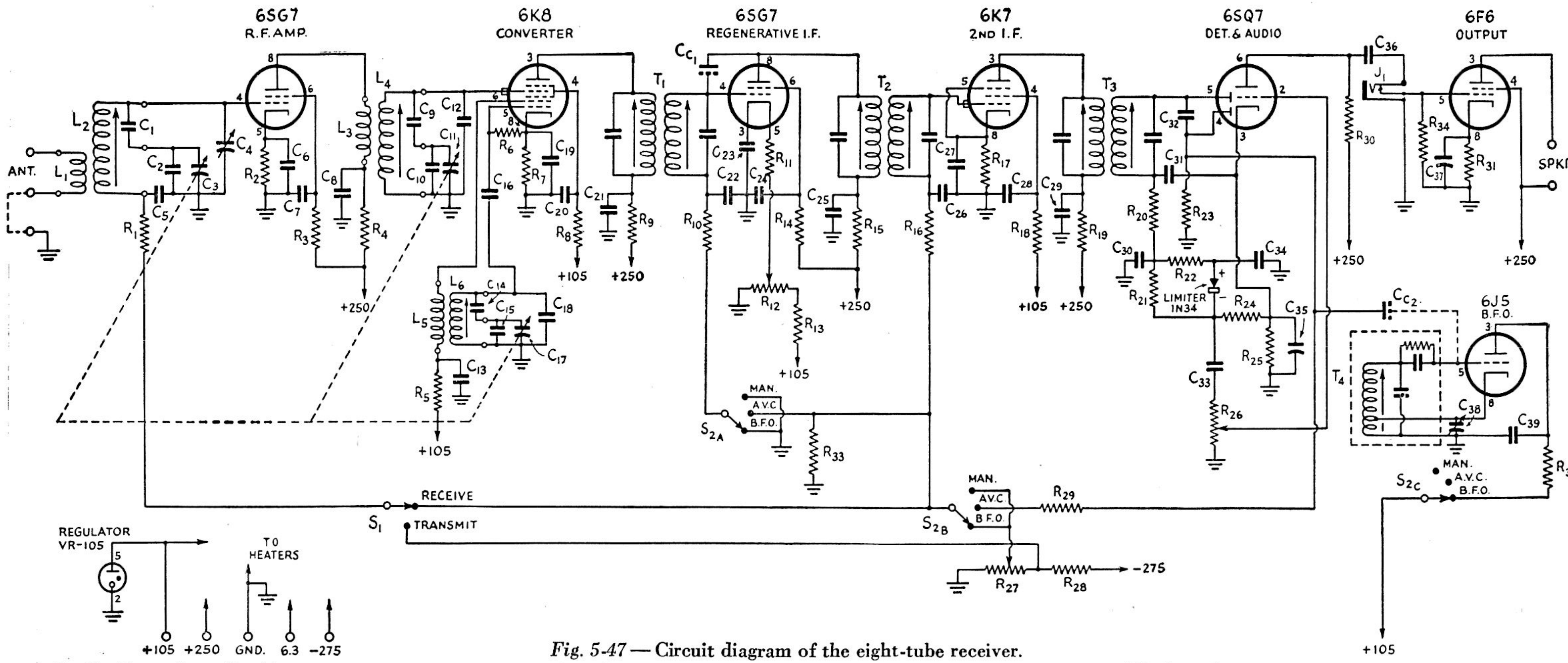



*Fig. 5-47* — **Circuit diagram of the eight-tube receiver.**

$C_1$, $C_9$, $C_{14}$ — See coil table.
$C_2$, $C_{10}$, $C_{12}$, $C_{18}$ — 10-$\mu\mu$fd. ceramic.
$C_3$, $C_{11}$ — 15-$\mu\mu$fd. midget variable (National UM-15).
$C_4$ — 15-$\mu\mu$fd. midget variable (Hammarlund HF-15).
$C_5$, $C_6$, $C_7$, $C_8$, $C_{13}$, $C_{19}$, $C_{20}$, $C_{21}$, $C_{22}$, $C_{23}$, $C_{24}$, $C_{25}$, $C_{26}$, $C_{27}$, $C_{28}$, $C_{29}$, $C_{39}$ — 0.01-$\mu$fd. mica.
$C_{15}$ — 37-$\mu\mu$fd. ceramic (10 and 27 in parallel).
$C_{16}$, $C_{30}$, $C_{32}$ — 100-$\mu\mu$fd. mica.
$C_{17}$ — 35-$\mu\mu$fd. midget variable (National UM-35).
$C_{31}$ — 220-$\mu\mu$fd. mica.
$C_{33}$ — 0.05-$\mu$fd. paper, 200 volts.
$C_{34}$ — 0.1-$\mu$fd. paper, 200 volts.
$C_{35}$, $C_{37}$ — 10-$\mu$fd. 25-volt electrolytic.
$C_{36}$ — 0.1-$\mu$fd. paper, 400 volts.
$C_{38}$ — 35-$\mu\mu$fd. midget variable (Hammarlund HF-35).
$C_{C1}$, $C_{C2}$ — See text.
$R_1$, $R_{10}$, $R_{16}$, $R_{30}$ — 0.1 megohm.
$R_2$ — 68 ohms.
$R_3$, $R_{14}$ — 33,000 ohms.
$R_4$, $R_5$, $R_8$, $R_9$, $R_{15}$, $R_{18}$, $R_{19}$ — 470 ohms.
$R_6$, $R_{13}$, $R_{20}$, $R_{21}$ — 47,000 ohms.
$R_7$ — 220 ohms.
$R_{11}$ — 180 ohms.
$R_{12}$ — 2000-ohm wire-wound potentiometer.
$R_{17}$ — 330 ohms.
$R_{22}$, $R_{23}$, $R_{29}$, $R_{33}$ — 1.0 megohm.
$R_{24}$, $R_{28}$ — 0.15 megohm.
$R_{25}$ — 2700 ohms.
$R_{26}$ — 1.0-megohm carbon potentiometer.
$R_{27}$ — 25,000-ohm carbon potentiometer.
$R_{31}$ — 470 ohms, 1 watt.
$R_{32}$ — 27,000 ohms.
$R_{34}$ — 0.22 megohm.
All resistors ½ watt unless otherwise noted.
$L_1$ through $L_6$ — See coil table.
$J_1$ — Closed-circuit jack.
$S_1$ — S.p.d.t. toggle switch.
$S_{2A\text{-}B\text{-}C}$ — Three-pole 3-position wafer switch (Centralab 2507).
$T_1$, $T_2$ — 456-kc. interstage i.f. transformer, permeability-tuned (Millen 64456).
$T_3$ — 456-kc. diode transformer, permeability-tuned (Millen 64454).
$T_4$ — 456-kc. b.f.o. assembly, permeability-tuned (Millen 65456).

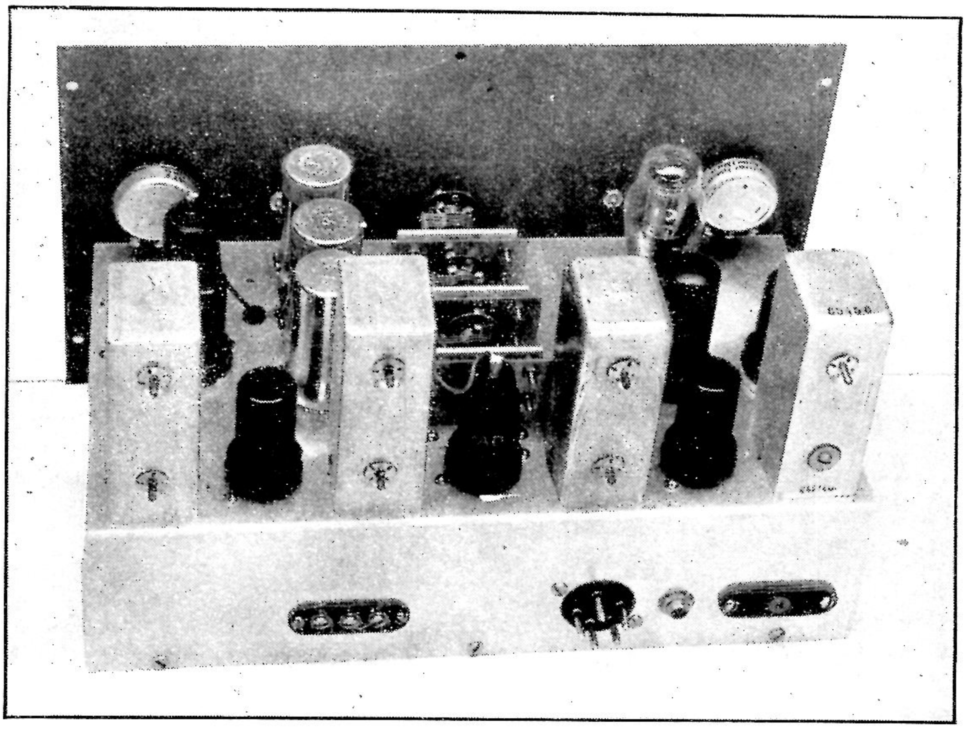


*Fig. 5-48* — This view of the eight-tube receiver chassis shows the mounting of the tuning condensers and the placement of most of the large components. The three shielded plug-in-coil assemblies can be seen to the left of the tuning gang. The 6K8 converter is the tube on the left nearest the panel.

The antenna terminal strip, power-supply plug, headphone jack and 'speaker terminals are mounted on the rear (foreground in this view) of the chassis.

### *Construction*

The receiver is built on an aluminum chassis mounted in a Par-Metal CA-202 cabinet, and a Millen 10035 dial is used for tuning. The chassis is made of 1/16-inch-thick stock, bent into a "U"-channel, and measures 13 inches wide and 7 1/4 inches deep on the top. It is 3 3/8 inches deep at the rear and 1/8 inch less at the front. The rear edge is reinforced with a piece of 3/8-inch square dural rod that is tapped for screws through the bottom of the cabinet, further to add to the strength of the structure when finally assembled. The various components that are common to the front lip of the chassis and the panel are used to tie the two together.

The shield panel used to mount the antenna compensator condenser is also made of 1/16-inch aluminum with a 5/8-inch lip on the side for mounting. Part of the lip must be cut away to clear wires and mounting plates on some sockets, so it is advisable to put in the panel after most of the assembly and wiring have been completed. Flexible couplings and bakelite rod couple the condenser to the panel bushing.

The three tuning condensers are mounted on individual brackets of 1/16-inch aluminum. The brackets measure 2 1/2 inches wide and 1 9/16 high, with 1/2-inch lips. A cover of thin aluminum — not shown in the photographs — slides over the condenser assembly to dress up the top view a bit. The dust cover is not necessary for satisfactory operation of the receiver.

Ceramic sockets are used for the plug-in coils and for the r.f. amplifier, converter and b.f.o. tubes. Mica condensers were used throughout the receiver for by-passing wherever feasible, because they lend themselves well to compact construction. Paper condensers could be used in the i.f. amplifier but they would crowd things a bit more.

In wiring the receiver, small tie-points were used wherever necessary to support the odd ends of resistors and condensers, and rubber grommets were used wherever wires run through the chassis, with the exception of the tuning-condenser leads. The latter leads, being of No. 14 wire, are self-supporting through the 5/16-inch clearance holes and do not require grommets. The same heavy wire was used for the grid and plate leads of the r.f. stage and the plate lead of the oscillator, to reduce the inductance in these leads. The tuning condensers are grounded back at the coil sockets and not above the chassis as might be the tendency. Screen, cathode and plate by-pass condensers are grounded at a single point for any tube wherever possible, although $C_2$ is grounded at the r.f.-coil socket, $C_8$ is grounded at the converter-coil socket, and $C_{13}$ is returned at the oscillator-coil socket. The plate and B+ leads from $T_1$ are brought back to the converter socket through shield braid, and $C_{21}$ is returned to ground at the converter socket.

The b.f.o. pitch condenser, $C_{38}$, is insulated from the chassis and panel by fiber washers, and the rotor is connected back to the tube socket by braid that shields the stator lead. This is done to reduce radiation from the b.f.o. which might get in at the front end of the i.f. amplifier.

The coils are wound on Millen 74001 permeability-tuned coil forms, according to the coil table. Series condensers are mounted inside the forms on all bands except the 80-meter range, where no condenser is required and the tuning condenser is jumped directly to the grid end of the coils. In building the coils, the washers are first drilled for the leads and then cemented to the form with Duco or other cement. The bottom washer is cemented close to the terminal pins, leaving just enough room

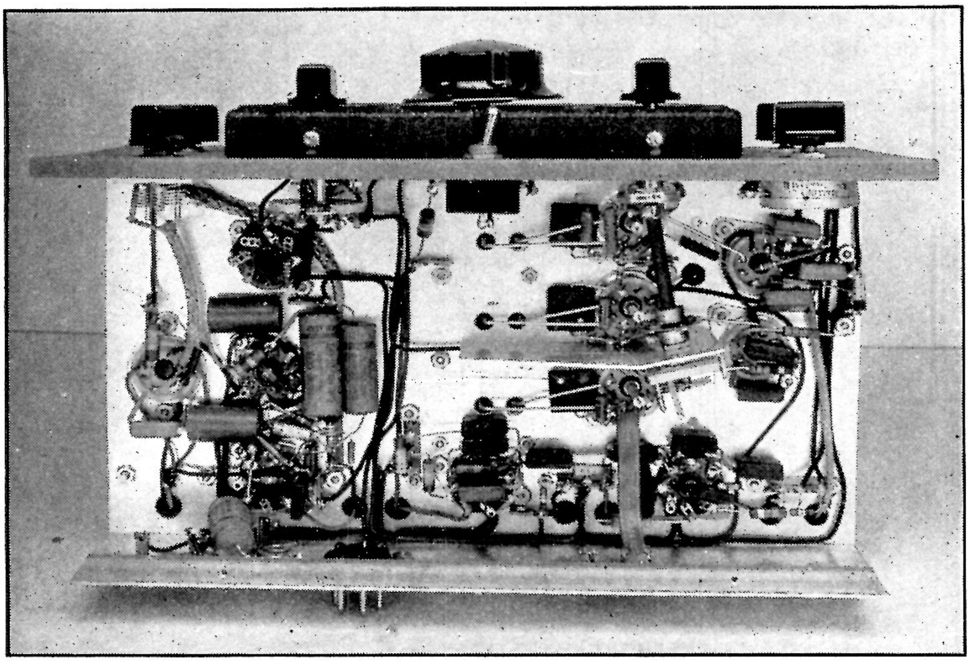


*Fig. 5-49* — The mica by-pass condensers used throughout the r.f. and i.f. stages are grouped around the sockets of their respective tubes. Tie-points are used wherever necessary to support small resistors and condensers. The antenna trimmer condenser is mounted on a bracket which also serves as shielding between the mixer- and r.f.-coil sockets, and it is offset to allow access to the trimmer screws on the coil forms. The plate and B+ leads from the first i.f. transformer, $T_1$, are run in shielded braid, as are the leads from the b.f.o. pitch-control condenser and the volume control.

to get the soldering iron in to fasten the coil ends and to leave room for the series condenser. The large coils, $L_2$, $L_4$ and $L_6$, were wound first in every case, and then a layer of polystyrene Scotch Tape wrapped over the coil, after which the smaller winding was put on and the ends of the windings soldered in place. Since for maximum range of adjustment it is desirable to allow the powdered-iron slug to be fully withdrawn from the coil, keeping the coils at the base end of the form allows the iron slug to travel out at the other end, under which condition the adjusting screw on the slug projects the least. To secure the wires after winding, drops of cement should be placed on them where they feed through the polystyrene washers.

### Alignment

If a signal generator is available, it can be used to align the i.f. amplifier on 455 kc. in the usual manner. If one is not available, the coupling at $C_{C1}$ can be increased to the point where the i.f. stage oscillates readily and the b.f.o. transformer is then tuned until a beat note is heard. The other transformers can then be aligned until the signal is loudest, after which $C_{C1}$ should be decreased until the i.f. oscillates with the regeneration control, $R_{12}$, about 5 degrees from maximum. The trimmers on $T_1$ then should be tuned to require maximum advancing of the regeneration control for oscillation, with a set value of $C_{C1}$. When properly tuned, the oscillation frequency of the i.f. stage and the frequency for maximum gain in the regenerative condition will be the same.

With a set of coils in the front end, set the tuning dial near the high-frequency end and tune in a strong signal or marker with the adjustment screw on the oscillator coil. The converter and r.f. coils can then be peaked, with the antenna compensator set at about half capacitance. Then tune to the other end of the band and see if you have enough bandspread. If the bandspread is inadequate, it means that $C_{14}$ is too large, and it should be reduced by using a smaller size of condenser or a combination that gives slightly less capacitance. The tracking of the converter and r.f. coils can be checked by repeaking the position of the slugs in the coils at the low-frequency end. If the converter- or r.f.-coil tuning slugs have to be advanced farther into the coil (to increase the inductance) it indicates that $C_9$ or $C_1$ should be larger. Tracking by the method described is at best a compromise, although to all intents and purposes the loss from some slight misalignment is completely unimportant. Another method would be to tap the tuning condensers on the coil in the familiar bandspreading manner, but this requires considerable time and patience. However, with the series condensers as used in this receiver, the tuning curve is more crowded at the high-frequency end of a range than at the low, and this would be reduced somewhat by the tapped-coil bandspread.

**COIL DATA FOR THE EIGHT-TUBE SUPERHETERODYNE**

| *Coil* | *3.5 Mc.* | *7 Mc.* | *14 Mc.* | *28 Mc.* |
|---|---|---|---|---|
| $L_1$ | 15 t. | 9 t. | 6 t. | 4 t. |
| $L_2$, $L_4$ | 76 t. | 33 t. | 19 t. | 8 t. |
| $C_1$, $C_9$ | short | 27 μμfd. | 15 μμfd. | 20 μμfd. |
| $L_3$ | 25 t. | 11 t. | 7 t. | 4 t. |
| $L_5$ | 10 t. | 8 t. | 4 t. | 2 t. |
| $L_6$ | 47 t. | 32 t. | 14 t. | 6 t. |
| $C_{14}$ | short | 42 μμfd. | 27 μμfd. | 51 μμfd. |

All coils wound on Millen 74001 forms, close-wound. 3.5-Mc. coils wound with No. 30 enam.; 7-Mc. coils wound with No. 30 d.s.c.; 14- and 28-Mc. coils wound with No. 30 d.s.c. on primaries and ticklers and No. 24 enam. on secondaries. $C_{14}$ for 7-Mc. range made by connecting 27- and 15-μμfd. condensers in parallel. $C_1$, $C_9$ and $C_{14}$, Erie Ceramicons, mounted in coil form.

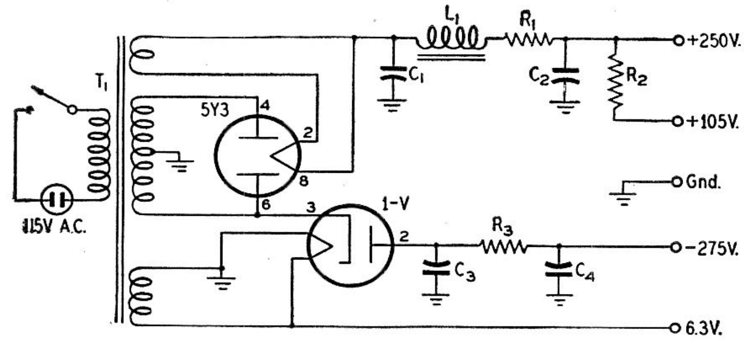



*Fig. 5-50* — Power-supply wiring diagram.

$C_1$, $C_2$ — 16-μfd. 450-volt electrolytic.
$C_3$, $C_4$ — 8-μfd. 450-volt electrolytic.
$R_1$ — 500 ohms, 10 watts, wire-wound.
$R_2$ — 5000 ohms, 10 watts, wire-wound.
$R_3$ — 0.1 megohm, 1 watt, composition.
$L_1$ — 30-henry 110-ma. filter choke (Stancor C-1001).
$T_1$ — 350-0-350 volts, 90 ma.; 5 volts at 3 amp., 6.3 volts at 3.5 amp.

♦

The adjustment of $L_5$ can be made, if deemed necessary, by lifting the cathode end of $R_6$ and inserting a 0–1 milliammeter. If the tickler coil has the right number of turns, the current will be from 0.15 to 0.2 ma., and it won't change appreciably over the band. Although such a grid-current check is a fine point and not really necessary, it is a simple way to determine that the oscillator portion is working, since the cold ends of $L_5$ and $L_6$ are at the same end of the form — the plug end — and this necessitates winding the two coils in opposite directions.

Some trouble may be experienced with oscillation in the r.f. stage at 28 Mc. However, a grounding strap of spring brass, mounted under one of the screws holding the mixer-coil socket to ground the shield when the coil is plugged in, will normally clear up the trouble. Inadequate coupling to the antenna will also let the r.f. stage oscillate under some tuning conditions, and close coupling is highly recommended for stability in this stage and also for best signal response. A 10-ohm resistor from $L_2$ to the grid of the 6SG7 will also do the trick.

It will be found that the over-all gain of the receiver is quite high on the lower-frequency bands, requiring that the r.f. gain be cut down to prevent overloading on strong signals. For c.w. reception, the regeneration control is advanced to the point just below oscillation and the b.f.o. is detuned slightly to give the familiar single-signal effect. For 'phone reception, $S_2$ is switched to "A.V.C." and volume-control adjustments made with the audio control, $R_{26}$. If desired, the regeneration control can be advanced until the i.f. is oscillating weakly, and then a heterodyne will be obtained on weak carriers, making them easy to spot. Strong carriers will pull the i.f. out of oscillation because the developed a.v.c. voltage reduces the gain, and hence a simple form of automatic selectivity control is obtained. If it is considered desirable to reduce the i.f. gain when switched to the "A.V.C." position, the regeneration control can be used for this purpose. The "MAN." position permits manual gain-control operation with the b.f.o. off.

The switch $S_1$ is used for receive-transmit and throws about 40 volts negative on the grid of the first r.f. stage.

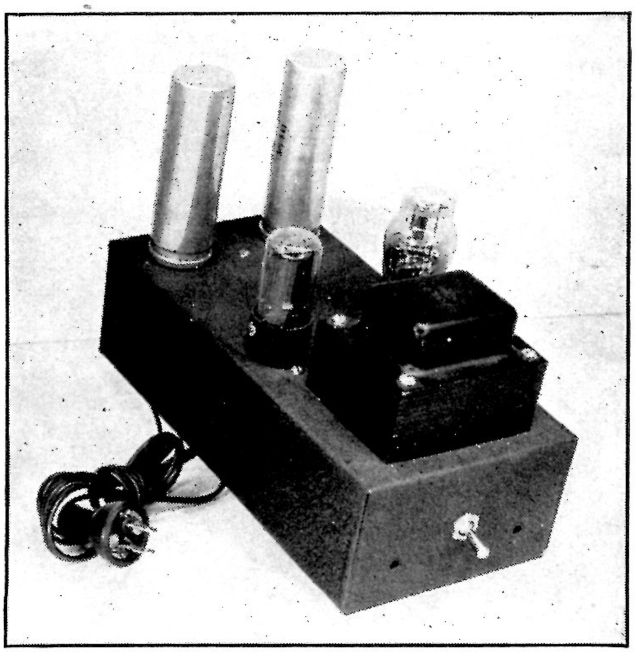


*Fig. 5-51* — Power supply for the eight-tube receiver. Two rectifiers are required because a separate supply is incorporated for gain-control purposes. The filter choke and the negative-supply filter condensers are mounted under the chassis. At the rear of the chassis is the socket for the power cable.

### *Power Supply*

A power supply suitable for the eight-tube receiver is shown in Figs. 5-50 and 5-51. An idea of the parts arrangement can be obtained from Fig. 5-51, although there is nothing critical about this portion of the receiver. If one wants a neat-looking station with no loose power supplies in sight, the power supply can be built into one corner of the loudspeaker cabinet.

## Improving Receiver Sensitivity

Early in this chapter it was pointed out that the sensitivity (signal-to-noise ratio) of a receiver on the higher frequencies above 20 Mc. is dependent upon the bandwidth of the receiver and the noise contributed by the "front end" of the receiver. Neglecting the fact that the image rejection is poor, a receiver with no r.f. stage is generally satisfactory, from a sensitivity point, in the 3.5- and 7-Mc. bands. However, as the frequency is increased and the atmospheric noise becomes less, the advantage of a good "front end" becomes apparent. Hence at 14 Mc. and higher it is worth while to use at least one stage of r.f. amplification ahead of the first detector for best sensitivity as well as image rejection. The multigrid converter tubes have very poor noise figures, and even the best pentodes and triodes are three or four times noisier when used as mixers as they are when used as amplifiers.

If the purpose of an r.f. amplifier is to improve the receiver noise figure at 14 Mc. and higher, a high-$G_m$ pentode or triode should be used. Among the pentodes, the best tubes are the 6AC7, 6AK5 and the 6SG7, in the order named. The 6AK5 takes the lead above 30 Mc. The 6J4, 6J6 and 7F8 are the best of the triodes. For best noise figure, the antenna circuit should be coupled a little heavier than optimum. This condition leads to poor selectivity in the antenna circuit, so it is rather futile to try to combine best sensitivity with maximum selectivity in this circuit.

When a receiver is satisfactory in every respect (stability and selectivity) except sensitivity on 14 and/or 28 Mc., the best solution for the amateur is to add a **preamplifier,** a stage or two of r.f. amplification designed expressly to improve the sensitivity. If image rejection is lacking in the receiver, some selectivity should be built into the preamplifier (it is then called a preselector). If, however, the receiver operation is poor on the higher frequencies but is satisfactory on the lower ones, a "converter" is the best solution.

Some commercial receivers that appear to lack sensitivity on the higher frequencies can be improved simply by tighter coupling to the antenna. Since the receiver manufacturer has no way to predict the type of antenna that will be used, he generally designs the input for some compromise value, usually around 300 or 400 ohms in the high-frequency ranges. If your antenna matches to something far different from this, the receiver effectiveness can be improved by proper matching. This can be accomplished by changing the antenna to the right value (as determined from the receiver instruction book) or by using a simple matching device as described later in this chapter. Overcoupling the input circuit will often improve sensitivity but it will, of course, always reduce the image-rejection contribution of the antenna circuit.

Commercial receivers can also be "hopped up" by substituting a high-$G_m$ tube in the first r.f. stage if one isn't already there. The amateur must be prepared to take the consequences, however, since the stage may oscillate, or not track without some modification. A simpler solution is to add the "hot" r.f. stage ahead of the receiver.

### *Regeneration*

Regeneration in the r.f. stage of a receiver (where only one stage exists) will often improve the sensitivity because the greater gain it provides serves to mask more completely the first-detector noise, and it also provides a measure of automatic matching to the antenna through tighter coupling. However, accurate ganging becomes a problem, because of the increased selectivity of the regenerative r.f. stage, and the receiver almost invariably becomes a two-handed-tuning device. Regeneration should not be overlooked as an expedient, however, and many amateurs have used it with considerable success. High-$G_m$ tubes are the best as regenerative amplifiers, and the feed-back should not be controlled by changing the operating voltages (which should be the same as for the tube used in a high-gain amplifier) but by changing the loading or the feed-back coupling. This is tricky and another reason why regeneration is not too widely used.

### *Gain Control*

In a receiver front end designed for best signal-to-noise ratio, it is advantageous in the reception of weak c.w. signals to eliminate the gain control from the first r.f. stage and allow it to run "wide open" all of the time. If the first stage is controlled along with the i.f. (and other r.f. stages, if any), the signal-to-noise ratio of the receiver will suffer. As the gain is reduced, the $G_m$ of the first tube is reduced, and its noise figure becomes higher. An elaborate receiver might well have separate gain controls for the first r.f. stage and for all i.f. stages.

## A Dual-Triode Preamplifier

The 6J6 preamplifier shown in Figs. 5-52, 5-54 and 5-55 is a simple device that can be used ahead of a receiver on 14 or 28 Mc. to improve the sensitivity. It is designed to be built for one band only and left fixed, although coil dimensions are given for both the 10- and 20-meter bands.

As can be seen from the circuit diagram in Fig. 5-53, the input circuit uses a reactance network, $C_1C_2$, to enable the operator to adjust the input of the preamplifier to whatever impedance his antenna system works into best. Higher impedances require that the ratio of $C_1$ to $C_2$ be made larger. Final peaking is done with $L_1$, and the only trimming done during actual operation is in the output circuit, $L_2C_5$. $C_5$ is made variable and is the only control easily accessible to the operator. This circuit provides some additional selectivity ahead of the receiver for image rejection, but the input circuit is coupled heavily to the antenna and hence adds no selectivity. The antenna circuit

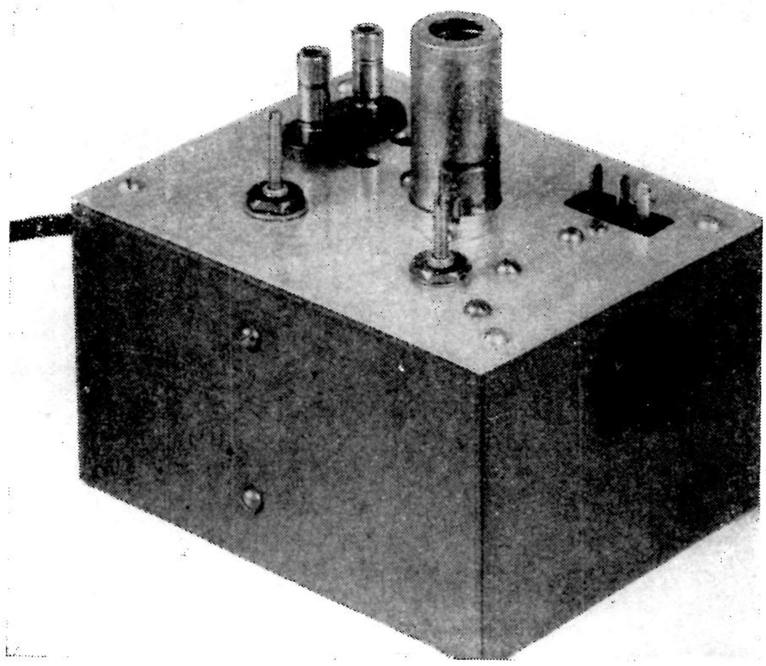


*Fig. 5-52* — A one-tube 6J6 preamplifier for boosting receiver sensitivity on 14 or 28 Mc. Note the two holes in front of the antenna posts through which the antenna-coupling condensers are adjusted.

requires no retuning over the bands. With this type of input matching, an antenna with a "flat" line works best, since there are some impedances presented by tuned lines that are difficult to match with this arrangement.

The first half of the 6J6 acts as a cathode-follower stage, and its output signal is developed across $R_1$. The second half of the tube acts as a grounded-grid r.f. amplifier and hence requires no neutralization. With the output of the preamplifier connected to a receiver but with the input disconnected from the antenna, it is possible to make the amplifier oscillate, but connecting an antenna and tuning the input properly results in a stable amplifier. A regenerative condition in the amplifier will require retuning of the input circuit over an amateur band, and is to be avoided.

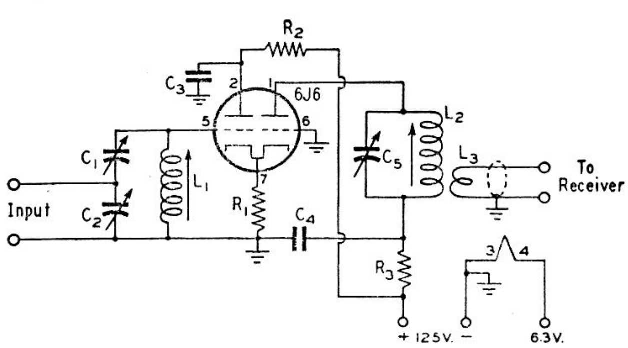



*Fig. 5-53* — Circuit diagram of the 6J6 cathode-coupled preamplifier.

$C_1$, $C_2$ — 3- to 30-μμfd. mica trimmer (National M-SO).
$C_3$, $C_4$ — 0.001-μfd. mica, smallest size.
$C_5$ — 35-μμfd. midget variable (Millen 20035).
$R_1$ — 470 ohms, ½-watt carbon.
$R_2$, $R_3$ — 1500 ohms, ½-watt carbon.
$L_1$ — 28 Mc.: 17 turns No. 22 d.c.c., close-wound.
— 14 Mc.: 34 turns No. 30 d.c.c., close-wound.
$L_2$ — 28 Mc.: 10 turns No. 18 enam., space-wound to fill form.
— 14 Mc.: 20 turns No. 22 d.c.c., close-wound.
$L_3$ — 3 turns No. 18 flexible hook-up wire close-wound over ground end of $L_2$.
Coils are wound on National XR-50 forms.

### *Construction*

The construction of the unit is a simple matter. With the single exception of the interstage shield, all of the components are mounted on the 4 × 5-inch piece of aluminum used to replace one side of the 3 × 4 × 5-inch box. This makes it easy to work on the unit, since all of the construction work can be done with the chassis removed from the box. The two condensers, $C_1$ and $C_2$, are supported by No. 12 tinned wire soldered to the input terminals (a National FWH assembly) and a lug on a small National GS-10 stand-off insulator. The output tuning condenser, $C_5$, is mounted on a small aluminum bracket. All r.f.-circuit grounds are brought to a single lug under one of the 6J6 socket mounting screws. The power leads are brought out through a Jones P-303-AB miniature plug, and the lip of the case must be cut out slightly to clear one bracket of this plug. The output cable, a length of RG-58/U, is secured to a small tie-point which also serves as a tie-point for the ends of $L_3$. The outer conductor of the cable is grounded by the tie-point mounting bracket.

The interstage shield is fastened to one side of the case only, and it is juggled into place after the chassis is fastened down. The interstage shield has a notch to clear the r.f.-circuit ground lug, and another notch holds the output cable in place. A small hole is necessary to pass the lead from $L_1$ to the grid of the tube. This grid lead must be connected after the shield is in place, but it is the only connection that can't be made beforehand. The grounded grid lead, from Pin 6, runs across the socket to the socket center shield, to Pin 3, and to ground.

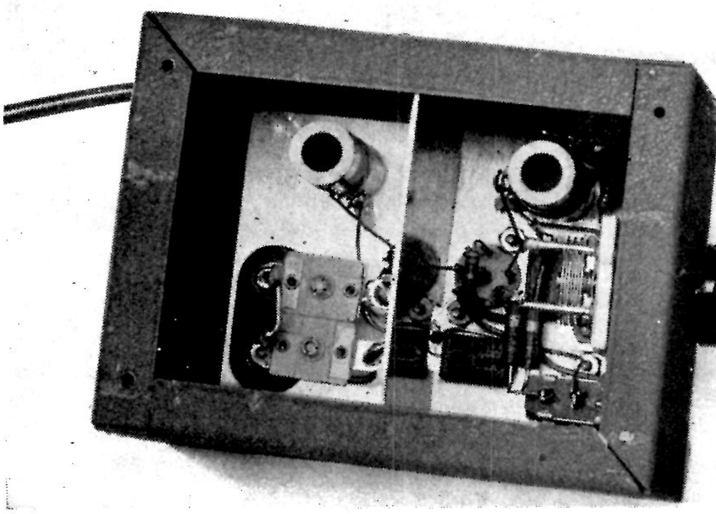


*Fig. 5-54* — A view of the underside of the preamplifier shows the shield partition in place.

necessary to resort to an external tuned circuit link-coupled to the input of the preamplifier. With a reasonably "flat" line, no difficulty should be encountered.

If the input circuit is too lightly loaded by the antenna, the preamplifier may be regenerative. While it will be sensitive in this condition, the regeneration will sharpen the input circuit so much that it will be necessary to retune $L_1$ several times across the band. Since the best noise figure generally obtains when the antenna coupling is slightly more than "critical coupling," there is no advantage in the regenerative condition except possibly the improved image rejection. The disadvantage of having to retune $L_1$ is obvious, however.

### *Adjustment*

The permeability-tuned coils and the adjustable input condensers make adjusting the preamplifier a fairly easy job. Connecting the amplifier to the receiver, $C_5$ is set at about half scale and the core of $L_2$ adjusted for resonance, as indicated by a slight increase in noise in the receiver. Then different ratios of $C_1$ to $C_2$ are tried, resonating the circuit with the slug in $L_1$, until best results are obtained. If $C_1$ is small compared with $C_2$, the loading on the input circuit will be light, and if it is too light the amplifier will oscillate. This is remedied by increasing the capacity of $C_1$ (or decreasing that of $C_2$) and reresonating $L_1$. If the receiving antenna uses a tuned line, some combinations may occur where proper loading cannot be obtained, in which case it may be

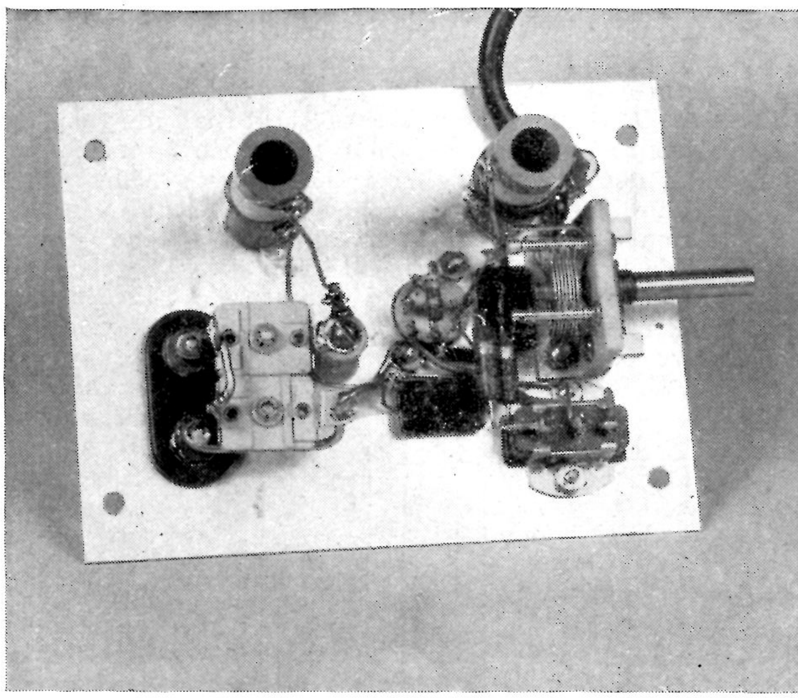


*Fig. 5-55* — The preamplifier chassis removed from the case shows the parts arrangement. All r.f. grounds are made to one common point.

## An Antenna-Coupling Unit for Receiving

It will often be found advantageous on the 14- and 28-Mc. bands to tune (or match) the receiving-antenna feedline to the receiver, in order to get the most out of the antenna. A compact unit for this purpose is shown in Fig. 5-56. The wiring diagram, Fig. 5-57, shows that the unit is a simple pi-section coupler. Through proper selection of condensers and inductances, a match can be obtained over a wide range of values. It can be placed close to the receiver and left connected all of the time, since it will have little or no effect on the lower frequencies. A short length of 300-ohm Twin-Lead is convenient for connecting the coupler to the receiver.

The antenna coupler is built in a 3 × 4 × 5-inch metal cabinet. All of the components except the two pairs of terminals are mounted on one panel. The condensers are mounted off the panel by the spacers furnished with the condensers, and a clearance hole for the shaft prevents any short-circuit to the panel. The

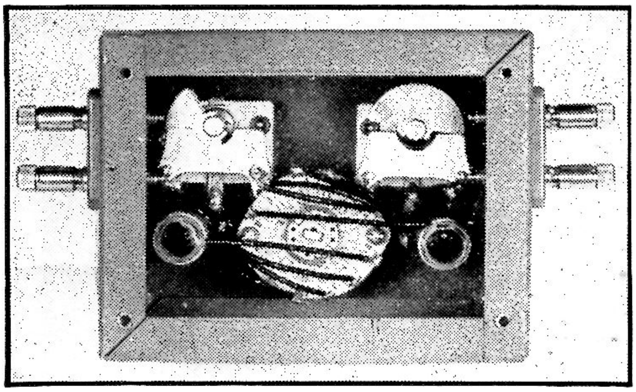


*Fig. 5-56* — Rear view of the antenna-coupling unit. The two coils can be seen directly below the two tuning condensers.

coils, wound on National PRD-2 polystyrene forms, are fastened to the panel with brass screws, and the coils should be wound on the forms as far away as possible from the mounting end. If this still leaves the coil ends within ½ inch of the panel, the forms should be spaced away from the panel by National XP-6 buttons. The switch should be wired so that the switching sequence puts in, in each coil, 0 turns, 2½ turns, 6½ turns, 14½ turns and 30 turns. All of the wiring, with the exception of the input and output terminals, can be done with the panel removed from the box.

The unit is adjusted for maximum signal by switching to different coil positions and adjusting $C_1$ and $C_2$. It will not be necessary to retrim the condensers except when going from one end of a band to the other, and when the unit is not in use, as on 7 and 3.5 Mc., the coils should be switched out of the circuit and the condensers set at minimum.

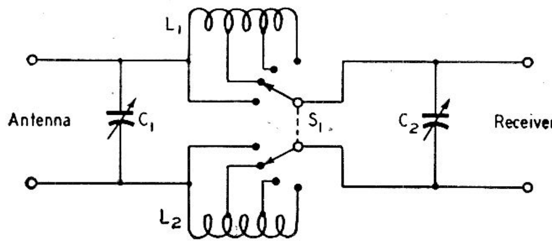



*Fig. 5-57* — Circuit diagram of the coupling unit.
$C_1$, $C_2$ — 100-μμfd. midget variable (Millen 22100).
$L_1$, $L_2$ — 30 turns No. 18 d.c.c. close-wound on ½-inch diameter polystyrene form, tapped at 2½, 6½ and 14½ turns.
$S_1$ — 2-circuit 5-position single-section ceramic wafer switch (Mallory 173C).

## Extending the Tuning Range

As mentioned earlier, when a receiver doesn't cover a particular frequency range, either in fact or in satisfactory performance, a simple solution is to use a **converter.** A converter is another "front end" for the receiver, and it is made to tune the proper range or to give the necessary performance. It works into the receiver at some frequency between 1.6 and 10 Mc. and thus forms with the receiver a "triple-detection" superhet.

There are several different types of converters in vogue at the present time. The commonest type, since it is the oldest, uses a regular tunable oscillator, mixer, and r.f. stages as desired, and works into the receiver at a fixed frequency. A second type uses broad-banded r.f. stages in the r.f. and mixer stages of the converter, and only the oscillator is tuned. Since the frequency the converter works into is high (7 Mc. or more), little or no trouble with images is experienced, despite the broad-band r.f. stages. A third type of converter uses broad-banded r.f. and output stages and a fixed-frequency oscillator (self- or crystal-controlled). The tuning is done with the receiver the converter is connected to. This is an excellent system if the receiver itself is well shielded and has no external pick-up of its own. Many war-surplus receivers fall in this category. A fourth type of converter uses a fixed oscillator with ganged mixer and r.f. stages, and requires two-handed tuning, for the r.f. stages and for the receiver. The r.f. tuning is not critical, however, unless there are many stages.

The broad-banded r.f. stages have the advantage that they can be built with short leads, since no tuning capacitors are required and the unit can be tuned initially by trimming the inductances. They are a little more prone to cross-modulation than the gang-tuned r.f. stages, however, because of the lack of selectivity. The fourth type of converter, although the most difficult to build, is probably the most satisfactory, particularly if a crystal-controlled high-frequency oscillator is used. It not only has the advantage of the best selectivity and protection against images and cross-modulation, but the crystal gives it a stability unobtainable with self-controlled oscillators. Amateurs who specialize in operation on 28 and 50 Mc. often develop good converters for use ahead of conventional communications receivers, and the extra trouble often pays off in outstanding performance for the station.

While converters can extend the operating range of an existing receiver, their greatest advantage probably lies in the opportunity they give for getting the best performance on any one band. By selecting the best tubes and techniques for any particular band, the amateur is assured of top receiver performance. With separate converters for each of several bands, changes can be made in any one without disabling or impairing the receiver performance on another band. The use of converters ahead of the low-frequency receiver is rapidly becoming standard practice on the bands above 14 Mc.

# A One-Tube Converter for 10 and 11 Meters

The 10- and 11-meter converter shown in Figs. 5-58 and 5-60 is a simple unit that can be built in a few hours, for a cost of less than ten dollars. The converter uses a fixed-frequency oscillator and tunable input and output circuits. The fixed oscillator frequency is selected to take advantage of the calibration and band-spread offered by the communications receiver into which the converter works. Because of the light current consumption — 10 to 12 ma. — it is usually possible to operate the converter from the receiver power supply.

The circuit diagram, Fig. 5-59, shows that a Type 6BE6 miniature pentagrid-converter tube is used. The tuning range of the oscillator allows the oscillator to be set 4 to 6 Mc. below the frequency of the signal (input) circuit, and the receiver into which the converter works must be able to cover the range 4–6 Mc.

A Hartley circuit is used in the oscillator portion of the 6BE6. Coil $L_3$ is connected in parallel with condensers $C_2$ and $C_4$, and the frequency of the oscillator is determined by the values of these three components. The frequency of the oscillator must remain fixed after the converter has once been adjusted, and, as a result, stability is an important requirement. This condition is obtained by using a high-$C$ tank circuit, with the 100-$\mu\mu$fd. condenser, $C_4$, providing the major portion of the capacity. The variable condenser, $C_2$, is used as a vernier control for selection of a spot-frequency within the oscillator-frequency range. Feed-back control for the oscillator is obtained by moving the 6BE6 cathode tap on $L_3$. Bias voltage for the oscillator is developed across resistor $R_1$, and $C_6$ is the grid-blocking condenser. Condenser $C_6$ keeps the screen grid at ground r.f. potential, and the dropping resistor, $R_2$, reduces the receiver supply voltage to 100 volts — the value recommended for the 6BE6 screen grid. The exact value for this resistor cannot be suggested at this time because the receiver supply voltage must be known before the resistance can be calculated. However, the resistor will carry about 7 ma., and it will probably have a resistance somewhere between 10,000 and 22,000 ohms.

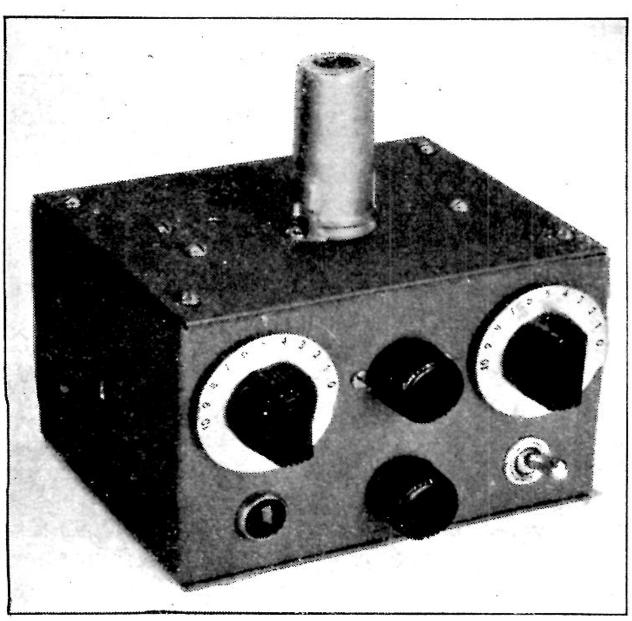


*Fig. 5-58* — A front view of the ten-meter converter. The components and controls on the front wall of the case, from left to right, are as follows: top row, r.f. tuning control, oscillator tuning knob, and i.f.-circuit control; bottom row, dial-light assembly, antenna change-over switch, and filament switch.

The input circuit consists of coils $L_1$ and $L_2$ and condenser $C_1$. The antenna coil, $L_1$, is center-tapped to allow changing from the doublet to a single-wire type of antenna without the necessity for grounding one of the input terminals.

The output circuit uses a parallel tank circuit, $C_3L_4$, an output link, $L_5$, and a decoupling network formed by condenser $C_7$ and resistor $R_3$.

Antenna change-over and stand-by switching is done with the selector switch, $S_{1A\text{-}B\text{-}C\text{-}D}$. When set at one of the two positions, sections $A$ and $B$ will connect the antenna to the converter input coil while section $C$ will connect the output link, $L_5$, to the output jack, $J_1$. At the same time, section $D$ will complete the high-voltage connection between the input jack, $J_2$, and the plate and screen circuits. When the selector switch is thrown to the second position the antenna will be connected to the receiver and plate and screen voltage will be removed from the 6BE6. This action of disconnecting the antenna and high voltage during transmission periods prevents converter-tube overload and damage to the input coils that might be caused by the strong transmitter signal. A toggle switch, $S_2$, is used as the heater on-off control.

### Construction

A utility box, measuring 3 × 4 × 5 inches, serves as the chassis and cabinet for the converter. The variable condensers, switches, pilot-light assembly and jacks should be mounted on the front and rear walls as shown in Fig. 5-60. The condensers are mounted in line on the front wall, with the shafts centered exactly 1 inch down from the top of the box. The pilot-light assembly and switches are mounted below the condensers and, in each case, are centered 11/16 inch above the bottom edge of the case.

The tube socket is mounted on the top cover of the utility box and is located 1⅝ inches from the front edge. Holes to pass the coil-form mounting screws are drilled on either side of the tube socket; these holes are ⅞ inch in from the ends of the cover. A tie-point strip is mounted to the rear and right of the tube socket.

Wiring of the unit will be greatly simplified if the wiring is divided into two jobs. The first half includes the wiring associated with the parts mounted on the case walls. This includes the jumper connections on the selector switch and the connections between this switch and the input and output jacks and terminals. Amphenol 300-ohm Twin-Lead is used between the antenna terminals and switch sections *A* and *B* but ordinary hook-up wire, twisted to form a low-impedance line, can be used. The lead from the switch to the output jack should be placed up against the rolled-over edge of the box, to obtain as much shielding as possible. The pilot light and toggle switch can be wired at this time, and a 6-inch lead should be left hanging from the switch side of the pilot light so that the tube filament circuit can be completed when the unit is assembled. The plate by-pass condenser, $C_7$, can be connected between the rotor terminals of the i.f. and oscillator condensers, and the decoupling resistor, $R_3$, can be mounted between $C_3$ and section *D* of the selector switch.

The input and output coils should now be wound on the forms suggested in the parts list. Holes, separated by the recommended distance, are drilled straight through the forms, and the ends of the windings are pulled through these holes and cemented in place. The antenna coil is wound directly below the grounded end of the grid coil, $L_2$, and the output link is wound over the cold end of $L_4$. It will not be possible to pass the top end of the output link, $L_5$, through a hole because $L_4$ is directly below this winding and, as a result, the free end of the link should be held in place with Scotch Tape or cement until the coil is mounted and wired. The oscillator coil, $L_3$, can be wound on a dowel or tube of 5/8-inch diameter; the coil will expand to a 3/4-inch diameter when it has been slipped off the form.

The tube socket, tie-point strip and coils are now mounted in place on the box cover. Soldering lugs are placed under each of the tube-socket mounting nuts. The oscillator coil is soldered between one of the lugs and one of the tie-point terminals. Condenser $C_4$ is connected across the ends of $L_3$, and the grid resistor, grid-blocking condenser and screen by-pass are wired into the circuit. If the receiver supply voltage is known at this time it is possible to calculate the correct value for the screen-dropping resistor, and the resistor can be mounted on the tie-point strip. The resistor value is obtained from the equation

$$R \text{ (ohms)} = \frac{\text{supply voltage} - 100}{0.0073}$$

Example: Supply voltage 250; the resistor value is $\frac{250 - 100}{0.0073}$ = 20,500 ohms. Anything within 10% of this figure would be satisfactory.

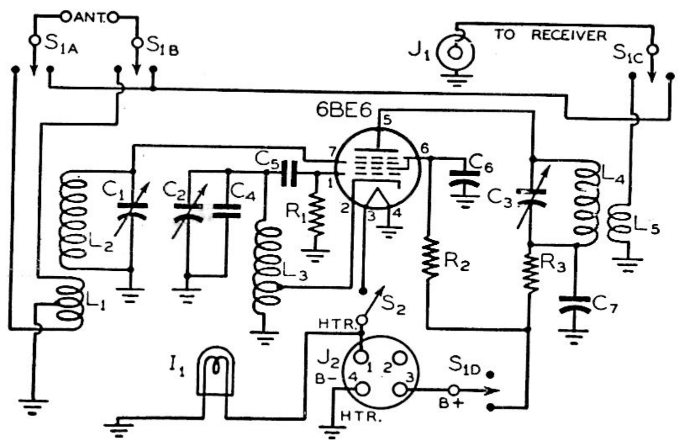



*Fig. 5-59* — Circuit diagram of the low-cost ten-meter converter.

$C_1$, $C_2$ — 15-$\mu\mu$fd. variable (Millen 20015).
$C_3$ — 75-$\mu\mu$fd. variable (Millen 20075).
$C_4$ — 100-$\mu\mu$fd. silver mica.
$C_5$ — 47-$\mu\mu$fd. mica.
$C_6$, $C_7$ — 0.01-$\mu$fd. paper.
$R_1$ — 22,000 ohms, 1/2 watt.
$R_2$ — Screen resistor; see text.
$R_3$ — 1,000 ohms, 1 watt.
$L_1$ — 5 turns No. 22 d.c.c., 9/16-inch diam., close-wound and center-tapped.
$L_2$ — 13 turns No. 22 d.c.c., 9/16-inch diam., 7/8 inch long.
$L_3$ — 6 turns No. 14 tinned, 3/4-inch diam., 3/4 inch long. Cathode tap 1 3/4 inches from cold end.
$L_4$ — 78 turns No. 32 d.c.c., 9/16-inch diam., 1 1/4 inches long.
$L_5$ — 10 turns No. 32 d.c.c., close-wound.
Coils $L_1$, $L_2$, $L_4$ and $L_5$ wound on National Type PRE-3 forms.
$I_1$ — 6.3-volt pilot-lamp-and-socket assembly.
$J_1$ — Panel-mounting female socket (Jones S-101).
$J_2$ — Panel-mounting male socket (Amphenol 86-CP4).
$S_{1A\text{-}B\text{-}C\text{-}D}$ — 4-pole double-throw selector switch (Mallory 3242J).
$S_2$ — S.p.s.t. toggle switch.

An 8-inch lead should be connected to the high-voltage end of the screen-resistor mounting terminal; the free end of this lead will be connected to the selector switch during the final stage of the wiring. The grounded ends of $L_2$ and $L_5$, and the center-tap of $L_1$, are connected to the grounded soldering lugs, and 2-inch tinned wire leads are connected to the following points: one to each soldering lug and one each to Pins 5 and 7 of the tube socket. A connection is now made between the cathode prong of the tube socket and the tap on coil $L_3$, and a connection is made between the screen dropping-resistor, $R_2$, and the screen-grid pin (No. 6) of the socket.

The top cover is now attached to the case and the wiring completed. Few connections remain to be made and, in each case, wires are already provided and soldered in place at one end. After the wiring has been completed it should be given a final check before the testing is started, paying special attention to the heater and plate circuits. Extreme care must be taken while soldering leads that terminate at the ends of $L_1$, $L_2$, $L_4$ and $L_5$. These coils are wound on polystyrene forms which melt and lose shape if subjected to intense heat for any length of time.

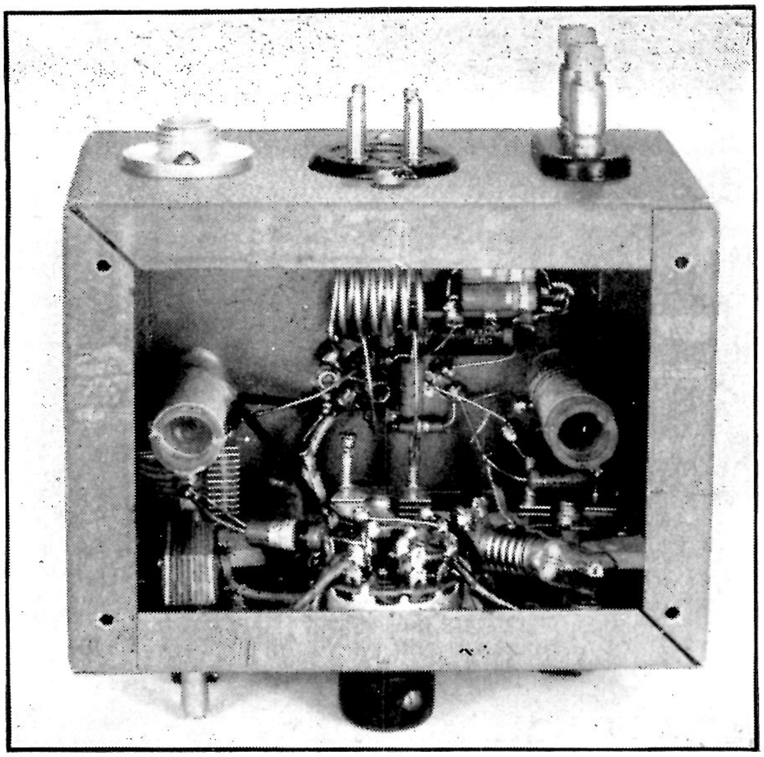


*Fig. 5-60* — An inside view of the ten-meter converter. The r.f. and i.f. coils are at the right and left ends of the box, respectively. The oscillator coil may be seen to the rear of the tube socket. This view also shows the arrangement of the components mounted on the front wall of the case and the location of the input and output connectors which are mounted on the rear wall. The plate by-pass condenser is in a vertical position between the oscillator and the i.f. tuning condensers.

### *Testing*

Adjustment of the converter is convenient if a test oscillator is available, but it is not necessary. Power for the unit can be obtained from the receiver with which the converter is to be used, or from a separate power supply. The converter requires 6.3 volts at 0.45 ampere for the heater and pilot lamp, and 200 to 250 volts d.c. at 10 to 12 ma. for the plate and screen.

After the power supply has been connected, it is advisable to check the screen and plate voltages with a voltmeter. It may be necessary to change the screen-dropping resistor, $R_2$, if the voltage at Pin 6 isn't between 90 and 110.

A coaxial or shielded cable should be connected from the converter output jack to the receiver input terminals. The cable must be shielded to avoid the pick-up of unwanted signals. If your transmitter uses VFO, set it to 28 Mc. and your receiver to 4 Mc. If you don't have VFO but use crystal control, set the receiver to your crystal frequency minus 24 Mc. If, for example, your crystal gives a harmonic at 28,650 kc., set the receiver to 4650 kc. The converter oscillator condenser, $C_2$, should now be adjusted until the VFO or crystal harmonic can be heard. If the harmonic can't be heard, run a wire from the antenna posts of the converter close to the transmitter oscillator. If the signal from the transmitter oscillator is too loud, reduce the length of the wire or remove it entirely. When the signal is reasonably weak in the converter, the input and output tuning capacitors, $C_1$ and $C_3$, can be tuned to make sure that the coils don't need trimming to bring the tuning ranges within the bands.

Once the converter has been carefully set up on a known frequency within the 10- or 11-meter bands, $C_2$ is left fixed and the tuning is done with the receiver. The frequency of the incoming signal can be read directly from the receiver, by adding 24 to the receiver frequency in Mc. For example, a 28-Mc. signal will tune at 4 Mc., and a 29.250-Mc. signal will fall at 5.250 Mc. When tuning the 11-meter band, the setting of $C_2$ is changed so that a signal frequency of 27 Mc. corresponds to 4.0 Mc. on the receiver.

The converter, when properly aligned and working into an average receiver, gives a signal-to-noise ratio of 10 to 1 with an input signal of about 10 microvolts. In operation, $C_1$ and $C_3$ need not be touched over a tuning range of about 150 or 200 kc. on the receiver. Therefore, these controls should be touched up at intervals if the entire 10-meter band is being combed.

## A Bandpass Converter for 14, 28, and 50 Mc.

The converter shown in Figs. 5-62, 5-63, 5-64 and 5-65 will give reception in the 14-, 28- and 50-Mc. bands with any receiver capable of tuning to 7.3 Mc. To simplify construction, the r.f. stages are fixed-tuned and only the local oscillator is tuned when running across a band. The bandwidth of the r.f. stages is sufficient to accept any signal over an amateur band without noticeable attenuation. The broad-banding is obtained by loading the circuits with resistors to reduce the $Q$, using a minimum of capacity for the same reason, and then "staggering" the circuits; i.e., tuning them to slightly different frequencies so that the resultant passband is broad and nearly flat within the required range. The input circuit, from the antenna, must be broad, and this can only be obtained by heavy coupling to the antenna. This condition coincides with the condition for best signal transfer.

As can be seen from the wiring diagram in Fig. 5-61, the only tuning controls in the r.f.

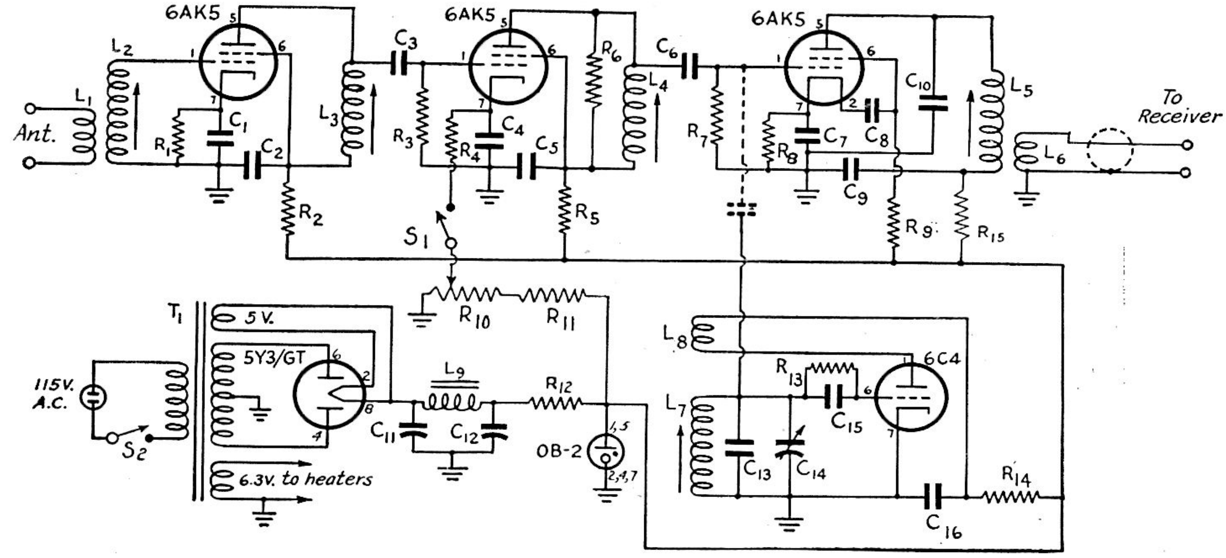



*Fig. 5-61* — Circuit diagram of the bandpass converter.

$C_1$, $C_2$, $C_4$, $C_5$, $C_7$, $C_8$, $C_{16}$ — 0.001-$\mu$fd. postage-stamp mica.
$C_3$, $C_6$ — 100-$\mu\mu$fd. postage-stamp mica.
$C_9$ — 0.01-$\mu$fd. mica.
$C_{10}$, $C_{15}$ — 51-$\mu\mu$fd. ceramic (Erie N150).
$C_{11}$, $C_{12}$ — 16-$\mu$fd. 450-volt electrolytic.
$C_{13}$ — 27-$\mu\mu$fd. ceramic (Erie N150) across $C_{14}$ plus additional capacity mounted in $L_7L_8$ form. See coil table.
$C_{14}$ — 11-$\mu\mu$fd. midget variable (Hammarlund HF-15 with one stator plate removed).
$R_1$, $R_4$ — 180 ohms.
$R_2$, $R_5$, $R_{14}$, $R_{15}$ — 270 ohms.
$R_3$, $R_6$, $R_8$ — 6800 ohms.
$R_7$ — 1.5 megohms.
$R_9$ — 1.0 megohm.
$R_{10}$ — 2000-ohm potentiometer, wire-wound.
$R_{11}$ — 10,000 ohms, 2 watts.
$R_{12}$ — 1250 ohms, 10 watts, wire-wound.
$R_{13}$ — 51,000 ohms, 1 watt.

All resistors ½ watt unless otherwise specified.

$L_1$–$L_8$ — See coil table.
$L_9$ — 8-henry 50-ma. filter choke (Stancor C-1279).
$S_1$ — S.p.s.t. rotary switch.
$S_2$ — S.p.s.t. switch, mounted on $R_{10}$.
$T_1$ — 300–0–300 volts, 50-ma. power transformer, with 5- and 6.3-volt windings (UTC R-6).

stages are the powdered-iron slugs of the coils. These are used to resonate the coils with the circuit capacities to the signal frequency. The loading resistors, $R_3$ and $R_6$, are used to broaden the circuits. The plate and screen voltages are the same on each r.f. amplifier tube, to reduce the number of by-pass condensers, and filter resistors are used to prevent over-all feed-back through the common power lead. Another possible source of over-all feed-back is the heater circuit, and in this converter the "hot" heater lead to the input stage was run in shield braid to reduce the possibility of feed-back.

The oscillator is a straight plate-tickler type using a 6C4, and it is coupled to the mixer through a capacity shown as dashed lines in the diagram. Actually the coupling capacitor consists of a short length of wire near the grid of the mixer tube.

The output frequency is 7.3 Mc. approximately, and this is the frequency to which $C_{10}L_5$ is tuned. If a frequency slightly below 7.0 Mc. is used, there is a possibility that the fourth harmonic of the receiver high-frequency oscillator will find its way into the converter when operating in the 28-Mc. band, resulting in a constant signal that has only nuisance value. A low-impedance shielded

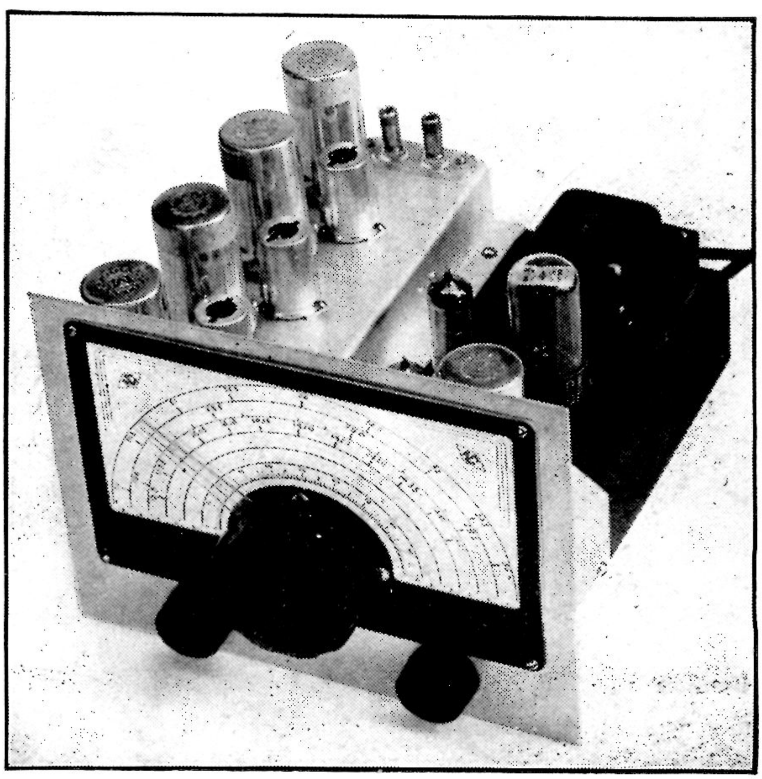


*Fig. 5-62* — A 28-Mc. converter that uses fixed-tuned r.f. stages and thus eliminates the ganging problem. The knob at left is for the "send-receive" switch, and the right-hand knob is for gain control.

**COIL DATA FOR THE BANDPASS CONVERTER**

| *Coil* | *14 Mc.* | *28 Mc.* | *50 Mc.* |
|---|---|---|---|
| $L_1$ | 13 t. No. 26 d.c.c. | 8 t. No. 26 d.c.c. | 5 t. No. 26 d.c.c. |
| $L_2$ | 35 t. No. 24 d.c.c. | 23 t. No. 24 d.c.c. | 8 t. No. 24 d.c.c. spaced wire diam. |
| $L_3$, $L_4$ | 25 t. No. 24 d.c.c. | 8½ t. No. 24 d.c.c. | 5 t. No. 24 d.c.c. spaced twice wire diam. |
| $L_5$ | 37 t. No. 26 enam. | Same | Same |
| $L_6$ | 9 t. No. 26 enam. | Same | Same |
| $L_7$ | 4 t. No. 24 d.c.c. (spaced to occupy ¼ inch) | 7 t. No. 20 enam. | 2 t. No. 24 d.c.c. spaced wire diam. |
| $L_8$ | 3 t. No. 26 d.c.c. | 3 t. No. 26 d.c.c. | 2 t. No. 24 d.c.c. |
| $C_{13}$ | 150 $\mu\mu$fd. | 27 $\mu\mu$fd. | 22 $\mu\mu$fd. |

$L_1$ wound over ground end of $L_2$, tape insulation. $L_8$ spaced from $L_7$ by washer thickness. All coils close-wound unless otherwise specified. All coils wound on Millen 74001 permeability-tuned forms.

line feeds the 7.3-Mc. output into the communications receiver. The communications receiver furnishes the necessary selectivity.

The cathode bias of the second r.f. amplifier is varied by the gain control, $R_{10}$, to avoid blocking by strong signals. The send-receive switch, $S_1$, is used to turn off the converter during transmission periods. The power switch, $S_2$, is mounted on the gain control and is used to turn off the power to the converter.

The power supply is regulated, using the miniature equivalent of the VR-105, and the stabilized 105 volts is fed to all stages.

### *Construction*

The r.f. stages and mixer are built as a separate unit on a strip of aluminum, to furnish a chassis in which the grounds are more certain than they would be on a black-crackled steel chassis, and it also makes a well-shielded amplifier when mounted on the steel chassis.

The steel chassis is a standard 7 × 11 × 2-inch affair. A panel is used to support the National ACN dial, and to reduce metal work on the steel chassis the panel is supported away from the chassis by an aluminum bracket on one side and by two of the screws that fasten the dial to the panel. Holes in the chassis allow access to the tuning slugs of the r.f. coils.

The tuning condenser is mounted on a small aluminum bracket fastened to the chassis by two screws and to the condenser by the shaft bushing. This results in a rigid mount that contributes considerably to the mechanical stability of the oscillator.

The construction of the aluminum channel is apparent from Fig. 5-64. It is 3 inches wide and 1¼ inches high, and is bolted to the side of the steel chassis and to the top. A small strip of bakelite, supported away from the side by screws and small spacers, is used to support the power-supply end of the filter resistors, $R_2$, $R_5$ and $R_{15}$. The ends are fed through small holes in the bakelite and then wrapped around the strip before being soldered together.

In the heater circuits of the miniature tubes, Pin 4 is grounded to a lug under the nut fastening the socket, and Pin 3 is the "hot" heater lead. In the case of the input 6AK5, the hot heater lead was led back in shield braid, and the braid was grounded at the lug grounding Pin 4, and to lugs at two other points along the way. These latter lugs are under the nuts fastening the sockets for $L_3$ and the output coil, $L_5L_6$.

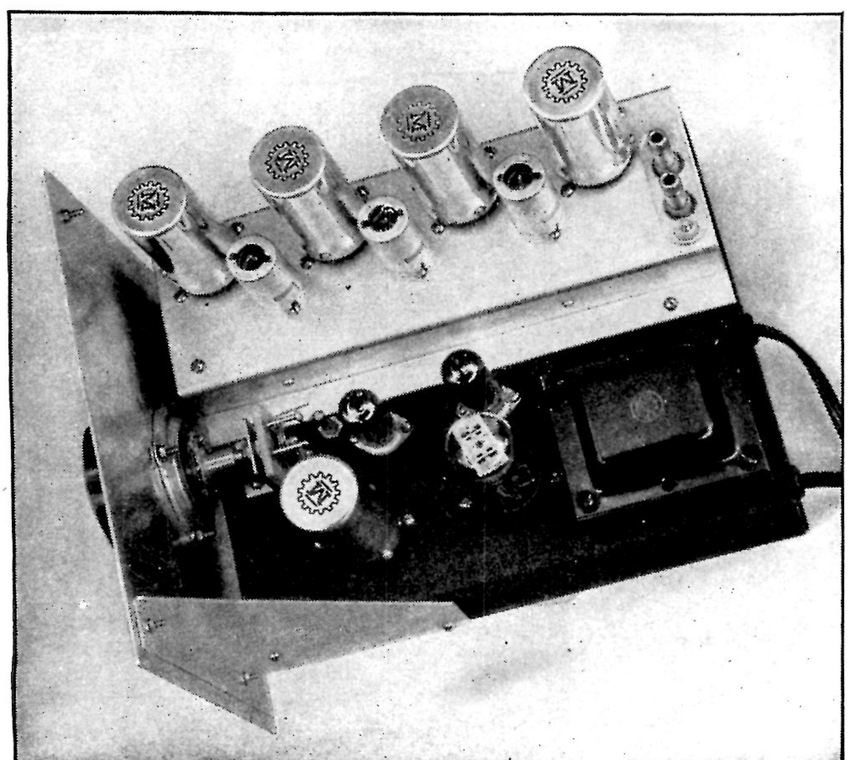


*Fig. 5-63* — Another view of the converter showing the r.f. sub-chassis. Note the bracket on the tuning condenser, used to avoid back-lash.

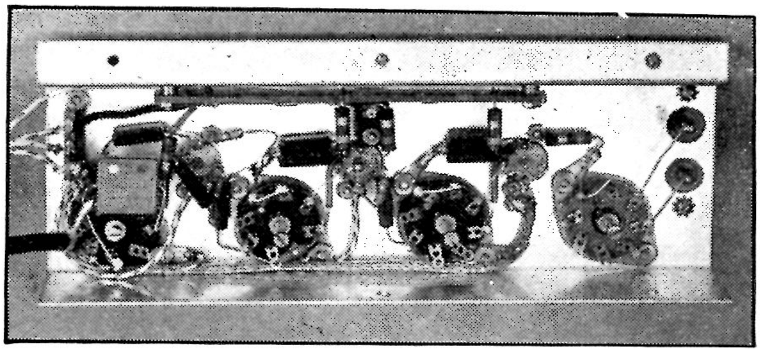


*Fig 5-64* — The straightforward arrangement of the r.f. components is shown in this view of the subchassis. The straight side is screwed to the side of the chassis.

♦

The cathode and screen/plate bypass condensers are grounded to lugs under nuts holding the sockets of their respective plate coils. Since it doesn't matter where the cathode resistors are grounded, they are returned to lugs under the coil sockets ahead of them. Pins 1 and 2 of the coil sockets are grounded to the lugs just mentioned, the No. 3 pins of the coil sockets for $L_3$, $L_4$ and $L_5$ go to the plates of their respective tubes, and the No. 4 pins of the same sockets are connected to the screen pins on the tube sockets. The grid condensers, $C_3$ and $C_6$, are tied from Pin 7 on the coil sockets to the grid pins on the tube sockets.

The oscillator and power-supply wiring on the steel chassis is conventional, with the exception of the oscillator coupling condenser. A small National TPB bushing is mounted on the chassis where it will be parallel to the lead on the grid side of $R_7$. This bushing is connected to the stator of $C_{14}$ and the "hot" side of $L_7$ by a heavy wire, and coupling is obtained by the capacity between this bushing and the grid lead of the mixer stage. The output cable from $L_6$ is a length of RG-59/U 70-ohm cable. If one of the free points on the 0B-2 voltage-regulator tube socket is used as a tie-point for $C_{12}$ and $L_9$, as was done in this case, be sure to clip off the pin on the tube. If this isn't done, a discharge will be obtained inside the tube, since the free pin projects inside the tube envelope and acts as an anode.

The coils for the converter are wound on Millen 74001 tuned plug-in coil forms. The coils are started on the form about ⅛ inch above the lower limit of travel of the iron slug. In the case of $L_3$ and $L_4$, one end of the winding is connected to Pin 4 and the other to Pin 7. A jumper is then run from Pin 7 to Pin 3. This jumper has the effect of tapping down the plate on the coil, since the jumper has some reactance at these frequencies. In the case of the oscillator coil, the padding condenser, $C_{13}$, is mounted inside the coil, although it could be mounted on the coil socket. The tickler, $L_8$, is wound on the form away from the slug end. The mixer output capacitor, $C_{10}$, is mounted on the socket. All coils are securely fastened with coil dope, and this is particularly important in the case of the oscillator coil assembly, to insure long-time stability.

### *Alignment*

After the wiring has been completed and checked, the oscillator should be checked first. Put a voltmeter across $R_{14}$ and see if the voltage increases slightly when the grid of the oscillator tube is touched. If it does, it shows that the circuit is oscillating, and the coil can be tuned to frequency with the iron slug.

Couple the output of the converter to a communications receiver on 7.3 Mc. and adjust the slug of $L_5$ for maximum noise in the receiver, with power to the converter and the converter gain control at minimum. Some kind of signal will be needed with which to establish the oscillator frequency accurately, and this signal can be a harmonic from the station transmitter or a test generator. For 28-Mc. alignment, set the signal source at about 28.5 Mc. and the tuning dial at 35 and adjust the slug on the oscillator coil until the signal is heard. Short the input of the receiver with a carbon resistor equal in value to the impedance of the antenna line. Having established the tuning range — and checking it at other points if available — peak $L_2$, $L_3$ and $L_4$ on noise. Tuning across the band, the output noise should peak near the center of the range and

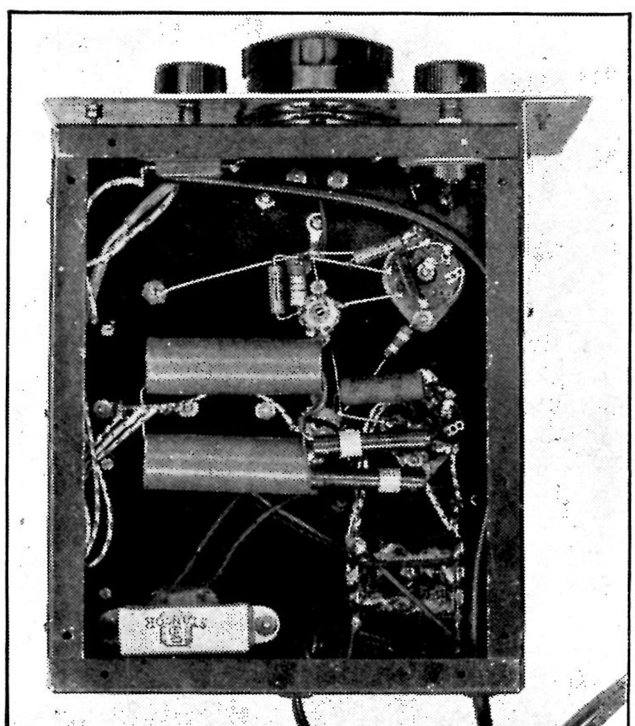


*Fig. 5-65* — A view underneath the chassis shows the polystyrene bushing used to couple from the oscillator to the mixer. The panel is mounted away from the chassis to simplify mounting of the dial. The tuning screws of the r.f. coil project through holes in the chassis.

fall off slightly at either end. By increasing the inductance of $L_4$ — running the slug in — and decreasing the inductance of $L_3$, it will be possible to get practically uniform noise output over the entire range. It will be found that $L_2$ tunes very broadly when loaded by the resistor or the antenna, and its resonance should be checked with this load disconnected, to make certain that the coil can be made to tune through resonance. A sharp increase in the noise will serve as an indication, and it may be found necessary to retard the gain control for this test, to prevent oscillation in the r.f. stages.

If any queer burbles or sudden peaks of noise are encountered, it indicates regeneration in the r.f. stages. If this is encountered, the r.f. stages can be worked on while removed from the chassis, since there will be enough stray oscillator output to the mixer to receive signals, and the various plate- and heater-supply leads can be investigated with a 0.001-$\mu$fd. mica condenser until the source of feed-back is found. Poor grounds can also give trouble.

Under normal conditions, the gain of the communications receiver following the converter will have to be reduced considerably, since the gain of the converter runs around 40 db. It will be found to require very little antenna for normal pick-up, but in order to give it every break it should be used with the best antenna available. Some experiment with the input coupling may be necessary if a tuned antenna is used, but this might be only a tuned circuit with a link line running to the converter input.

# Tuning a Receiver

### *C.W. Reception*

For making code signals audible, the beat oscillator should be set to a frequency slightly different from the intermediate frequency. To adjust the beat-oscillator frequency, first tune in a moderately-weak but steady carrier with the beat oscillator turned off. Adjust the receiver tuning for maximum signal strength, as indicated by maximum hiss. Then turn on the beat oscillator and adjust its frequency (leaving the receiver tuning unchanged) to give a suitable beat-note. The beat oscillator need not subsequently be touched, except for occasional checking to make certain the frequency has not drifted from the initial setting. The b.f.o. may be set on either the high- or low-frequency side of zero beat.

The use of a.v.c. is not generally satisfactory in c.w. reception, except in receivers expressly designed for the purpose, because the rectified beat-oscillator voltage in the second detector circuit also operates the a.v.c. circuit. This gives a constant reduction in gain and prevents utilization of the full sensitivity of the receiver. Hence the gain should be manually adjusted to give suitable audio-frequency output.

To avoid overloading in the i.f. circuits, it is usually better to control the i.f. and r.f. gain and keep the audio gain at a fixed value than to use the a.f. gain control as a volume control and leave the r.f. gain fixed at its highest level, except when there are few loud signals on the band and a low noise level.

### *Tuning with the Crystal Filter*

If the receiver is equipped with a crystal filter the tuning instructions in the preceding paragraph still apply, but more care must be used both in the initial adjustment of the beat oscillator and in tuning. The beat oscillator is set as described above, but with the crystal filter in operation and adjusted to its sharpest position, if variable selectivity is available. The initial adjustment should be made with the phasing control in the intermediate position. After it is completed, the beat oscillator should be left set and the receiver tuned to the other side of zero beat (audio-frequency image) on the same carrier to give a beat-note of the same tone. This beat will be considerably weaker than the first, and may be "phased out" almost completely by careful adjustment of the phasing control. This is the adjustment for normal operation; it will be found that one side of zero beat has practically disappeared, leaving maximum response on the desired side.

An interfering signal having a beat-note differing from that of the a.f. image can be similarly phased out, provided its carrier frequency is not too near the desired carrier.

Depending upon the filter design, maximum selectivity may cause the dots and dashes to lengthen out so that they seem to "run together." It must be emphasized that, to realize the benefits of the crystal filter in reducing interference, it is necessary to do *all* tuning with it in the circuit. Its selectivity is so high that it is often impossible to find the desired station quickly, should the filter be switched in only when interference is present.

### *'Phone Reception*

In reception of 'phone signals, the normal procedure is to set the r.f. and i.f. gain at maximum, switch on the a.v.c., and use the audio gain control for setting the volume. This insures maximum effectiveness of the a.v.c. system in compensating for fading and maintaining constant audio output on either strong or weak signals. On occasion a strong signal close to the frequency of a weaker desired station may take control of the a.v.c., in which case the weaker station will practically disappear because of the reduced gain. In this case better reception may result if the a.v.c. is switched off, using the manual r.f. gain control to set the gain at a point that prevents "blocking" by the stronger signal.

A crystal filter will do much toward reducing interference in 'phone reception. Although the high selectivity cuts sidebands and thereby reduces the audio output, especially at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility even though the "quality" of the transmission may suffer. As in the case of c.w. reception, it is advisable to do all tuning with the filter in the circuit. Variable-selectivity filters permit a choice of selectivity to suit interference conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat-note equal to the frequency difference. Such a heterodyne can be reduced by adjustment of the phasing control in the crystal filter. It cannot be prevented in a "straight" superheterodyne having no crystal filter.

A tone control often will be of help in reducing the effects of high-pitched heterodynes, sideband splatter and noise, by cutting off the higher audio frequencies. This, like sideband cutting with high selectivity, causes some reduction in naturalness.

### *Spurious Responses*

Spurious responses can be recognized without a great deal of difficulty. Often it is possible to identify an image by the nature of the transmitting station, if the frequency assignments applying to the frequency to which the receiver is tuned are known. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat-note with the desired signal and is actually on the same frequency, the beat-note will not change as the receiver is tuned through the signal; but if the interfering signal is an image, the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be turned off for this test. Using a crystal filter with the beat oscillator on, an image will peak on the side of zero beat opposite that on which the desired signal peaks.

Harmonic response can be recognized by the "tuning rate," or movement of the tuning dial required to give a specified change in beat-note. Signals getting into the i.f. via high-frequency oscillator harmonics tune more rapidly (less dial movement) through a given change in beat-note than do signals received by normal means.

Harmonics of the beat oscillator can be recognized by the tuning rate of the beat-oscillator pitch control. A smaller movement of the control will suffice for a given change in beat-note than is necessary with legitimate signals.

## Building a Receiver

In building a receiver, as in most other pieces of radio gear, following a few basic laws during the construction will more often than not result in a reliable job that won't have a lot of "bugs" which require cleaning out before the receiver will perform properly. The more complex a circuit is, and the more gain it has, the better are the chances for trouble, but a large number of tubes doesn't necessarily indicate that a receiver will be difficult to make work properly.

### *Feed-Back*

Probably the greatest source of trouble in receivers is from undesirable feed-back. Feed-back giving rise to regeneration and oscillation can occur in any part of the receiver that has gain: the r.f., i.f. and audio amplifiers. It can occur in a single stage or it may appear as an over-all feed-back through several stages that are on the same frequency. To avoid feed-back in a single stage, the output must be isolated from the input in every way possible, with the vacuum tube furnishing the only coupling between the two circuits. For example, an oscillation can be obtained in a r.f. or i.f. stage if there is any undue capacitive or inductive coupling between output and input circuits, if there is too high an impedance between cathode and ground or screen and ground, or if there is any appreciable impedance through which the grid and plate currents can flow in common. This simply means good shielding of coils and condensers in r.f. and i.f. circuits, the use of good by-pass condensers (mica at 14 Mc. and higher, and with short leads), and returning all by-pass condensers (grid, cathode, plate and screen) with short leads to one spot on the chassis. If single-ended tubes are used, the screen or cathode by-pass condenser should be mounted across the socket, to serve as a shield between grid and plate pins. Less care is required as the frequency is lowered, but in high-impedance circuits like those found in high-gain audio amplifiers, it is sometimes necessary to shield grid and plate leads and to be careful not to run them close together. If this precaution is not observed, serious audio feed-back may result.

To avoid over-all feed-back in a multistage amplifier, strict attention must be paid to avoid running any part of the output circuit back near the input circuit without first filtering it carefully. Since the signal-carrying parts of the circuit (the "hot" grid and plate leads) can't be filtered, the best design for any multistage amplifier is a straight line, to keep the output as far away from the input as possible. For example, an r.f. amplifier might run along a chassis in a straight line, run into a mixer where the frequency is changed, and then the i.f. amplifier could be run back parallel to the r.f. amplifier, provided there was a very large frequency difference between the r.f. and the i.f. amplifiers. However, to avoid any possible coupling, it would be better to run the i.f. amplifier off at right angles to the r.f.-amplifier line, just to be on the safe side. Good shielding is important in preventing over-all oscillation in high-gain-per-stage amplifiers, but it becomes less important when the stage gain drops to a low value. In a high-gain amplifier, the power leads (including the heater circuit) are common to all stages, and they can provide the over-all coupling if they aren't properly filtered. Good by-passing and the use of series isolating resistors will generally eliminate any possibility of coupling through the power leads. R.f. chokes are used in the heater leads where necessary, because the voltage drop through resistors would be too great.

### *Instability*

Receivers that drift rapidly as the tubes warm up and take hours to "settle down" are a nuisance. Thermal drift is a difficult thing to eliminate, but it can be minimized by using ceramic instead of bakelite insulation in the r.f. circuits (particularly the oscillator), a large cabinet compared with the chassis (to provide for good radiation of developed heat), minimizing the number of high-wattage resistors in the receiver itself and putting them in the power supply, and not mounting the oscillator coils and tuning condenser too close to a tube.

Sensitivity to vibration and shock is also a bother, and can be minimized by using good mechanical support for coils and tuning condensers, a heavy chassis, and by not hanging any of the oscillator-circuit components in the air on long leads. Tie-points should be used wherever necessary to avoid long leads on components in the oscillator circuits. Stiff long wires used for wiring components are no good if they can vibrate, and stiff short leads are excellent because they can't be made to vibrate. Susceptibility to vibration and shock is to be avoided in the oscillators of a receiver and in the first stage of a high-gain audio amplifier.

### *Hum*

A steady a.c. hum in the background of a receiver is a nuisance when listening closely for weak signals. If it appears when no signals are present, it is introduced in the audio circuit and comes from running a grid lead too close to an a.c. heater lead. Shielding or a change in location of the wire will generally eliminate it. It is a good idea to keep all audio grid leads close against the chassis if they are unshielded, to reduce their chances of picking up a.c. hum.

Hum present only on a carrier indicates a.c. modulation of the high-frequency oscillator. It can be avoided by using an oscillator circuit with a grounded cathode, by keeping the grid lead of the oscillator away from the a.c. heater leads, and by making sure that any transformer mounted on the same chassis with the high-frequency oscillator doesn't have loose laminations and a tendency to vibrate.

### *Smooth Tuning*

Smooth tuning is a great convenience to the operator, and can be obtained by taking pains with the mounting of the dial and tuning condensers. They should have good alignment and no back-lash. If the condensers are mounted off the chassis on posts instead of brackets, it is almost impossible to avoid some back-lash unless the posts have extra-wide bases. The condensers should be selected with good wiping contacts to the rotor, since with age the rotor contacts can be a source of erratic tuning. All joints in the oscillator tuning circuit should be carefully soldered, since a loose connection or "rosin joint" can develop trouble that is sometimes hard to locate.

## Servicing Superhet Receivers

### *I.F. Alignment*

A calibrated signal generator or test oscillator is a very useful device for initial alignment of an i.f. amplifier. Some means for measuring the output of the receiver is required. If the receiver has a tuning meter, its indications will serve the purpose. Lacking an S-meter, a high-resistance voltmeter or preferably a vacuum-tube voltmeter can be connected across the second-detector load resistor, if the second detector is a diode. Alternatively, if the signal generator is a modulated type, an a.c. voltmeter can be connected across the primary of the transformer feeding the 'speaker, or from the plate of the last audio amplifier through a 0.1-$\mu$fd. blocking condenser to the receiver chassis. Lacking an a.c. voltmeter, the audio output can be judged by ear, although this method is not as accurate as the others. If the tuning meter is used as an indication, the a.v.c. of the receiver should be turned on, but any other indication requires that it be turned off. Lacking a test oscillator, a steady carrier tuned through the input of the receiver (if the job is one of just touching up the i.f. amplifier) will be suitable. However, with no oscillator and tuning an amplifier for the first time, one's only recourse is to try to peak the i.f. transformers on "noise," a difficult task if the transformers are badly off resonance, as they are apt to be. It would be much better to spend a little time and haywire together a simple oscillator for test purposes.

Initial alignment of a new i.f. amplifier is as follows: The test oscillator is set to the correct frequency, and its output is connected to the grid of the last i.f. amplifier tube and to the chassis. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum output, as shown by the indicating device being used. The oscillator output lead is then clipped on to the grid of the next-to-the-last i.f. amplifier tube, and the second-from-the-last transformer trimmer adjustments are peaked for maximum output. This process is continued, working back from the second detector, until all of the i.f. transformers have been aligned. It will be necessary to reduce the output of the test oscillator as more of the i.f. amplifier is brought into use, because the increased gain is likely to cause overloading and consequent inaccurate adjustments. It is desirable in all cases to use the minimum oscillator signal that will give useful output readings. The i.f. transformer in the plate circuit of the mixer is aligned with the signal introduced to the grid of the mixer. Since the tuned circuit feeding the mixer grid may have a very low impedance at the i.f., it may be necessary to boost the test generator output or to disconnect the circuit temporarily from the mixer grid.

If the i.f. amplifier has a crystal filter, the filter should first be switched out and the alignment carried out as above, setting the test oscillator as closely as possible to the crystal frequency. When this is completed, the crystal should be switched in and the oscillator frequency varied back and forth over a small range either side of the crystal frequency to find the exact frequency, as indicated by a sharp rise in output. Leaving the test oscillator set on the crystal peak, the i.f. trimmers should be realigned for maximum output. The necessary readjustment should be small. The oscillator frequency should be checked frequently to make sure it has not drifted from the crystal peak.

A modulated signal is not of much value for aligning a crystal-filter i.f. amplifier, since the high selectivity cuts sidebands and the results may be inaccurate if the audio output is used as the tuning indication. Lacking the a.v.c. tuning meter, the transformers may be conveniently aligned by ear, using a weak unmodulated signal adjusted to the crystal peak. Switch on the beat oscillator, adjust to a suitable tone, and align the i.f. transformers for maximum audio output.

An amplifier that is only slightly out of alignment, as a result of normal drift or aging, can be realigned by using any steady signal, such as a local broadcast station, instead of the test oscillator. One's 100-kc. standard makes an excellent signal source for "touching up" an i.f. amplifier. Allow the receiver to warm up thoroughly, tune in the signal, and trim the i.f. for maximum output.

If you bought your receiver instead of making it, be sure to read the instruction book carefully before attempting to realign the receiver. Most instruction books include alignment details, and any little special tricks that are peculiar to that particular type of receiver will also be described.

### *R.F. Alignment*

The objective in aligning the r.f. circuits of a gang-tuned receiver is to secure adequate tracking over each tuning range. The adjustment may be carried out with a test oscillator of suitable frequency range, with harmonics from your 100-kc. standard or other known oscillator, or even on noise or such signals as may be heard. First set the tuning dial at the high-frequency end of the range in use. Then set the test oscillator to the frequency indicated by the receiver dial. The test-oscillator output may be connected to the antenna terminals of the receiver for this test. Adjust the oscillator trimmer condenser in the receiver to give maximum response on the test-oscillator signal, then reset the receiver dial to the low-frequency end of the range. Set the test-oscillator frequency near the frequency indicated by the receiver dial and carefully tune the test oscillator until its signal is heard in the receiver. If the frequency of the signal as indicated by the test-oscillator calibration is higher than that indicated by the receiver dial, more inductance (or more capacity in the tracking condenser) is needed in the receiver oscillator circuit; if the frequency is lower, less inductance (less tracking capacity) is required in the receiver oscillator. Most commercial receivers provide some means for varying the inductance of the coils or the capacity of the tracking condenser, to permit aligning the receiver tuning with the dial calibration. Set the test oscillator to the frequency indicated by the receiver dial, and then adjust the tracking capacity or inductance of the receiver oscillator coil to obtain maximum response. After making this adjustment, recheck the high-frequency end of the scale as previously described. It may be necessary to go back and forth between the ends of the range several times before the proper combination of inductance and capacity is secured. In many cases, better over-all tracking will result if frequencies near but not actually at the ends of the tuning range are selected, instead of taking the extreme dial settings.

After the oscillator range is properly adjusted, set the receiver and test oscillator to the high-frequency end of the range. Adjust the mixer trimmer condenser for maximum hiss or signal, then the r.f. trimmers. Reset the tuning dial and test oscillator to the low-frequency end of the range, and repeat; if the circuits are properly designed, no change in trimmer settings should be necessary. If it is necessary to increase the trimmer capacity in any circuit, it indicates that more inductance is needed; if less capacity resonates the circuit, less inductance is required.

Tracking seldom is perfect throughout a tuning range, so that a check of alignment at intermediate points in the range may show it to be slightly off. Normally the gain variation from this cause will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned for maximum performance in that region, even though the ends of the frequency range as a whole may be slightly out of alignment.

### *Oscillation in R.F. or I.F. Amplifiers*

Oscillation in high-frequency amplifier and mixer circuits may be evidenced by squeals or "birdies" as the tuning is varied, or by complete lack of audible output if the oscillation is strong enough to cause the a.v.c. system to reduce the receiver gain drastically. Oscillation can be caused by poor connections in the common ground circuits. Inadequate or defective by-pass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. A metal tube with an ungrounded shell will cause trouble. Improper screen-grid voltage, resulting from a shorted or too-low screen-grid series resistor, also may be responsible for such instability.

Oscillation in the i.f. circuits is independent of high-frequency tuning, and is indicated by a continuous squeal that appears when the gain is advanced with the c.w. beat oscillator on. It can result from defects in i.f.-amplifier circuits similar to those above. Inadequate cathode by-pass capacitance is a common cause of such oscillation. An additional by-pass condenser of 0.1 to 0.25 $\mu$fd. often will remedy the trouble. Similar treatment can be applied to the screen-grid and plate by-pass filters of i.f. stages.

### *Instability*

"Birdies" or a mushy hiss occurring with tuning of the high-frequency oscillator may indicate that the oscillator is "squegging" or oscillating simultaneously at high and low frequencies. This may be caused by a defective tube, too-high oscillator plate or screen-grid voltage, excessive feed-back, or too-high grid-leak resistance.

A varying beat-note in c.w. reception indicates instability in either the h.f. oscillator or beat oscillator, usually the former. The stability of the beat oscillator can be checked by introducing a signal of intermediate frequency (from a test oscillator) into the i.f. amplifier; if the beat-note is unstable, the trouble is in the beat oscillator. Poor connections or defective parts are the likely cause. Instability in the high-frequency oscillator may be the result of poor circuit design, loose connections, defective tubes or circuit components, or poor voltage regulation in the oscillator plate-and/or screen-supply circuits. Mixer pulling of the oscillator circuit also will cause the beat-note to "chirp" on strong c.w. signals because the oscillator load changes slightly.

In 'phone reception with a.v.c., a peculiar type of instability ("motorboating") may appear if the h.f.-oscillator frequency is sensitive to changes in plate voltage. As the a.v.c. voltage rises the electrode currents of the controlled tubes decrease, decreasing the load on the power supply and causing its output voltage to rise. Since this increases the voltage applied to the oscillator, its frequency changes correspondingly, throwing the signal off the peak of the i.f. resonance curve and reducing the a.v.c. voltage, thus tending to restore the original conditions. The process then repeats itself, at a rate determined by the signal strength and the time constant of the power-supply circuits. This effect is most pronounced with high i.f. selectivity, as when a crystal filter is used, and can be cured by making the oscillator relatively insensitive to voltage changes and by regulating the plate-voltage supply.

## Narrow-Band Frequency- and Phase-Modulation Reception

The only FM (frequency modulation) and PM (phase modulation) in general use on the frequencies below 30 Mc. is of the "narrow-band" variety, in which the channel width of the signal is no greater than that of an AM signal. Normal wide-band FM (and PM) receiving techniques, which involve the use of limiter and discriminator circuits, are not generally applicable to the reception of narrow-band FM (and PM), and most amateurs use their regular communications receivers with no modification.

### *FM Reception*

In the reception of FM signals by a normal communications receiver, the a.v.c. is switched off and the incoming signal is not tuned "on the nose," as indicated by maximum reading of the S-meter, but slightly off to one side or the other. This puts the carrier of the incoming signal on one side or the other of the i.f. selectivity characteristic (see Fig. 5-1). As the frequency of the signal changes back and forth over a small range with modulation, these variations in frequency are translated to variations in amplitude, and the consequent AM is detected in the normal manner. The signal is tuned in (on one side or the other of maximum carrier strength) until the audio quality appears to be best. The audio output from the signal depends on the *slope* of the i.f. characteristic and the amount of *swing* (deviation) of the signal. If the audio is too weak, the transmitting operator should be advised to increase his swing slightly, and if the audio quality is bad ("splashy" and with serious distortion on volume peaks) he should be advised to reduce his swing. Coöperation between transmitting and receiving operators is a necessity for best audio quality. The transmitting station should always be advised immediately if at any time his bandwidth exceeds that of an AM signal, since this is a violation of the FCC regulations, except in those portions where wide-band FM is permitted.

### *PM Reception*

PM signals can be received in the same way that NFM (narrow-band FM) signals are, except that in this case the audio output will appear to be lacking in "lows," because of the differences in the deviation-*vs.*-audio characteristics of the two systems. This can be remedied to a considerable degree by advancing the tone control of the receiver to the point where more nearly normal speech output is obtained.

NPM signals can also be received on communications receivers by making use of the crystal filter, in which case there is no need for audio compensation. The crystal filter should be set to the sharpest position and the carrier should be tuned in on the crystal peak, *not* set off to one side. The phasing condenser should be set not for exact neutralization but to give a rejection notch at some convenient side frequency such as 1000 cycles off resonance. There is considerable attenuation of the side bands with such tuning, but it is not *selective* attenuation except for the added selectivity of the i.f. amplifier, and it can readily be overcome by using additional audio gain. FM signals received through the crystal filter in this fashion will have a "boomy" characteristic because the lower frequencies are accentuated.

CHAPTER 6

# High-Frequency Transmitters

A transmitter for the low-frequency amateur bands (4 to 30 Mc.) consists essentially of an oscillator and usually, but not always, one or more amplifiers, one of which feeds radio-frequency power to the antenna or transmission line.

The oscillator, fed by a source of d.c. power, such as rectified 60-cycle house current, a d.c. generator or batteries, serves to generate a relatively small amount of power at a selected radio frequency. The amplifiers perform the function of increasing the r.f. power at that frequency, or a multiple of it, to the level desired before feeding it to the antenna system. In other words, a transmitter is a device that converts d.c. power to power at radio frequencies. Several typical transmitter arrangements are shown in the diagrams of Fig. 6-1. At A, the oscillator feeds directly into the antenna. At B, the output of the oscillator is fed to an amplifier which feeds the antenna. In the arrangements of C, D and E, the oscillator output is fed to a frequency-multiplying amplifier which doubles the frequency before passing the energy along to the next stage. As the diagram indicates, it is often possible to operate more than one transmitter stage from a single power supply.

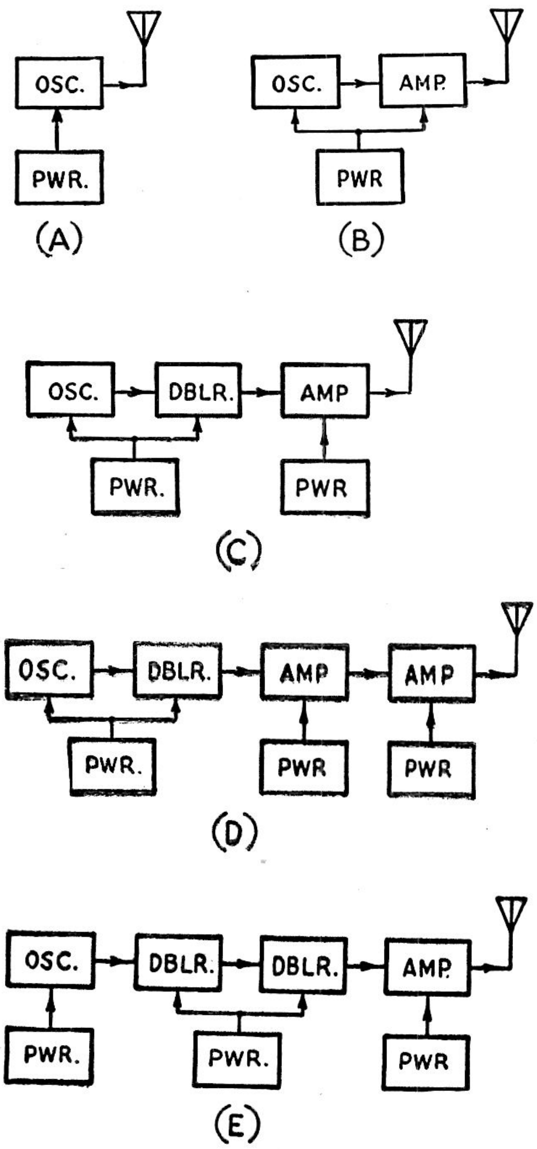



***Fig. 6-1*** **— Block diagrams showing typical combinations of oscillator and amplifiers and' power-supply arrangements for transmitters. A wide selection is possible depending upon the number of bands in which operation is desired and the power output.**

In a telegraph (c.w.) transmitter, means is provided for breaking the transmitted energy up into the characters of the Continental (International Morse) code. In 'phone transmitters, power at audio voice frequencies is combined with the d.c. input so as to superimpose (modulate) those frequencies on the steady r.f. output (carrier) from a transmitter which otherwise is essentially the same as that most often used for telegraphy.

To minimize interference when a large number of stations must work in one frequency band, the power output of a transmitter must be as stable in frequency and as free from spurious radiations as the state of the art permits. The steady r.f. output must be free from amplitude variations attributable to ripple from the plate power supply or other causes, its frequency should be unaffected by variations in supply voltages or inadvertent changes in circuit constants, and there should be no radiation on other than the intended frequency which might cause interference to other radio services.

The design of any particular transmitter is based primarily upon the power-output level desired and the number of bands to be covered and their frequencies. Added to these are the factors of operating convenience and space restrictions. Portable and mobile equipment require special consideration in design.

# Variable-Frequency Oscillators

Two general classes of oscillators are used in amateur transmitters. A crystal-controlled oscillator is a fixed-frequency oscillator. The frequency generated is held within very close limits (a few cycles per megacyle) by a quartz crystal. The frequency is determined almost entirely by the thickness of the crystal. Other constants in the circuit have relatively little effect. The frequency of a self-controlled or variable-frequency oscillator (VFO) is determined principally by the values of inductance and capacitance which make up the oscillator tank circuit.

The disadvantage of the crystal type of oscillator is that a different crystal must be used for each frequency desired in any one amateur band. By making the inductance, capacitance, or both variable in the self-controlled oscillator, it may be operated at any frequency desired at the turn of a dial, in the manner of a receiver. The disadvantage of a VFO is that much care must be exercised in the design and construction if the frequency stability is to approach that of a crystal-controlled oscillator.

Although the trend in recent years has been toward the VFO with its greater flexibility, the crystal oscillator still is widely used by beginners and in certain types of operation because of the ease with which frequency stability and calibration may be maintained.

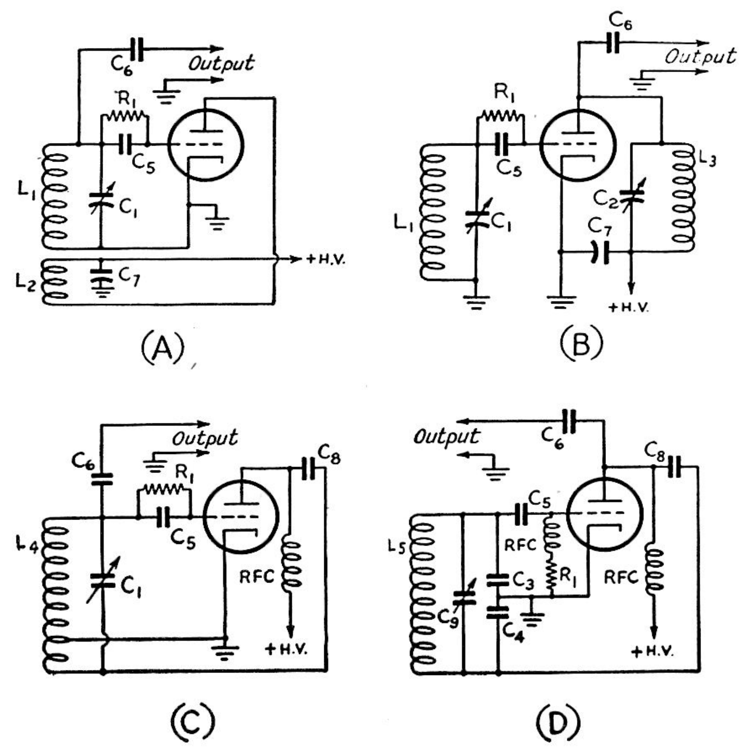
## VFO CIRCUITS

### *Tickler-Feed-Back Circuit*

Fig. 6-2 shows several types of self-controlled oscillator circuits. In the tickler-feed-back circuit of A, the frequency is determined principally by the values of $C_1$ and $L_1$. Feed-back is provided by the coupling between $L_1$ and $L_2$. It may be adjusted by varying the number of turns in $L_2$ or by changing the coupling between $L_1$ and $L_2$. When connected as shown, the two coils must be wound in the same direction. If they are wound in opposite directions, the connections to *one* of the coils must be reversed.

### *T.G.T.P. Oscillator*

In the tuned-grid tuned-plate (t.g.t.p.) circuit of Fig. 6-2B, the $L_1C_1$ tank has the higher $Q$ because of the relatively high value of $C_1$. Therefore it forms the principal control over frequency. $L_3C_2$ is used primarily for adjustment of feed-back, although its effect upon frequency is considerable. Feed-back is through the plate-grid capacitance of the tube. $L_1$



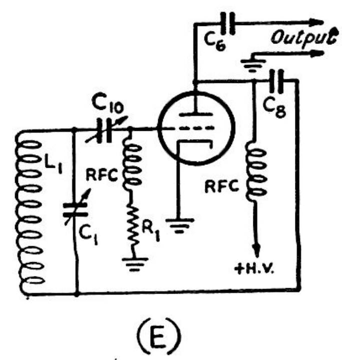




*Fig. 6-2* — VFO circuits. A — Tickler feed-back. B — Tuned-grid tuned-plate. C — Hartley. D — Colpitts. E — Ultraudion.

Approximate values are as follows:

- $C_1$ — Tank tuning condenser. For 3.5 Mc. — 500 $\mu\mu$fd. or more total, including any fixed capacitance which may be used for bandspread purposes.
- $C_2$ — Plate tank condenser — 100-$\mu\mu$fd. variable.
- $C_3$ — Tank condenser — 0.003-$\mu$fd. mica for 3.5 Mc.
- $C_4$ — Tank condenser — 0.001-$\mu$fd. mica for 3.5 Mc.
- $C_5$ — Grid condenser — 100 $\mu\mu$fd. or less, mica.
- $C_6$ — Output coupling condenser — 100 $\mu\mu$fd. or less, mica.
- $C_7$ — Plate by-pass condenser — 0.01-$\mu$fd. paper.
- $C_8$ — Plate blocking condenser — 0.001-$\mu$fd. mica.
- $C_9$ — Tuning condenser — 250-$\mu\mu$fd. variable.
- $C_{10}$ — Grid condenser — 100 $\mu\mu$fd. or less, variable.
- $R_1$ — Grid leak — 50,000 ohms.
- $L_1$ — Tank coil — 4.3 $\mu$hy. for 3.5 Mc. with 500 $\mu\mu$fd.
- $L_2$ — Tickler winding — approximately one-third the number of turns in $L_1$, wound over bottom portion of $L_1$ or adjacent to it.
- $L_3$ — Plate tank coil — 22 $\mu$hy. for 3.5 Mc. with 100-$\mu\mu$fd. condenser.
- $L_4$ — Tank coil — same as $L_1$, tapped approximately one-third from plate end.
- $L_5$ — Tank coil — 3.3 $\mu$hy. for 3.5 Mc. with capacitance values given for $C_3$, $C_4$ and $C_9$.
- RFC — Parallel-feed r.f. choke — 2.5 mh.

and $L_3$ are not coupled. For positive feed-back, $L_3C_2$ must be tuned to a frequency higher than that of $L_1C_1$. Maximum feed-back occurs when the two circuits are tuned close (but not exactly) to the same frequency. Power output is taken from the plate tank circuit, $L_3C_2$.

### *Hartley*

The Hartley circuit of Fig. 6-2C is quite similar to the tickler-feed-back circuit of A, except that the tank condenser is connected across both plate and grid portions of a single winding. The frequency is determined principally by the values of $L_4$ and $C_1$. Feed-back may be adjusted by moving the cathode tap which is the r.f. ground point.

### *Colpitts*

The Colpitts circuit of Fig. 6-2D also is similar, the principal difference being that the voltage division between the grid and plate portions of the circuit is by capacitive means rather than inductive. Frequency is determined chiefly by the values of $L_5$, and the resultant of $C_3$ and $C_4$ in series and $C_9$ in parallel. Feed-back may be adjusted by changing the ratio of $C_3$ to $C_4$, $C_9$ being used to adjust frequency.

### *Ultraudion*

A variation of the Colpitts is the ultraudion circuit of Fig. 6-2E. The grid-cathode and plate-cathode capacitances of the tube serve as the capacitive voltage divider. The circuit is tuned to the desired frequency by $C_1$. Feed-back may be adjusted within limits by alteration of the value of the grid coupling condenser, $C_{10}$. (See "Frequency-Multiplying Oscillators" this chapter for additional VFO circuits.)

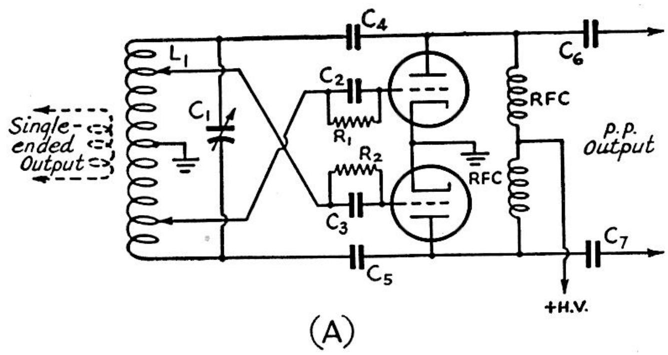
### *Push-Pull VFO Circuits*

Two tubes may be used in push-pull in an oscillator. The two tubes provide an increase in power output over that obtainable from a single tube. Since they are effectively in series across the tuned circuit, their effect on frequency stability is less than that of a single tube. So far as tube effects are concerned, the same order of stability may be obtained with half the total capacitance in the tank circuit required for a single tube.

The Hartley and tickler-feed-back circuits lend themselves most readily to a push-pull arrangement, as shown in Fig. 6-3. In the Hartley of A, excitation may be adjusted by moving the grid taps, keeping them in positions symmetrical in respect to the center-tap. In the tickler-feed-back circuit of B, excitation is adjusted by changing the number of turns in $L_3$, keeping the number of turns equal either side of the center-tap. $L_3$ should be wound over the center portion of $L_2$.

Capacitance coupling may be used if a push-pull oscillator is to feed a push-pull amplifier, but inductive coupling, which assures a better circuit balance, is preferable if the oscillator is to be coupled to a single-tube amplifier. (See "Interstage Coupling" this chapter.)

(See "Frequency-Multiplying Oscillators" this chapter for additional VFO circuits.)



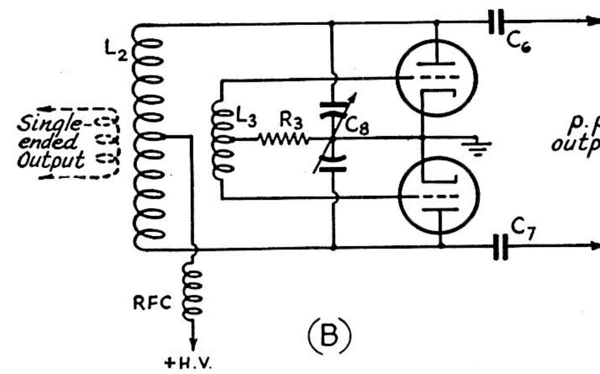




*Fig. 6-3* — **Push-pull VFO circuits. A — Hartley. B — Tickler feed-back.**

**Approximate values are as follows:**

**$C_1$ — Tank condenser — 500 $\mu\mu$fd. total, including any fixed capacitance used for bandspread purposes, for 3.5 Mc.**
**$C_2$, $C_3$ — Grid condenser — 100 $\mu\mu$fd. or less, mica.**
**$C_4$, $C_5$ — Plate blocking condenser — 0.001-$\mu$fd. mica.**
**$C_6$, $C_7$ — Output coupling condenser — 100 $\mu\mu$fd. or less, mica.**
**$C_8$ — Tank condenser — 500 $\mu\mu$fd. per section, including any fixed capacitance used for bandspread purposes, for 3.5 Mc.**
**$R_1$, $R_2$ — Grid leak — 50,000 ohms.**
**$R_3$ — Grid leak — 25,000 ohms.**
**$L_1$ — Tank coil — 4.3 $\mu$hy., tapped at center, for 3.5 Mc., and 500 $\mu\mu$fd.**
**$L_2$ — Tank coil — 8.6 $\mu$hy., tapped at center, for 3.5 Mc., and 250 $\mu\mu$fd. total across coil.**
**$L_3$ — Tickler coil — approximately one-third number of turns in $L_2$, tapped at center, wound over center of $L_2$ or between halves of $L_2$.**
**RFC — R.f. choke — 2.5 mh.**

### *Factors in VFO Design and Construction*

To provide satisfactory performance on the air, considerable care must be exercised in the design and construction of a VFO. Since the frequency depends upon the $L$ and $C$ in the circuit, anything which operates to change these values will cause a change in frequency. For stability which will approach that of which a crystal oscillator is capable, the values of inductance and capacitance must be held within extremely small tolerances.

It is not too difficult to provide a satisfactory coil and condenser for the tank circuit. But the tube must be connected across this circuit and its effect upon frequency is by no means negligible nor easily controlled. The tube has the effect of a capacitance which can be made to hold satisfactorily constant only with great care. It is obvious too that the connection of any reactive load, such as an antenna or the input of an amplifier stage, will change the frequency, since this load must be connected across the frequency-determining circuit, thereby changing the net value of inductance or capacitance as the case may be. An antenna and feeders cannot be held sufficiently rigid to prevent changes in their capacitance. Under practical operating conditions the input circuit of an amplifier may develop changes in the reactance which it presents across the oscillator circuit, especially while it is being tuned or alternately connected and disconnected, which it is in effect if the amplifier is keyed. Isolating amplifiers (see "R.F. Power Amplifiers" this chapter) can be used to reduce these effects.

### *Chirp*

Variations in the plate voltage of the oscillator tube cause changes of appreciable magnitude in the effective input capacitance of the tube. If the oscillator can be run continuously during transmission, this effect can be made negligible by the use of a regulated plate-voltage supply. But if the oscillator must be keyed for break-in work, an objectionable shift in frequency with keying **(chirp)** can be avoided only by reducing the time constant of the keying circuit to the point where the change in frequency between zero-voltage and full-voltage conditions takes place so rapidly that the ear cannot detect it. The time constant is reduced by minimizing any capacitance which may appear across the key contacts, including by-pass condensers in the transmitter. Unfortunately, as discussed in Chapter Eight, a certain time lag is required to eliminate clicks. Therefore the measures necessary for the elimination of chirps and clicks are in opposition. A compromise is usually necessary. It is possible that the keying of an amplifier may constitute little improvement over oscillator keying, for reasons previously given, unless sufficient isolation is provided between the oscillator and the keyed stage.

### *Drift*

The effects of temperature are characterized by a slow drift or creep in frequency. Part of this change, especially for the first few minutes after power is applied to the oscillator, may be attributable to change in tube-electrode capacitance as the tube heats up. But over a protracted period of time, drift is a result of small changes with temperature in physical dimensions of the coil and condenser in the tank circuit. Good design dictates that these components be isolated as much as possible from the heat developed in the tubes and power-supply equipment. With care, frequency drift can be brought within satisfactory limits by mounting the tubes external to the enclosure surrounding the tank coil and condensers and the use of zero-temperature mica condensers for all tank capacitance other than that required for tuning purposes, by providing ventilation and by keeping the power input to the oscillator at a minimum — not more than a few watts. Where maximum stability with temperature change is desired, temperature-compensating condensers may be used to form part of the tank-circuit capacitance.

### *Mechanical Considerations*

Any mechanical vibration which causes a change in the capacitance across the tank circuit will cause a corresponding change in frequency. This should be minimized by solid construction, secure wiring and by cushioning the mounting of the oscillator unit against shocks. The oscillator should be thoroughly shielded from the strong r.f. fields of the antenna and adjacent high-power amplifier stages which may, through overloading of the oscillator grid, cause roughening of the oscillator signal.

### *VFO Tank Capacitance and Tuning Systems*

All of the previously-mentioned effects upon the frequency of an oscillator may be minimized by the use of high capacitance in the tank circuit, thus making uncontrollable capacitance changes a small percentage of the total circuit capacitance. The same effect can be obtained by tapping the tube elements across only a portion of the circuit, as shown in Fig. 6-4, so that the tube capacitance shunts only a part of the tank circuit. A tube of low or medium amplification factor is preferable as an oscillator because changes in loading — tuning of the output circuit of the oscillator or following amplifier stages as well as plate-voltage variations — have less influence upon the effective input capacitance of the tube than with a high-$\mu$ tube.

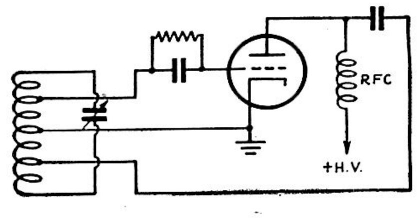



**Fig. 6-4** — Oscillator tubes are sometimes connected across only a portion of the tank coil to reduce the effects of the tube on the tuned circuit. In this Hartley circuit, the grid and plate connections are made to taps instead of to the ends of the coil.

Because it is considered easier to maintain percentage stability at lower frequencies, VFOs usually are designed to operate at a frequency not higher than the 3.5-Mc. band, the higher-frequency bands being reached by frequency-multiplier stages which are discussed under

"Frequency-Multiplying Amplifiers." At 3.5 Mc., a tank capacitance of 500 to 1000 $\mu\mu$fd. is considered adequate, with values increased in proportion if the oscillator is designed to operate at lower frequencies.

Any of the bandspread tuning systems used in receivers may be applied to the oscillator circuits which have been under discussion. The parallel-condenser system is used most widely since it lends itself well to the high-*C* circuits which should be used in VFOs. Plug-in coils for changing oscillator frequency ranges are not recommended because experience has shown that the coil contacts may become the source of frequency instability.

## VFO ADJUSTMENT

### *Tuning Characteristics*

With the exception of the t.g.t.p., all normally-operating VFO circuits will function quite uniformly, over the range of an amateur band at least, as soon as plate voltage is applied. If, through incorrect adjustment of excitation or overloading the circuit does not oscillate, the plate current will be the zero-bias value for the tube at the plate voltage at which it is being operated, falling to a lower value when oscillation takes place. If the oscillator is functioning, touching the grid with the finger will cause a variation in plate current. The value of plate current to be expected with a given tube when oscillating depends upon such factors as plate voltage, grid-leak resistance, excitation adjustment and loading. It should remain essentially constant with reasonable changes in tuning capacitance. With normal excitation adjustment, the plate current should show an increase when the load is connected. Excitation and grid-leak resistance should be adjusted for maximum frequency stability — not maximum output.

With the t.g.t.p. circuit, oscillation takes place only when the plate tank circuit is tuned to a frequency higher than that to which the grid circuit is tuned, and maximum output usually occurs when the two are tuned close (but not exactly) to the same frequency. If the plate tuning condenser has sufficient range to tune the circuit to a frequency lower than that of the grid circuit, oscillation will cease and the plate current will be relatively high, as shown at the left in Fig. 6-5. As the plate circuit is tuned past the point of resonance with the grid circuit in the high-frequency direction, the plate current will drop suddenly (point *A*) indicating the starting of oscillation, then dip rapidly to a minimum (point *C*) where the power output is greatest. As the tuning capacitance is decreased further, the plate current will rise gradually to point *B* where it will jump to a higher value, indicating that oscillation has ceased. For maximum frequency stability, the circuit should be tuned in the region *D-E*.

When the oscillator is loaded, the characteristic is similar (see dashed curve in Fig. 6-5),

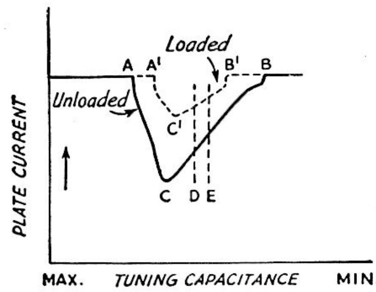



*Fig. 6-5* — General tuning characteristic of the t.g.t.p. VFO circuit and triode, tetrode or pentode crystal oscillators. As the capacitance of the plate tank condenser is decreased from maximum, oscillation will start suddenly, the plate current dropping abruptly, as shown at point *A*. As the capacitance is decreased further, the plate current will dip to a minimum, *C*, and then gradually rise to point *B* where an abrupt rise in plate current will indicate that oscillation has ceased. Maximum output will be obtained at point *C*, but the oscillator should be adjusted for operation in the *D-E* region for best frequency stability.

but the minimum plate-current dip is much less pronounced and the range of plate tuning over which the circuit will oscillate becomes less as the loading is increased.

If the oscillator is delivering sufficient power, an r.f. indicator should show maximum r.f. voltage at the grid and plate terminals and decrease to zero voltage at the cathode in all of the circuits shown in Fig. 6-2. The Hartley circuit of Fig. 6-2C, the Colpitts of D and the ultraudion circuit of E have the disadvantage in construction that the tuning control (tuning-condenser rotor shaft) must be insulated.

### *Checking VFO Stability*

A VFO should be checked thoroughly before it is placed in regular operation on the air. Since succeeding amplifier stages may affect the signal characteristics, final tests should be made with the complete transmitter in operation. Almost any VFO will show signals of good quality and stability when it is running free and not connected to a load. A well-isolated monitor is a necessity. Perhaps the most convenient, as well as one of the most satisfactory, monitoring arrangements is a well-shielded receiver combined with a crystal oscillator. (See "Crystal Oscillators.") The crystal frequency should lie in the band of the lowest frequency to be checked and in the frequency range where its harmonics will fall in the higher-frequency bands. The receiver b.f.o. is turned off and the VFO signal is tuned to beat with the signal from the crystal oscillator instead. In this way any receiver instability caused by overloading of the input circuits which may result in "pulling" of the h.f. oscillator in the receiver, or by a change in line voltage to the receiver when the transmitter is keyed, will not affect the reliability of the check. Most present-day crystals have a sufficiently-low temperature coefficient to give a satisfactory check on drift as well as on chirp and signal quality if they are not overloaded.

Harmonics of the crystal may be used to

beat with the transmitter signal when monitoring at the higher frequencies. Since any chirp observed at the lower frequencies will be magnified at the higher frequencies, most-accurate checking can be done by monitoring at the higher frequencies.

The distance between the crystal oscillator and receiver should be adjusted to give a good beat between the crystal oscillator and the transmitter signal. When using harmonics of the crystal oscillator, it may be necessary to attach a piece of wire to the oscillator as an antenna to give sufficient signal in the receiver.

Checks may show that the stability is sufficiently good to permit oscillator keying at the lower frequencies, where break-in operation is of greater value, but that chirp becomes objectionable at the higher frequencies. If further improvement does not seem possible, it would be logical in this case to use oscillator keying at the lower frequencies and amplifier keying at the higher frequencies.

# Crystal Oscillators

## ● SIMPLE CIRCUITS

The simplest crystal-controlled oscillator circuits are shown in Fig. 6-6. Since the crystal is the equivalent of a tuned circuit of fixed frequency, it will be observed that each of the crystal circuits is essentially the equivalent of one of the VFO circuits of Fig. 6-2.

### *Triode, Tetrode and Pentode Oscillators*

The triode crystal circuit of Fig. 6-6A is the equivalent of the t.g.t.p. circuit in which the crystal replaces the tuned grid circuit. The pentode circuit of B is the same except for the substitution of a screen-grid tube for the triode. This circuit sometimes is operated with the suppressor by-passed and raised to a positive voltage of about 50 instead of grounded as shown. The same circuit is used for tetrodes, such as the 6V6 and 6L6, the suppressor connection being omitted. Tuning characteristics are similar to those of the t.g.t.p. circuit shown in Fig. 6-5.

### *Pierce Oscillator*

The circuit shown in Fig. 6-6C is known as the Pierce circuit. It is the equivalent of the ultraudion variation of the Colpitts. The crystal replaces the single tuned circuit and thus this oscillator requires no tuning adjustment and will work without change in values over a wide range of crystal frequencies. Since excitation otherwise is not adjustable, the condenser $C_5$ sometimes is required to obtain satisfactory operation. Less power is obtainable from the Pierce circuit than from the others because the crystal is directly in the power-delivering circuit which limits the r.f. voltage that may be developed without danger to the crystal. Triodes also may be used in this circuit. (See "Frequency-Multiplying Oscillators" this chapter for additional crystal-oscillator circuits.)

## ● POWER LIMITATIONS OF CRYSTALS

Although high-power crystal oscillators sometimes are used in cases where utmost simplicity must be the paramount consideration, the oscillator normally should be considered as a frequency-generating device only, with power output of secondary importance. While crystal-controlled oscillators are much more tolerant than VFOs in respect to temperature changes, the danger of crystal fracture, as well as drift, places a limitation on the amount of power output obtainable. With the

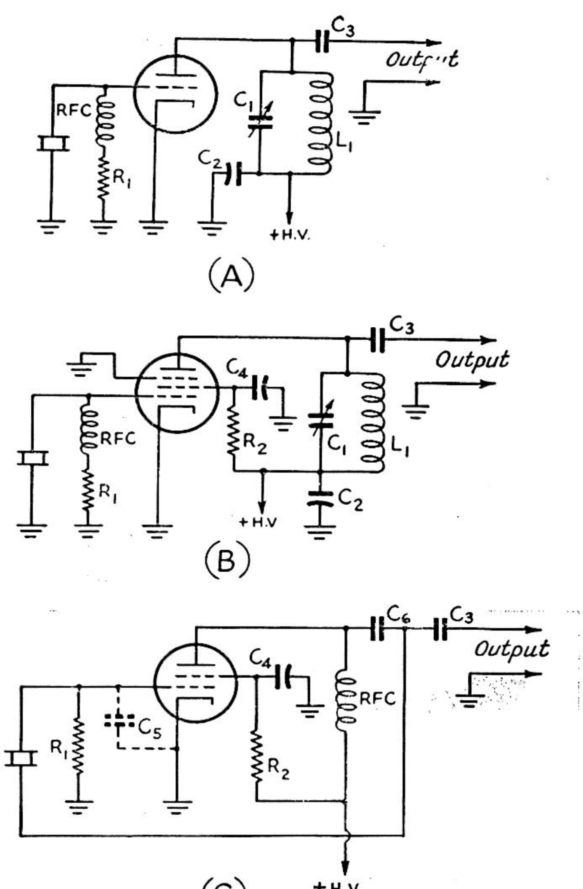



*Fig. 6-6* — Simple crystal-oscillator circuits. A — Triode. B — Tetrode or pentode. C — Tetrode Pierce. Other crystal-oscillator circuits are shown in Fig. 6-18.

Approximate values are as follows:

$C_1$ — Tank condenser — 100-$\mu\mu$fd. variable.
$C_2$ — Plate by-pass condenser — 0.01-$\mu$fd. paper.
$C_3$ — Output coupling condenser — 100 $\mu\mu$fd. or less, mica.
$C_4$ — Screen by-pass condenser — 0.01-$\mu$fd. paper.
$C_5$ — Feed-back condenser — 50 to 100 $\mu\mu$fd.
$C_6$ — Plate blocking condenser — 0.001-$\mu$fd. mica.
$R_1$ — Grid leak — 50,000 ohms.
$R_2$ — Screen voltage-dropping resistor — 25,000 to 50,000 ohms.
$L_1$ — Tank coil — 22 $\mu$hy. for 3.5 Mc.; 7.5 $\mu$hy. for 7 Mc.
RFC — Parallel-feed r.f. choke — 2.5 mh.

large prewar-type crystals, triode crystal oscillators may be operated with proper adjustment at plate voltages as high as 200 or 250, but the voltage should be reduced to 150 to 200 for the new-type smaller crystals. Low- or medium-$\mu$ triodes are preferable as oscillators. Beam tetrodes or pentodes, with their high power-sensitivity and reduced grid-plate capacitance, require less voltage across the crystal than a triode for the same amount of output. Therefore the larger-type crystals can be operated with plate voltages of 300 or 400 with power output up to 10 or 15 watts if required. The smaller crystals may take plate voltages up to 300 or 325 before showing marked instability. However, as stated previously, it is always advisable to limit the input to the oscillator and depend upon amplifiers for the desired power.

With the simple crystal-oscillator circuits shown in Fig. 6-6, excitation, and therefore r.f. voltage across the crystal, is greatest when the oscillator is unloaded and it is under this condition that danger to the crystal is greatest.

## GRINDING CRYSTALS

Crystal blanks, cut to approximate frequency, are available at very reasonable prices. With proper equipment and a little care, these blanks can be ground to the desired frequency. Complete crystal-grinding equipment includes several components. First necessity is a flat piece of plate glass, about 4 inches square or larger. To hold the crystal flat while grinding a flat "button" (shown in Fig. 6-7), also of plate glass, either round or square and slightly larger than the crystal, is required. Both pieces may be obtained at glass stores. Two grades of abrasive, No. 303 emery for surface grinding and No. 600 Carborundum for edge grinding

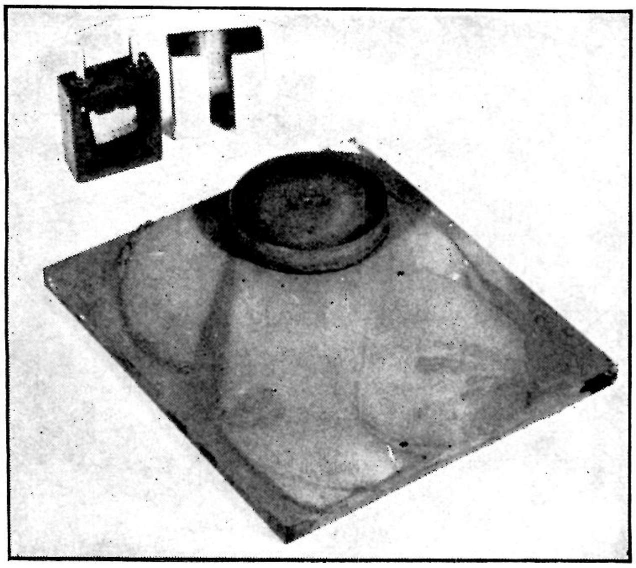


*Fig. 6-7* — **The equipment necessary for grinding a crystal blank to frequency. A piece of plate glass and a "button" of the same material are essential. The "quick-change" adaptation for the crystal holder is a convenience. Not shown, but also convenient, are a small paint brush for spreading abrasive and a toothbrush for scrubbing.**

and beveling are obtainable from hardware stores or opticians'-supply houses. A small paint brush is handy for moistening the abrasive and spreading it around the lapping plate. To facilitate frequent checking of frequency during the grinding process, the quick-change holder shown in Fig. 6-8 is desirable. It consists of an FT243 holder with a sliding cover fashioned from sheet metal. Soap, warm water and a toothbrush are used to clean and rinse

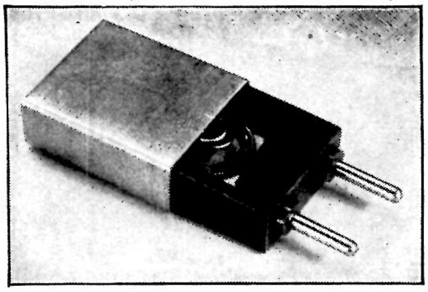


*Fig. 6-8* — **The quick-change crystal holder with sliding cover.**

the crystal. Lintless cloth from an optician's or a clean towel can be used for drying.

Present-day electrodes have raised lands on each corner, as shown in Fig. 6-9, and the crystal should lie at least halfway across these lands and should not be larger than the electrode. The electrodes should be cleaned as carefully as the crystal. Before final assembly both crystal and electrodes should be handled carefully by the corners or edges after their last good scrubbing.

### *How To Grind*

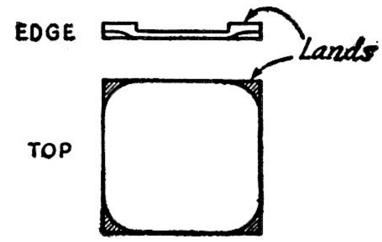
The actual grinding is done as follows: Spread the 303 abrasive over an area about a half inch square on the lapping plate, wet the brush, mix water into the spot and spread the abrasive over the lapping plate. Always keep the abrasive moist. Take the button, put a drop of water at its center, and press the dry crystal blank over the drop of water. There should be just enough water in the drop so that it squeezes out under the edges of the blank, where it is wiped away. Place the button, blank down, on the emery and put the index finger in the center of the button. Use just enough pressure to move the button in a figure-8 pattern. This motion is used because it helps keep the blank flat.

After grinding through ten or fifteen "8s" the blank should be rechecked for frequency and activity. The blank's activity is a term used in crystal making to describe how strongly a crystal will oscillate. This may be indicated by the magnitude of the dip in the plate current, grid current to the next stage, or rectified grid current in the crystal oscillator. (See "Maximum Grid Current" this chapter.) It is nearly impossible to tell how much change in frequency will occur during the grinding of a crystal, because pressure on the button, the amount of abrasive, and the area of the "8" all will vary the frequency. The frequency change probably will be between 200 and 1000 cycles per "8," using a 7-Mc. crystal. The crystal can be moved along faster as the operator becomes more familiar with the technique, but for the beginner frequent checks of activity are in order so that any drop can be corrected.

To grind a crystal successfully the activity must be good when the crystal is brought to the desired frequency. There are several ways to raise the activity. Assuming that, with careful grinding on a flat plate with a flat button, the two faces of the crystal are parallel, the major cause of low activity will be dirt or moisture on the crystal or electrodes. Before checking activity the crystal should be scrubbed carefully with the toothbrush, using warm water and soap. Wipe the crystal clean and be sure that the electrodes are clean and dry. If the activity is still down the next thing is to bevel all eight edges of the crystal. The beveling can be done with either fine or coarse abrasive, but is usually more effective with the coarse. Beveling, incidentally, will also raise the frequency because of the quartz ground off during the process.

Although beveling will usually improve the activity, another method — and probably the simplest — is to change electrodes. The land heights on the electrodes have a critical effect on activity. If the center of the crystal becomes too high and the lands are so low that the center of the crystal touches the center of the electrodes, the crystal will stop oscillating.

*Fig. 6-9* — The ½ × ½-inch electrodes used in modern crystal holders, showing the lands at the corners between which the crystal is firmly held.




The last step — and the most drastic method of raising activity — is to edge-grind adjacent edges. This grinding is best done with coarse abrasive and should be followed by a slight bevel to remove any chips which may remain. By checking the crystal frequently, a drop in activity can be corrected by the above methods. If the crystal is ground too far and goes completely dead, the frequency may be too high when the crystal is again reactivated.

Most crystals produced in the last five years or so have been brought to the desired frequency by an etching process. This process is not only a convenient means of quantity production, but it also results in a completely-clean surface for the crystal, which plays an important part in the activity of the crystal and the maximum power it will handle without overheating. Therefore regrinding may impair the performance of etched crystals, since regrinding destroys the etched surface.

# R.F. Power Amplifiers

## SCREEN-GRID AMPLIFIERS

Since the power output from an oscillator is limited for reasons previously stated, r.f. power greater than 10 or 15 watts, if the oscillator is crystal-controlled, or a watt or two if VFO is used, usually is obtained by feeding the output of the oscillator into one or more amplifiers as may be required to raise the power level to that desired before feeding it to the antenna.

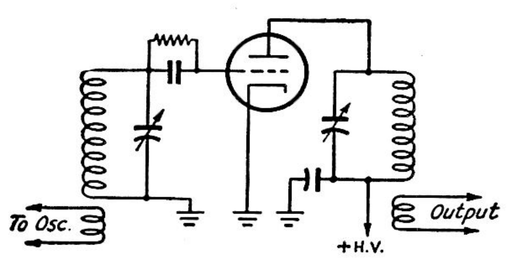



*Fig. 6-10* — Fundamental r.f. power-amplifier circuit. Means must be provided to prevent oscillation since the circuit is the same as that for a t.g.t.p. oscillator. See text for discussion.

Fig. 6-10 shows a fundamental amplifier circuit. The oscillator output is fed into the grid circuit of the amplifier. Power output is taken from the plate circuit. Both grid and plate circuits are tuned to the frequency of the oscillator. It will be noticed, however, that this fundamental circuit is the same as the circuit shown for the tuned-grid tuned-plate oscillator of Fig. 6-2B. Therefore the amplifier circuit itself will function as an oscillator, independently of the oscillator feeding it, unless measures are taken to make this condition impossible.

Feed-back is through the path provided by the grid-plate capacitance of the tube. Therefore an obvious measure is to decrease this capacitance so that the feed-back is not sufficient to support oscillation. This can be done by using a well-screened tube in the amplifier, as shown in Fig. 6-11A, and arranging the amplifier-circuit components, particularly the grid and plate tank coils, so that no other coupling for feed-back from output to input circuits exists. Tubes such as the 6K7, 6AG7, 2E25, 807, 814 and 4-125A and similar types can be used successfully in this way. The less-completely screened tubes, such as the 6F6, 6V6, 6L6 and 813 usually require other means of stabilization.

The equivalent push-pull circuit is shown in Fig. 6-11B. The r.f. chokes are used so that the centers of the coils will not be grounded for r.f. through the power supply, thus setting up two distinct tank circuits, each consisting of half of the coil and one section of the split-stator condenser. The circuits for pentodes are the same with the addition of the grounded suppressor grid.

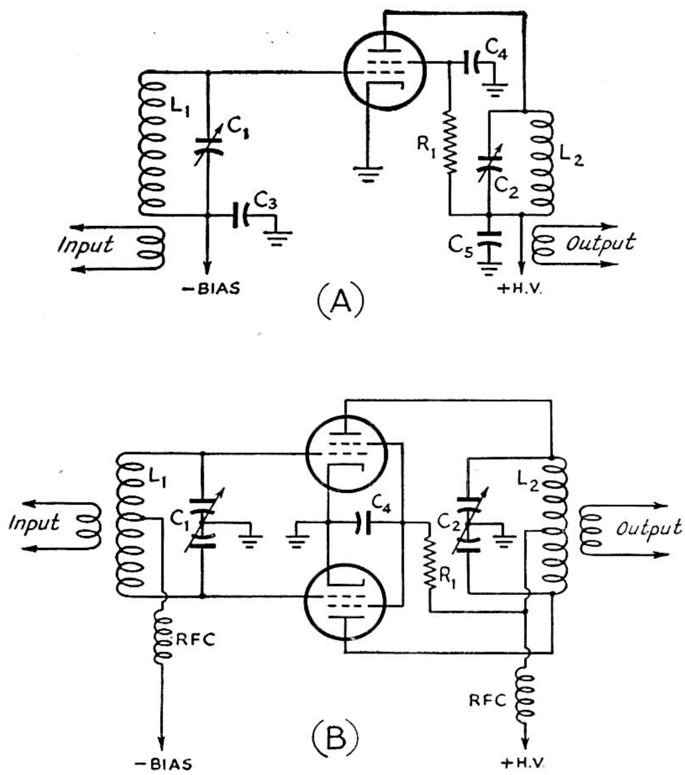



*Fig. 6-11* — Screen-grid amplifier circuits. A — Single-tube amplifier. B — Push-pull.
$L_1C_1$ (grid tank) and $L_2C_2$ (plate tank) are tuned to the frequency being fed to the amplifier.
$C_3$ — Grid by-pass condenser — 0.01-$\mu$fd. paper.
$C_4$ — Screen by-pass condenser — 0.01-$\mu$fd. paper.
$C_5$ — Plate by-pass condenser — 0.01-$\mu$fd. paper.
$R_1$ — Screen voltage-dropping resistor.
RFC — Isolating r.f. chokes — 1 to 2.5 mh.

## NEUTRALIZED TRIODE AMPLIFIERS

Another method of preventing feed-back is to neutralize the effect of the grid-plate capacitance by feeding, through a separate path to the grid, a voltage which is equal, but in opposite phase, to the voltage fed back from plate to grid through the plate-grid capacitance of the tube.

The most commonly-used circuits for this purpose are shown in Fig. 6-12. Amplifiers using these systems of neutralization are known as **plate-neutralized amplifiers.** In each case, the midpoint of the plate tank circuit, either coil or condenser, is grounded, so that the voltages at opposite ends of the tank are essentially equal, but 180 degrees out of phase.

The value of voltage fed back to neutralize the amplifier is adjusted to match that fed back through the grid-plate capacitance of the tube by adjusting the capacitance of the neutralizing condenser, $C_6$.

### *Capacitive Voltage Division*

In Fig. 6-12A, the division of voltage across the tank circuit is dependent upon the ratio of the capacitances of the two sections of the tank condenser. Since these capacitances are equal in a split-stator condenser, the voltages at the ends of the tank circuit in respect to the cathode, which is connected to the center of the tank circuit through ground, are equal. Therefore the neutralizing voltage is the same as the feed-back voltage when the capacitance of the neutralizing condenser is equal to the grid-plate capacitance of the tube.

In this circuit, the output (plate-cathode) capacitance of the tube shunts only one section of the tank condenser. Therefore that half of the tank circuit has a greater total capacitance across it than the other half. Since the added capacitance is fixed in value, this means that the ratio of capacitance across the two sections does not remain constant for all settings of the tank condenser, especially if the tube output capacitance is large. When the tank condenser is at maximum capacitance, the tube capacitance usually is a small percentage of the condenser capacitance and therefore the unbalance between the two sections is at a minimum. However, when the tank condenser is set near minimum capacitance, the tube capacitance may be as large or larger than the condenser capacitance and the unbalance is appreciable. Therefore a fixed adjustment of the neutralizing condenser may not hold neutralization for all settings of the tank condenser. This condition can be corrected by connecting a condenser, $C_7$, of capacitance equal to the tube output capacitance across the neutralizing-condenser side of the tank circuit, as shown in Fig. 6-13, to balance the tube capacitance which appears across the other side. The alternative is to adjust the inductance of the coil so that the circuit tunes to the desired frequency near the maximum-capacitance setting of the tank condenser, where the unbalance is less than near minimum capacitance.

### *Inductive Voltage Division*

In Fig. 6-12B, the voltage division for neutralization is dependent upon the ratio of inductances in the two sections of the coil. The coil usually is tapped at the center to give equal voltages at the ends of the tank circuit. However, the tap may be placed off center, thus making the voltage at the plate end of the circuit higher or lower than the voltage at the other end of the circuit in respect to cathode. In this case, the capacitance of the neutralizing condenser must be altered to compensate. If the plate end of the coil has more inductance than the other end, the capacitance of the neutralizing condenser must be greater than the grid-plate capacitance of the tube. If the plate end has less inductance than the other end, the neutralizing capacitance must be less than the grid-plate capacitance of the tube.

One disadvantage of the circuit of Fig. 6-12B is that when plug-in coils are used to change the amplifier tuning from one fre-

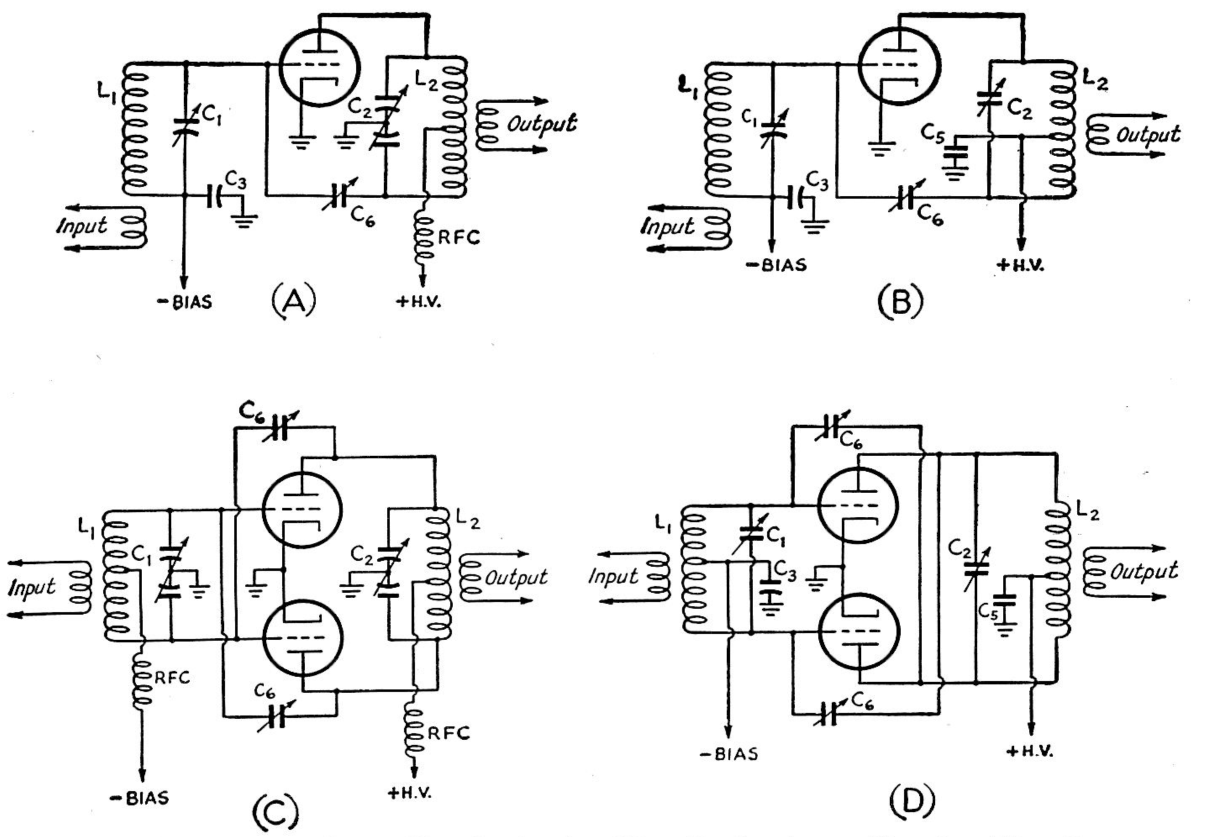



*Fig. 6-12* — Neutralized triode-amplifier circuits. A and B — Single-tube amplifier. C and D — Push-pull. $C_6$ should have approximately the same value as the plate-grid capacitance of the tube when the plate tank circuit is grounded at the center (see text for other conditions). All other values same as corresponding values given under Fig. 6-11.

quency band to another, it may be difficult to tap the coils for each band with sufficient similarity to avoid the necessity for readjustment of the neutralizing condenser each time coils are changed.

### *Push-Pull Neutralized Amplifiers*

Since a push-pull circuit is a balanced arrangement, it is more easily neutralized and a single adjustment of the neutralizing condenser will hold over a greater tuning range. Circuits with inductive and capacitive voltage division are shown in Fig. 6-12C and D. While the circuit of C usually is considered preferable, inductive voltage division, as shown at D, sometimes is employed to permit the use of single-section tank condensers. In theory the single-section condensers introduce an unbalance in respect to ground, but in practice, the unbalance may not be so serious that the circuit cannot be used successfully, at least at 3.5 and 7 Mc.

## ● OTHER NEUTRALIZING CIRCUITS

Additional, but less widely-used neutralizing circuits are shown in Fig. 6-14. The circuit of Fig. 6-14A is similar to that of Fig. 6-12A, except that the voltage division takes place in the grid circuit instead of the plate circuit. Any voltage which may be fed back to the grid circuit through the grid-plate capacitance of the tube is divided in the grid tank circuit so that half appears at the grid, while the other half is fed, 180 degrees out of phase, back to the plate. In another similar version the grid tank coil, instead of the condenser, is used as the voltage divider, the circuit being similar to Fig. 6-12B.

### *Link Neutralization*

The link neutralizing circuit of Fig. 6-14B sometimes is useful as an expedient to stabilize a screen-grid amplifier which is not sufficiently screened or shielded. It has the advantage that it may be added readily to an already-existing amplifier circuit without the necessity for a major alteration in either grid or tank circuits which would be required to shift the ground point to the center of the tank circuit. The link provides the path for coupling back the neutralizing voltage and proper phasing is dependent upon the polarization of the two link coils. Connections to one of the link coils may be switched to obtain correct polarization.

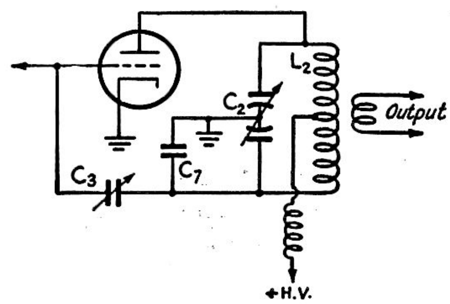



*Fig. 6-13* — In this neutralizing circuit, $C_7$, which has the same capacitance as the output capacitance of the tube, has been added to compensate for the tube capacitance across the upper half of the circuit.